

# Engineering Control Systems

A Practical Introduction for Mechanical,  
Civil, and Aerospace Engineers



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## Preface

This open-source text is designed to offer the reader a complete text on the basics of control theory within the context of Mechanical Engineering.

## Cover Art

Union Pacific No. 119 is a replica of the 4-4-0 “American” type steam locomotive that participated in the Golden Spike ceremony at Promontory Summit, Utah, on May 10, 1869, marking the completion of the first transcontinental railroad in the United States. The original locomotive was built in 1868 by Rogers Locomotive and Machine Works for the Union Pacific Railroad and was scrapped in 1903. The replica on the cover was built by O’Connor Engineering Laboratories and delivered to Promontory in 1979. The transcontinental railroad represents a significant milestone in industrial development and the practical applications of control theory.

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## Source Code

The source code for this text is available at <https://github.com/austindowney/Engineering-Control-Systems>.

# Contents

<b>1 Text Overview</b>	<b>7</b>
1.1 Basics of Control System . . . . .	8
1.2 Videos of control systems . . . . .	9
1.3 History . . . . .	9
1.4 Examples of control systems . . . . .	10
<b>2 Systems</b>	<b>12</b>
2.1 Basic Concepts in Vibrations . . . . .	12
2.2 Single Degree-of-Freedom Systems . . . . .	12
2.2.1 Spring-Mass Model . . . . .	13
2.2.2 Linear Springs . . . . .	14
2.2.3 Equation of Motion for an Oscillating System . . . . .	15
2.3 Forcing Function . . . . .	18
2.3.1 Step Function . . . . .	18
2.3.2 Pulse Function . . . . .	18
2.3.3 Impulse Function . . . . .	19
2.3.4 Ramp Function . . . . .	19
2.4 Introduction to Stability . . . . .	20
2.5 1 <sup>st</sup> Order Systems . . . . .	21
2.5.1 Spring-damper mechanism . . . . .	21
2.5.2 Homogeneous Equation (Free Response) . . . . .	22
2.5.3 Stability of 1 <sup>st</sup> -order systems . . . . .	23
2.5.4 Stability of 1 <sup>st</sup> -order Systems Under Forced Response . . . . .	24
2.6 2 <sup>nd</sup> Order Systems . . . . .	24
2.6.1 Homogeneous Equation (Free Vibration Response) . . . . .	25
2.7 Forced Vibration Response . . . . .	28
2.8 Spring-mass-damper oscillator . . . . .	29
2.9 Stability of 2nd order systems . . . . .	32
<b>3 Transfer Functions</b>	<b>37</b>
3.1 Laplace Transforms . . . . .	37
3.2 Transfer Function Method for Solving Vibrating Systems . . . . .	40
3.2.1 Free Vibration for Undamped Systems . . . . .	40
3.2.2 Forced Vibration (Impulse) for damped Systems . . . . .	41
3.2.3 Arbitrary Inputs to a System . . . . .	43
3.3 Laplace method in Controls . . . . .	45
3.3.1 Laplace transform of exponential function . . . . .	45
3.3.2 Laplace transform of step function . . . . .	47
3.3.3 Laplace transform of ramp function . . . . .	48
3.3.4 Laplace transform shifted step function . . . . .	49
3.3.5 Laplace transform pulse function . . . . .	50
3.3.6 Laplace transform impulse function . . . . .	51
3.4 Properties of Laplace Transforms . . . . .	53

3.4.1	Differentiation Property . . . . .	54
3.4.2	Integration Property . . . . .	55
3.4.3	Time Shift Property . . . . .	55
3.4.4	Final Value Theorem . . . . .	57
3.5	Convolutional property of Laplace Transforms . . . . .	58
3.6	Transfer Function for Response to Random Inputs . . . . .	58
3.6.1	Defining the transfer function $H(s)$ . . . . .	58
3.6.2	Transfer Function method (Steady-State solution) . . . . .	60
3.6.3	Response to Random Inputs . . . . .	63
3.7	Inverse Laplace Transform . . . . .	69
3.8	Dominant Poles . . . . .	72
3.8.1	Case A (Unstable): . . . . .	72
3.8.2	Case B (Stable): . . . . .	72
<b>4</b>	<b>Time Series Response</b> . . . . .	<b>75</b>
4.1	Transfer Functions Block Diagrams . . . . .	75
4.1.1	Create transfer function using numerator and denominator coefficients . . . . .	75
4.1.2	Create transfer function using Zeros, Poles, and Gain . . . . .	76
4.2	Order verse Type . . . . .	77
4.3	Time Response . . . . .	78
4.3.1	Time Series Response for a Step Function . . . . .	78
4.3.2	Time Series Response for a Impulse Function . . . . .	79
4.3.3	Time Series Response for a Ramp Function . . . . .	80
4.4	1 <sup>st</sup> Order System Time Response . . . . .	80
4.4.1	Step response of a 1 <sup>st</sup> Order System . . . . .	81
4.4.2	Impulse response of a 1 <sup>st</sup> Order System . . . . .	82
4.4.3	Ramp response of a 1 <sup>st</sup> Order System . . . . .	82
4.4.4	Summary of the First Order System Responses . . . . .	83
4.5	2 <sup>nd</sup> Order System Time Response . . . . .	85
4.5.1	Step Response for a 2 <sup>nd</sup> Order System . . . . .	86
4.5.2	Impulse response of a 2 <sup>nd</sup> Order System . . . . .	89
4.5.3	Ramp response of a 2 <sup>nd</sup> Order System . . . . .	91
4.5.4	Summary of the Second Order System Responses . . . . .	94
4.6	Stability of response . . . . .	95
4.6.1	Stability of 1 <sup>st</sup> -Order Responses . . . . .	96
4.6.2	Stability of 2 <sup>nd</sup> -Order Responses . . . . .	97
4.6.3	Stability of Higher-order Responses . . . . .	98
4.6.4	Absolute Stability . . . . .	100
4.6.5	Marginal Stability . . . . .	100
4.6.6	Relative Stability . . . . .	100
<b>5</b>	<b>Performance Indicators</b> . . . . .	<b>101</b>
5.1	1 <sup>st</sup> -order System Generic Performance Indicators . . . . .	101
5.1.1	Step response 1 <sup>st</sup> -order system performance indicators . . . . .	102
5.1.2	Impulse response 1 <sup>st</sup> -order system performance indicators . . . . .	103

5.1.3	Ramp response 1 <sup>st</sup> -order system performance indicators . . . . .	104
5.2	1 <sup>st</sup> -order System Specific Performance Indicators . . . . .	105
5.2.1	Impulse Response Specific Performance Indicators . . . . .	107
5.3	2 <sup>nd</sup> -order System Generic Performance Indicators . . . . .	108
5.3.1	Step response 2 <sup>nd</sup> -order system performance indicators . . . . .	109
5.3.2	Impulse response 2 <sup>nd</sup> -order system performance indicators . . . . .	111
5.3.3	Ramp response 2 <sup>nd</sup> -order system performance indicators . . . . .	112
5.4	2 <sup>nd</sup> -order System Specific Performance Indicators . . . . .	114
5.4.1	Procedure for a Step Response . . . . .	115
<b>6</b>	<b>System Identification</b>	<b>120</b>
6.1	Experimental Signal Processing . . . . .	120
6.2	System Identification for 1 <sup>st</sup> -order systems . . . . .	122
6.2.1	Option 1 Optimization . . . . .	122
6.2.2	Option 2 Graphical Methods . . . . .	122
6.2.3	Option 3 Performance Indicators . . . . .	125
6.3	System Identification for 2 <sup>nd</sup> -order systems . . . . .	125
6.3.1	Optimization . . . . .	126
6.3.2	Frequency Estimation . . . . .	126
6.3.3	Step Response Analysis ( $\zeta \ll 1$ ) . . . . .	127
6.3.4	Logarithmic decrement . . . . .	132
<b>7</b>	<b>Control Systems</b>	<b>137</b>
7.1	On-Off (Bang-Bang) Control . . . . .	137
7.2	Proportional-Integral-Derivative (PID) Control . . . . .	137
7.3	Control Systems . . . . .	141
7.4	Stability of feedback control systems and root locus . . . . .	167
7.5	Stability Criteria . . . . .	183
7.6	Feedback Controllers . . . . .	195
7.7	SIMULINK Aircraft Roll Motion . . . . .	222
7.8	Feedback Controllers (continued) . . . . .	252
7.9	SIMULINK Aircraft Roll Motion (continued) . . . . .	261
7.10	Instability Suppression with velocity feedback . . . . .	266
<b>8</b>	<b>Frequency Analysis</b>	<b>271</b>
8.1	Time Response Under Harmonic Excitations . . . . .	271
8.2	Frequency Response Function (FRF) . . . . .	277
8.3	Performance Indicators in Frequency Domains . . . . .	302
8.4	System Identification in Frequency Domains . . . . .	316
8.5	Frequency Domain Analysis of Feedback System Stability . . . . .	322
8.6	Stability Margins . . . . .	370
8.7	Phase Compensators . . . . .	386
8.8	Lag Compensators . . . . .	428
<b>9</b>	<b>Combined Analysis</b>	<b>433</b>

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<b>10 Control Systems Designer</b>	<b>446</b>
<b>12 State-Space Representations</b>	<b>483</b>
<b>12 Machine Learning Control</b>	<b>505</b>

# 1 Text Overview

- Systems, 1<sup>st</sup> order, 2<sup>nd</sup> order, and higher order.
  - Introduction to control systems
  - Differential equations
  - General solution
  - Free Response
  - Stability
  - Laplace Transforms
- Performance
  - Time response
  - Performance indicators
  - System identification
- Control Systems
  - Feedback
  - Stability
  - Controllers
- Frequency Analysis
  - FRF
  - Bode
  - Nyquist
  - Manguis
- Additional Topics
  - Combined Analysis
  - CSD
  - Single-input single-output (SISO)
  - Tool
  - State Space Models (introduction)

## 1.1 Basics of Control System

A control system consisting of interconnected components is designed to achieve a desired purpose. To understand the purpose of a control system, it is useful to examine examples of control systems through the course of history. These early systems incorporated many of the same ideas of feedback that are in use today.

Modern control engineering practice includes the use of control design strategies for improving manufacturing processes, the efficiency of energy use, advanced automobile control, including rapid transit, among others.

**System** - An interconnection of elements and devices for a desired purpose.

**Control System** - An interconnection of components forming a system configuration that will provide a desired response.

**Process** - The device, plant, or system under control. The input and output relationship represents the cause-and effect relationship of the process.

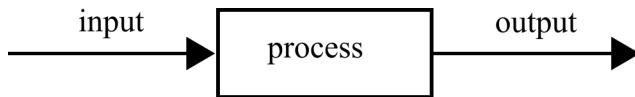


Figure 1.1: Process to be controlled.

**Open-Loop Control Systems** - utilize a controller or control actuator to obtain the desired response.

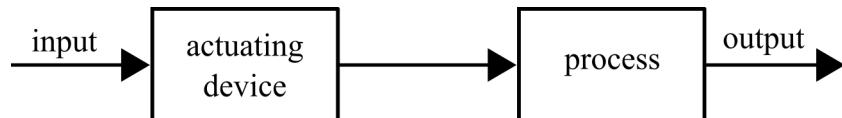


Figure 1.2: Open-loop control system (without feedback).

**Closed-Loop Control Systems** - utilizes feedback to compare the actual output to the desired output response.

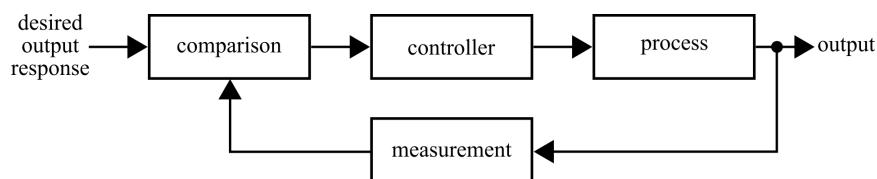


Figure 1.3: Closed-loop control system (with feedback).

**Multivariable Control System** - a control system with multiple desired response controlled by multiple output variables.

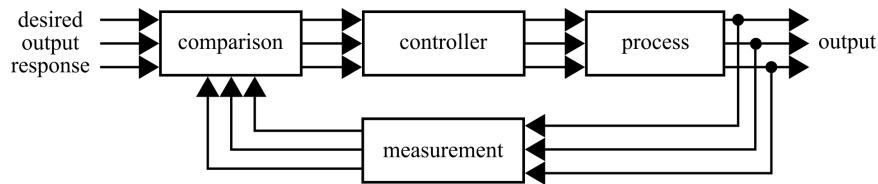


Figure 1.4: Multivariable control system with many inputs and many outputs to the system.

**Positive and negative feedback** - A positive feedback (also called Regenerative feedback) adds control outputs to the system in the same direction as the original perpetuation while a negative feedback (Degenerative feedback) provides a counteracting force to the original permutations.

## 1.2 Videos of control systems

### Ball and Plate PID control with 6 DOF Stewart platform

[https://www.youtube.com/watch?v=j40mVLc\\_oDw&t=95s&ab\\_channel=FullMotionDynamics](https://www.youtube.com/watch?v=j40mVLc_oDw&t=95s&ab_channel=FullMotionDynamics)

### Ball and Plate PID control with 6 DOF Stewart platform

[https://www.youtube.com/watch?v=meMWfva-Jio&ab\\_channel=StepanOzana](https://www.youtube.com/watch?v=meMWfva-Jio&ab_channel=StepanOzana)

### SpaceX first landing

[https://www.youtube.com/watch?v=1sJlFzUQVmY&ab\\_channel=BloombergQuicktake](https://www.youtube.com/watch?v=1sJlFzUQVmY&ab_channel=BloombergQuicktake)

## 1.3 History

- Greece (BC) - Float regulator mechanism.
- Holland (16th Century) - Temperature regulator.
- 18th Century James Watt's centrifugal governor for the speed control of a steam engine.
- 1920s Minorsky worked on automatic controllers for steering ships.
- 1930s Nyquist developed a method for analyzing the stability of controlled systems.
- 1940s Frequency response methods made it possible to design linear closed-loop control systems.
- 1950s Root-locus method due to Evans was fully developed.
- 1960s State space methods, optimal control, adaptive control.
- 1980s Learning controls are begun to investigated and developed.
- Present and on-going research fields. Recent application of modern control theory includes such non-engineering systems such as biological, biomedical, economic and socio-economic systems

## 1.4 Examples of control systems

Controllers in the form of centrifugal governors for steam engines set off the industrial revolution by allowing steam engines to produce reliable and controllable power. Figure 1.5 shows centrifugal governor on steam engines.

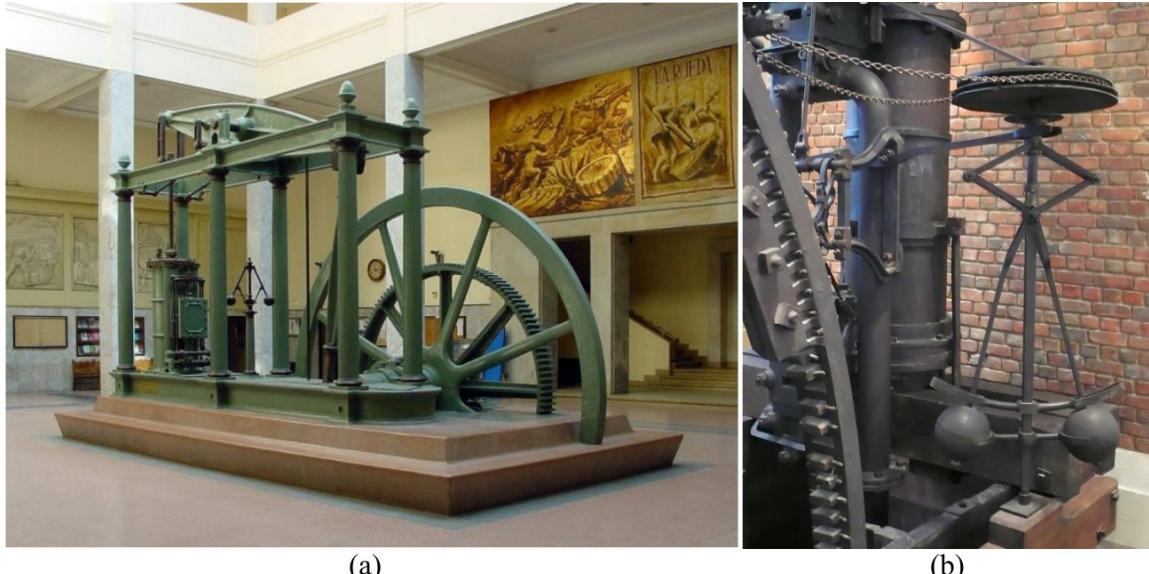
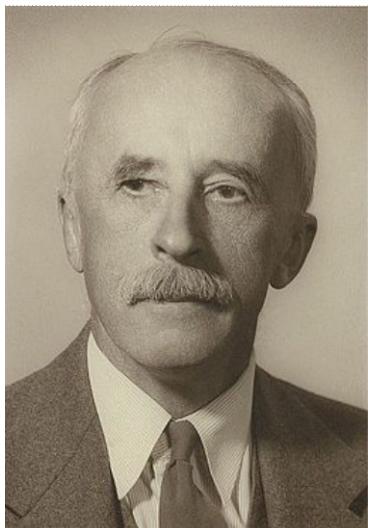


Figure 1.5: Centrifugal governors on a: (a) Boulton & Watt steam engine from 1788<sup>1</sup>, and; (b) D. Napier & Son from 1832 double-acting steam engine<sup>2</sup>.

<sup>1</sup>Dr. Mirko Junge, CC BY 3.0 <<https://creativecommons.org/licenses/by/3.0/>>, via Wikimedia Commons.

<sup>2</sup>Nicolás Pérez, CC BY-SA 3.0 <<http://creativecommons.org/licenses/by-sa/3.0/>>, via Wikimedia Commons.

Arguably, the Russian American engineer Nicolas Minorsky was the first to develop the theoretical analysis for the three-term control we now call PID. This was done in 1922 while he was researching and designing automatic ship steering for the US Navy. He based his work on watching how a ship's helmsman responds to wave loading on a ship, with a delayed input to the helm that not only considered the current ship course, but also past error and the desired rate of change for the ship. For a helmsman, the goal is stability, not absolute control, which simplifies how one thinks about the challenge of control.



(a)



(b)

Figure 1.6: Historical prospective of PID control showing: (a) Portrait of Nicolas Minorsky<sup>1</sup> and (b) the battleship USS New Mexico (BB-40) of the United States Navy which was the first to implement PID control in its steering<sup>2</sup>.

<sup>1</sup>Peter Minorsky, grandson of Nicolas Minorsky, CC BY-SA 1.0 <<https://creativecommons.org/licenses/by-sa/1.0/>>, via Wikimedia Commons

<sup>2</sup>U.S. Navy, Public domain, via Wikimedia Commons

## 2 Systems

### 2.1 Basic Concepts in Vibrations

The study of vibrations, within the broader field of classical mechanics, is the investigation of oscillations that occur about an equilibrium point. Vibrations, both desired and undesired, are present in all mechanical systems and can be helpful (e.g. a soil sieve, rotary sander) or destructive (e.g. an aircraft frame in resonance). The oscillations that form a vibrating system may be periodic (e.g., pendulum) or random (e.g. turbulence in an airplane), or a combination of the two.

Vibrations impact our daily lives in a variety of ways, from the sound made by banjo strings that vibrates between 140 and 400 Hz to the 4-6 Hz vibration felt by a passenger in a car seat. The consideration of the vibrations and their associated mathematical modeling are an important factor in the design of mechanical systems. In the material that follows, the fundamental theories of vibration are presented and modeled using fundamental physical principles such as Newton's three laws of motion. These models are analyzed using the mathematical tools of calculus and differential equations.

**Review 2.1** Newton's three laws of motion:

1. In an inertial frame of reference, an object either remains at rest or continues to move at a constant velocity, unless acted upon by a force.
2. In an inertial reference frame, the vector sum of the forces  $F$  on an object is equal to the mass  $m$  of that object multiplied by the acceleration of the object:  $F = ma$ . (It is assumed here that the mass  $m$  is constant)
3. When one body exerts a force on a second body, the second body simultaneously exerts a force equal in magnitude and opposite in direction on the first body.

### 2.2 Single Degree-of-Freedom Systems

In its simplest form, the phenomenon of vibration is the exchange of energy between potential and kinetic energy. Therefore, a vibrating system must have a component that stores potential energy. This component must also be capable of releasing the energy as kinetic energy. This kinetic energy is stored in the movement of a mass where the measure of this movement is the velocity of the system and the continuous interchange between potential and kinetic energy is the vibration of the system. The simplest vibrating systems can be modeled as a single-degree-of-freedom (1-DOF) system. In a 1-DOF system, one variable can describe the motion of a system. Potential examples of 1-DOF systems include:

1. yo-yo
2. pogo stick
3. door swinging on axis

#### 4. throttle (gas pedal)

Variables often used for describing 1-DOF systems are  $x(t)$ ,  $y(t)$ ,  $z(t)$ , and  $\theta(t)$ . Examples of 1-DOF systems are presented in figure 2.1 where the assumption of small displacements is made. Note: we will often drop the “( $t$ )” for simplicity in this material.

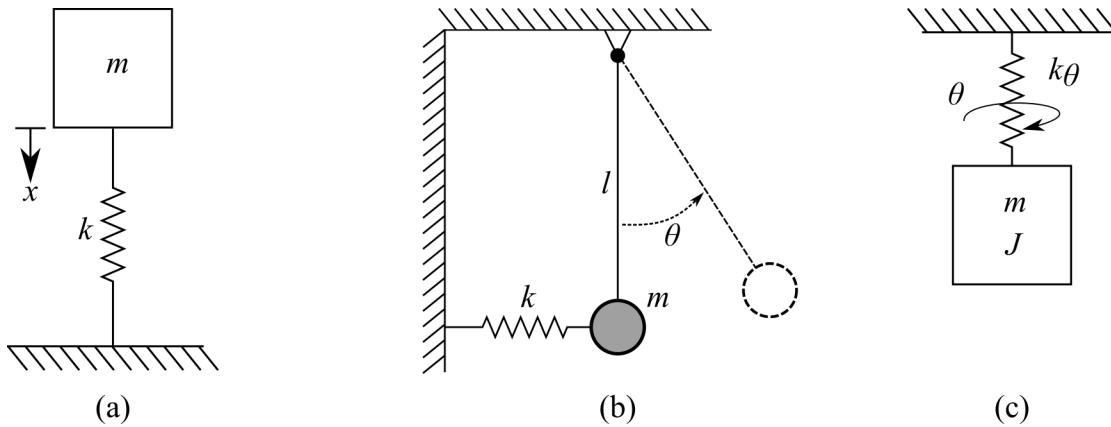


Figure 2.1: Examples of single degree of freedom (DOF) systems showing: (a) a vertical spring-mass system; (b) a simple pendulum; and (c) a rotational spring-mass system.

### 2.2.1 Spring-Mass Model

“All models are wrong, but some are useful”

George E.P. Box (1919 - 2013)

Newtonian physics describes the motion of particles in terms of displacement  $x$ , velocity  $\dot{x}$ , and acceleration  $\ddot{x}$  vectors. Moreover, from Newton’s second law of motion says that the change in the velocity of mass in motion is a product of the force acting on the mass. A simple way to express this phenomenon is through a spring-mass model as presented in figure 2.2. These spring-mass models neglect the mass of the spring and concentrate all the mass of the system into a single point. Note that in this case the force vector and mass-acceleration vectors lie on the same axis and as such are collinear. Therefore, these vectors can be easily treated as scalars simplifying the math used in the modeling of the system.

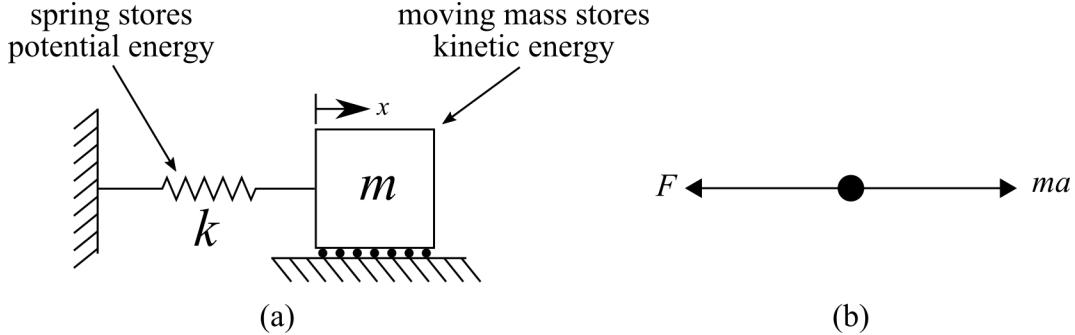


Figure 2.2: A single-degree-of-freedom (1-DOF) spring mass model showing: (a) annotated schematic of a mass-spring system; and (b) the equivalent free-body diagram represented as a point-mass system.

## 2.2.2 Linear Springs

Springs are mechanical devices that store energy, moreover, ideal spring is a theoretical representation of this mechanical device that is massless and responds with a linear increase in force for a unit increase in displacement (i.e.  $F = kx$ ). For simplicity, the spring in the spring-mass model considered here is assumed always ideal linear springs. A graphical representation of the idealized linear spring is presented in figure 2.3 where a unit force  $F$  applied to the free end of the spring results in a unite displacement  $x$  of the spring. The resulting mathematical relationships,  $F = kx$ , is known as Hooke's Law. Nonlinear springs add considerable complexity to the modeling of spring-mass systems, therefore, these are not considered in this introductory work.

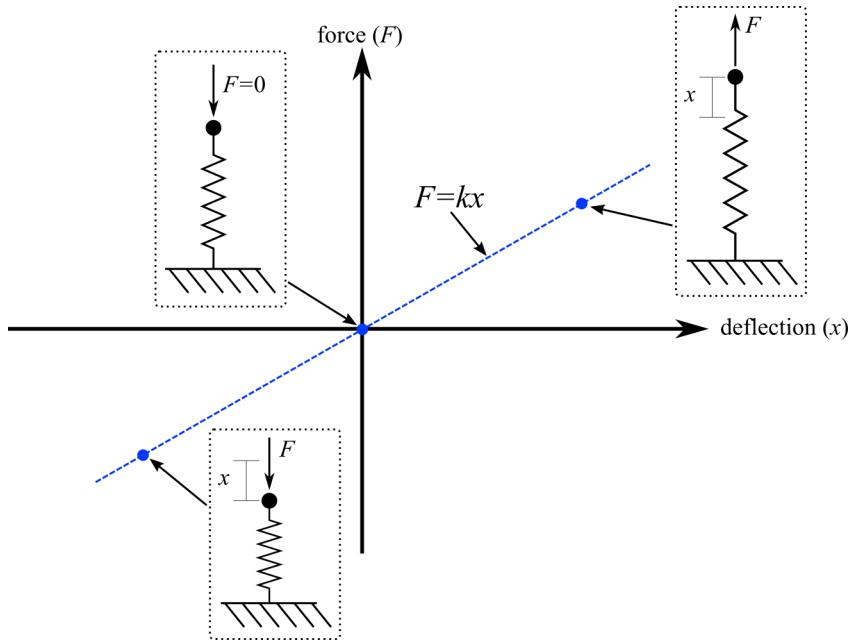


Figure 2.3: Force-displacement plot for a linear spring.

### 2.2.3 Equation of Motion for an Oscillating System

An Equation of Motion (EOM) is an equation that provides a basis for modeling a vibrating system about its equilibrium point and relates the transfer of the potential energy from the spring to the kinetic energy mass. In developing the EOM we assume that any surfaces are frictionless and as such, no energy is extracted from the vibrating system. Referencing the 1-DOF system in figure 2.4(a), and assuming the mass only moves in the  $x$  direction, the only force acting on the mass in the  $x$  direction is the force that results from the elongation of the spring as annotated in figure 2.4(b). Therefore, the sum of forces in along the  $x$  axis must equal the mass ( $m$ ) times the acceleration of the mass ( $\ddot{x}$ ).

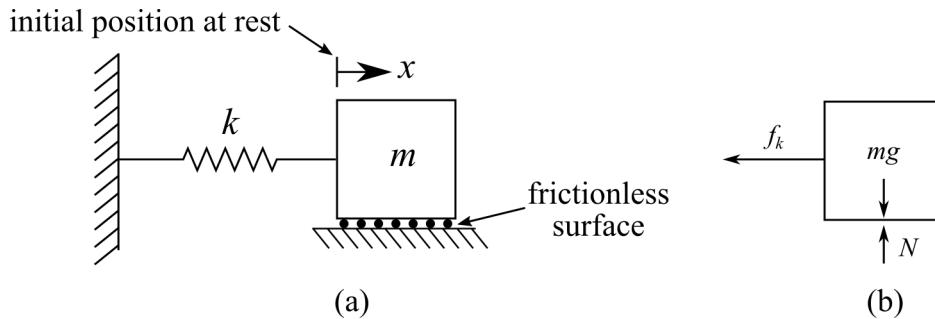


Figure 2.4: A spring mass model of a 1-DOF system showing: (a) a schematic of the system; (b) free-body diagram of the system at its initial position.

Considering that positive displacements are to the right, the standard form of the equation of motion for an undamped system without any excitation is expressed as:

$$s_1\ddot{x} + s_2x = 0 \quad (1)$$

where  $s_1$  and  $s_2$  are constants to be determined for the specific system. A systematic approach to obtaining free-body diagram (FBD) of a system under vibration can be expressed in three steps:

1. Draw a free-body diagram (FBD) at the system's equilibrium and displaced position (without a displacing force).
2. Apply Newton's second law to both FBDs (equilibrium and displaced).
3. Combine the equations to write the EOM in standard form with the forcing component on the right-hand side. For free vibration, the forcing component is 0.

Solving these three steps for 1-DOF system presented in figure 2.4 results in the EOM:

$$m\ddot{x} + kx = 0 \quad (2)$$

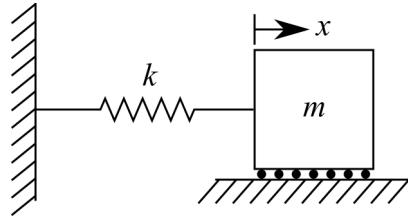
**Review 2.2** A second-order linear homogeneous differential equation has the form:

$$a\ddot{x} + b\dot{x} + cx = 0 \quad (3)$$

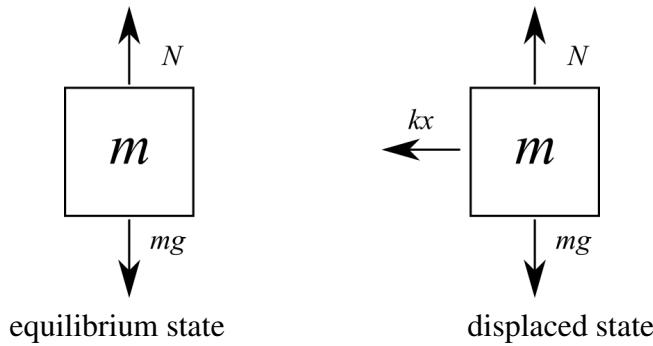
The EOM for a 1-DOF system under a free vibration is a second-order differential equation due to acceleration ( $\ddot{x}$ ) being the second derivative of displacement ( $x$ ) and homogeneous as

the forcing function (right-hand side of the equations) is zero. In EOM's current form,  $m = k$ ,  $b = 0$ , and  $c = k$ . In future work,  $b$  will account for damping in the vibrating system.

**Example 2.1** Considering the system:



**Step-1** Define the direction of displacement, and draw the FBD for the equilibrium and displaced state.



The equation for the equilibrium state is:

$$\stackrel{+}{\rightarrow} \sum F_x = 0 \quad (4)$$

and in the displaced state:

$$\stackrel{+}{\rightarrow} \sum F_x = -kx \quad (5)$$

This equation does not equal zero as the FBD does not account for the restoring force.

**Step-2** Apply Newton's second law (we want to store energy in the kinetic state) of motion to the sum of forces for the displaced position we get:

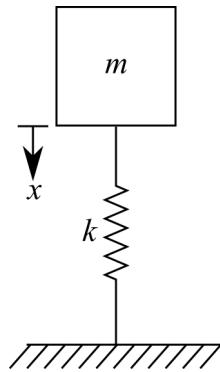
$$ma = m\ddot{x} = \stackrel{+}{\rightarrow} \sum F_x = -kx \quad (6)$$

$$m\ddot{x} = -kx \quad (7)$$

**Step-3** Rearrange in the Equation to construct an EOM:

$$m\ddot{x} + kx = 0 \quad (8)$$

**Example 2.2** Some systems will have an initial displacement, as the system will oscillate around this position we need to define the EOM about this position. Considering the system:



**Step-1** Define the direction of displacement, and draw the FBD for the equilibrium and displaced state.



The equation for the equilibrium state is:

$$+\downarrow \sum F_x = mg - k\delta = 0 \quad (9)$$

and in the displaced state:

$$+\downarrow \sum F_x = mg - k(\delta + x) \quad (10)$$

This equation does not equal zero as the FBD does not account for the restoring force.

**Step-2** Apply Newton's second law (we want to store energy in the kinetic state) of motion to the sum of forces for the displaced position we get:

$$m\ddot{x} = +\downarrow \sum F_x = mg - k\delta - kx \quad (11)$$

We can then use the information from the equilibrium state to cancel out some terms, this becomes:

$$m\ddot{x} = -kx \quad (12)$$

**Step-3** Rearrange in the Equation to construct an EOM:

$$m\ddot{x} + kx = 0 \quad (13)$$

## 2.3 Forcing Function

### 2.3.1 Step Function

A step function is a common loading situation and can represent the dropping of a load into a truck, a car going over a curve, or a motor starting up. The step function ( $\Phi$ ) is also known as the Heaviside function



Figure 2.5: Step function.

### 2.3.2 Pulse Function

Pulse function ( $\rho(t; \tau)$ ) consists of a step up at  $t = 0$  followed by a step down at  $t = \tau$ . The amplitude is  $1/\tau$  so that the area under the pulse is constant and equal to unity ( $A = 1$ ).

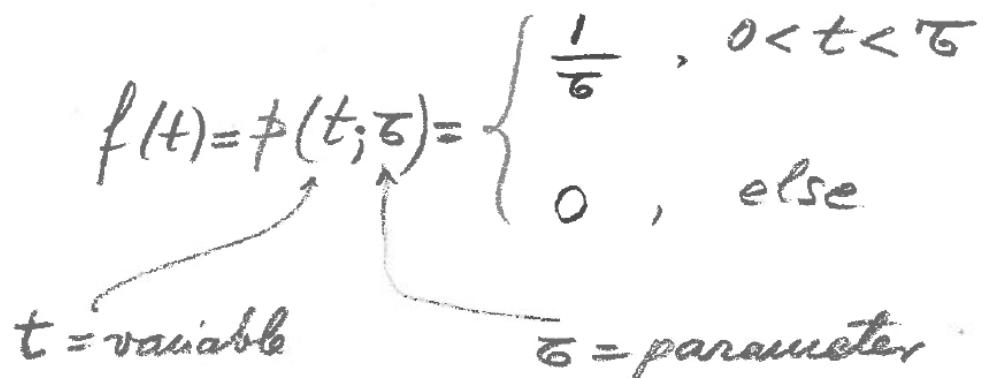
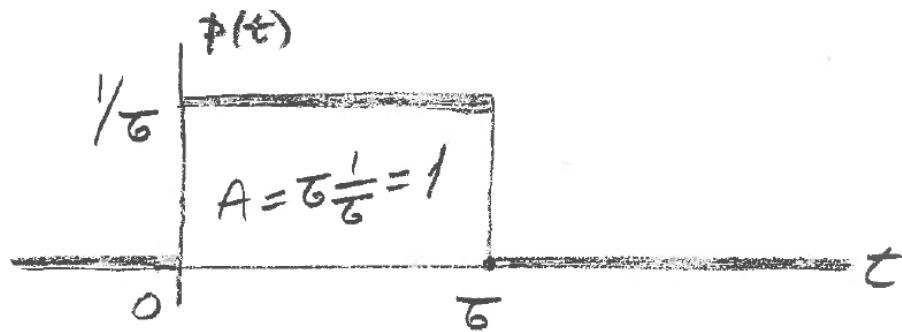


Figure 2.6: Pulse function.

### 2.3.3 Impulse Function

Shock loads on mechanical systems represent a very common source of vibration. These short-duration forces are also called an impulse. An impulse excitation is defined as a force that is applied for a very short, or infinitesimal, length of time. An impulse is a nonperiodic force that is represented by the symbol  $\delta$ . Impulse function ( $\delta(t)$ ) is also known as the “Dirac function”, “Dirac delta function”, “Dirac impulse function”, or “delta function”. The impulse function  $\delta(t)$  is obtained from the pulse function  $\rho(t; \tau)$  by letting  $\tau$  become infinitesimally small, i.e.,

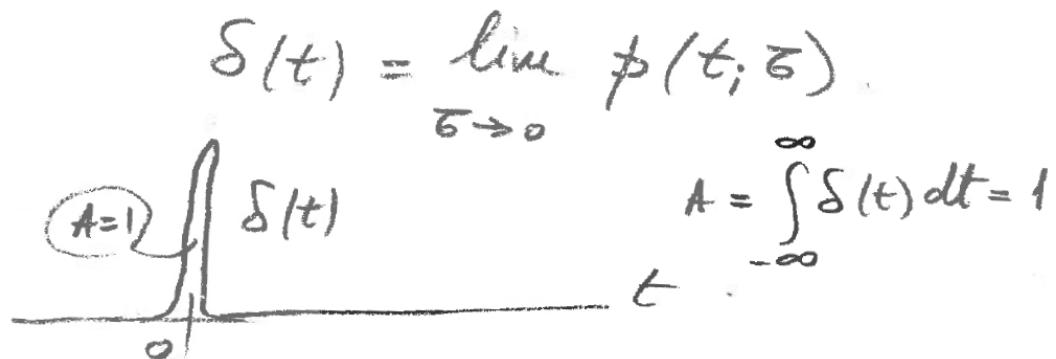
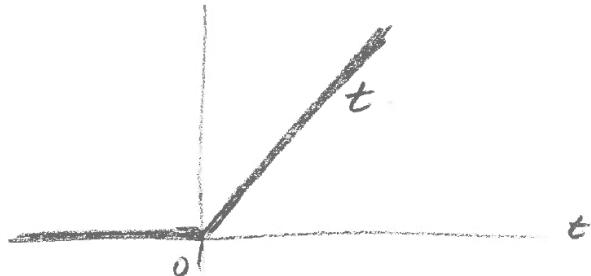


Figure 2.7: Impulse function.

The area under the *delta* function is equal to unity ( $A = 1$ ) just like the pulse function  $\rho(t; \tau)$ .

### 2.3.4 Ramp Function

The ramp function is zero for  $t < 0$  and equal to  $t$  for  $t > 0$ .



$$f(t) = \begin{cases} 0 & , t < 0 \\ t & , t > 0 \end{cases}$$

Figure 2.8: Ramp function.

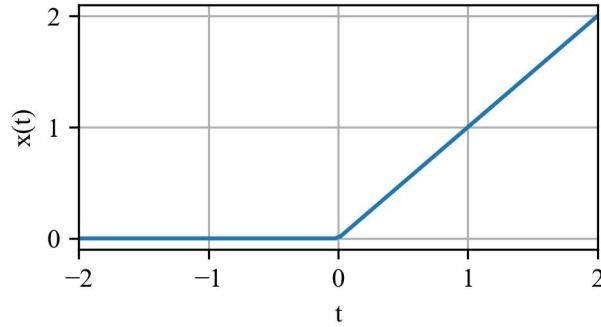


Figure 2.9: Ramp function.

$$f(t) = \begin{cases} t & t > 0 \\ 0 & t \leq 0 \end{cases} \quad (14)$$

## 2.4 Introduction to Stability

The stability of a system is explained through “poles” and “zeros”. The poles and the zeros of a system determine whether the system is stable, and how well the system performs.

1. A system at a pole has an output that is infinite even though the input to the system was finite (i.e. unstable).
2. A system at a zero has an output that is finite even though the input to the system was infinite (i.e. stable).

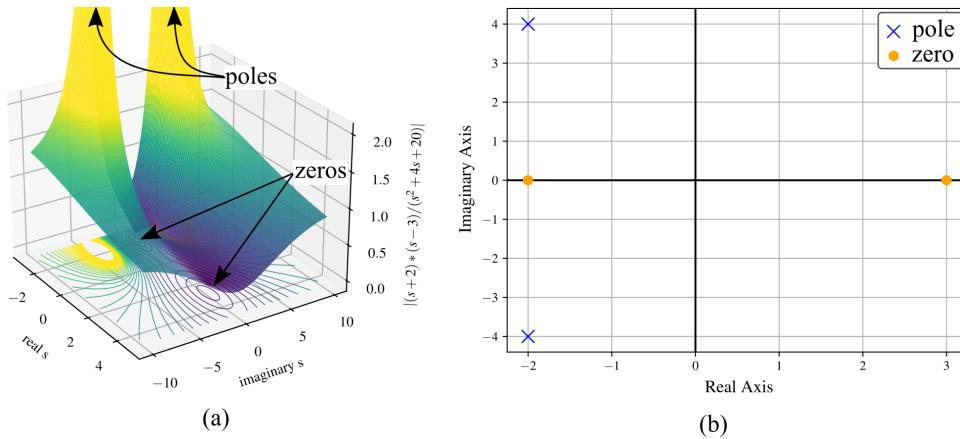


Figure 2.10: Poles and zeros.

Control systems can be designed simply by assigning specific values to the poles and zeros of the system. Physically realizable control systems must have a number of poles greater than the number of zeros. Systems that satisfy this relationship are called Proper. For a rational polynomial

transfer function: The poles of a transfer function are defined as the roots of the denominator polynomial of the transfer function. The zeros of a transfer function are simply the roots of the numerator polynomial. Visualizing of poles and zeros in the complex space can be helpful in understanding the stability of the system. Figure 2.10 shows the complex space for an arbitrary transfer function  $(s+2)(s-3)/s^2 + 4s + 20$  where figure 2.10(a) shows the 3D space while figure 2.10(b) just reports the poles and zeros of the transfer function. Figure 2.11 reports the transfer functions for 3 additional transfer functions.

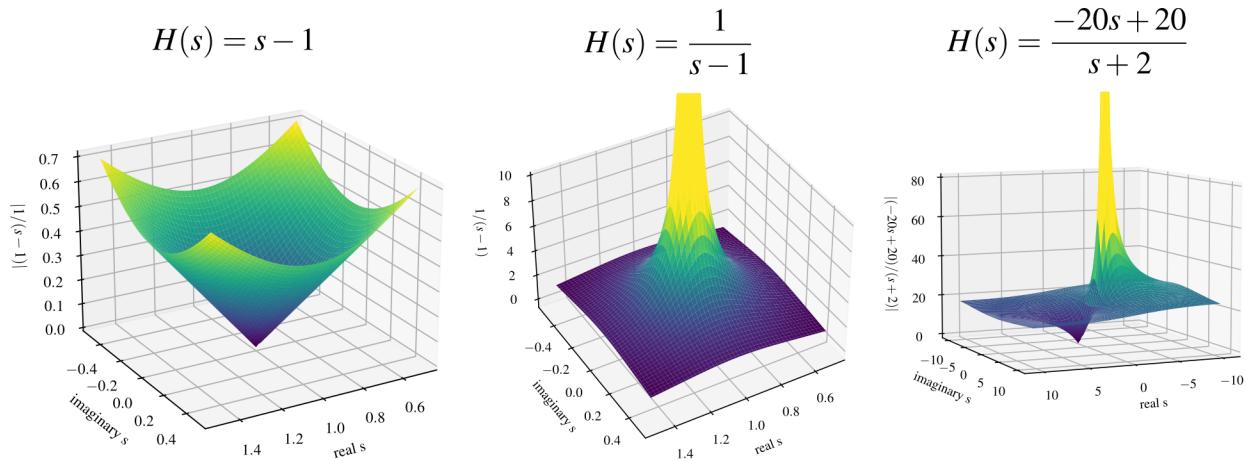


Figure 2.11: Poles and zeros in the complex space, plotted for 3 transfer functions.

## 2.5 1<sup>st</sup> Order Systems

### 2.5.1 Spring-damper mechanism

Consider the system:

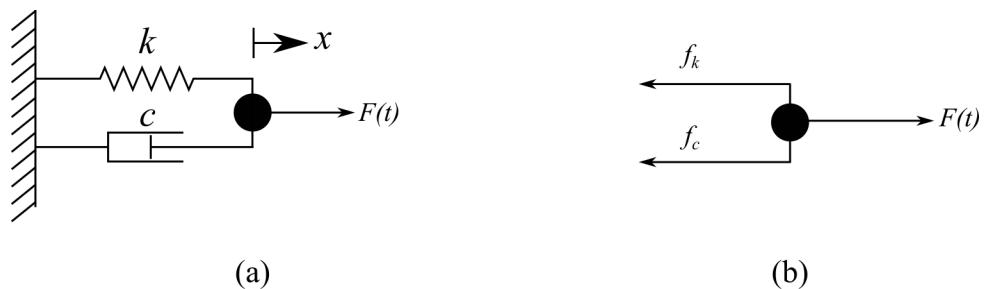


Figure 2.12: Damped 1-DOF system with an external force ( $F(t)$ ) applied, showing: (a) the system configuration; and (b) the free body diagram

Making the assumption that this system does not have any mass, the force balance equation is written as:

$$-kx - cx + f^* = 0 \quad (15)$$

this becomes the equation of motion

$$c\dot{x} + kx = f^* \quad (16)$$

normalizing the equation of motion by the stiffness  $k$ ,

$$\frac{c}{k}\dot{x} + x = \frac{1}{k}f^* \quad (17)$$

define the *time constant*  $T$  as

$$T = \frac{c}{k} \quad (18)$$

and the normalized forcing function  $f(t) = \frac{1}{k}f^*(t)$ . This results in the 1<sup>st</sup>-order ODE in standard form:

$$T\dot{x}(t) + x(t) = f(t) \quad (19)$$

A general solution for this ODE is written as:

$$x(t) = x_c(t) + x_p(t) \quad (20)$$

where:  $x_c$  is called the “complementary solution” and satisfies the homogeneous equation and satisfies the homogeneous equation

$$T\dot{x}_c + x_c = 0 \quad (21)$$

The homogeneous equation is obtained from equation 19 by making zero the right hand side. The complementary solution  $x_c$  is the free response.

$x_p(t)$  is called the “particular solution” and is used to satisfy the right-hand side of equation 19. Substituting equation 20 into equation 19 yields:

$$T(\dot{x}_c + \dot{x}_p) + x_c + x_p = f(t) \quad (22)$$

or

$$T\dot{x}_c + x_c + T\dot{x}_p + x_p = f(t) \quad (23)$$

As  $T\dot{x}_c + x_c = 0$ :

$$T\dot{x}_p + x_p = f(t) \quad (24)$$

This shows that:

- $T\dot{x}_c + x_c$  is the solution to the free response.
- $T\dot{x}_p + x_p$  is the solution to the forced response.

## 2.5.2 Homogeneous Equation (Free Response)

The Homogeneous equation is obtained from equation 19 and setting the right hand side to zero:

$$T\dot{x}_c + x_c = 0 \quad (25)$$

To solve, assume: \_\_\_\_\_

$$\begin{aligned} x_c(t) &= ce^{pt} \\ \dot{x}_c(t) &= pce^{pt} \end{aligned} \quad (26)$$

I really think  
 $c$  should be  $a$   
or it confuses  
with damping.

Inserting equation 26 into equation 25 gives:

$$T(pce^{pt}) + ce^{pt} = 0 \quad (27)$$

canceling out like terms results in the “characteristic equation”:

$$Tp + 1 = 0 \quad (28)$$

therefore:

$$p = \frac{-1}{T} \quad (29)$$

where this is a “pole” of the system. Inserting equation 29 into equation 26 results in the “free response” of the system:

$$x_c(t) = ce^{-t/T} \quad (30)$$

as shown in figure 2.13. Note that the lack of mass in the system prevents it from oscillate as there is no way for it to hold kinetic energy, and therefore no way for it to exchange energy between potential and kinetic forms.

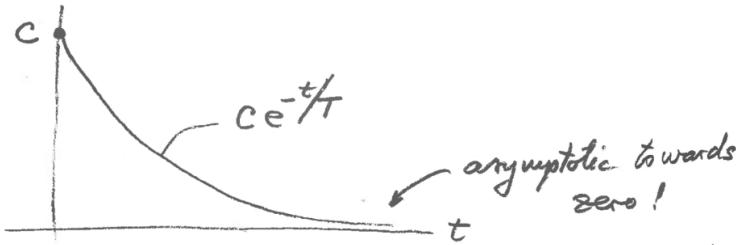


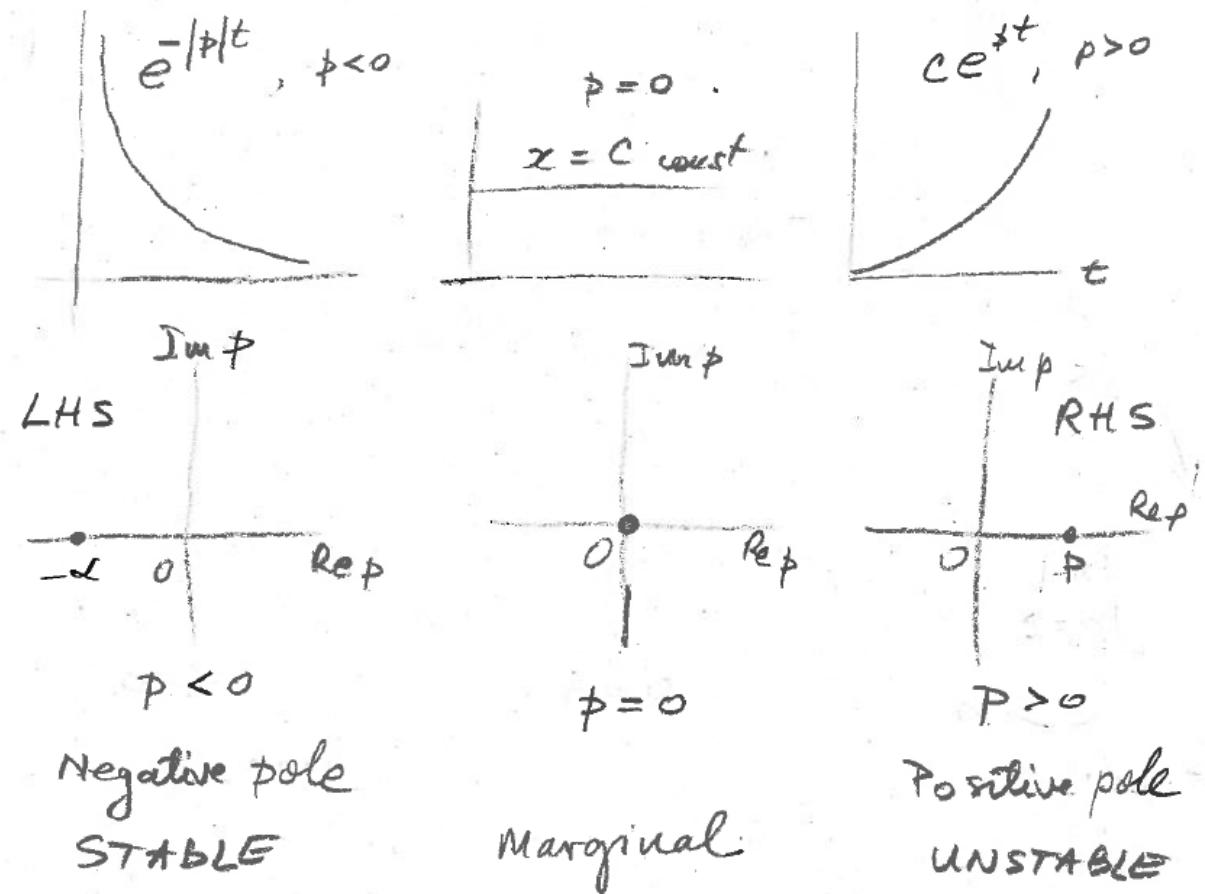
Figure 2.13: Free decay of a 1-DOF system

### 2.5.3 Stability of 1<sup>st</sup>-order systems

Using the general free response form eq 30 where  $p$  is a root of the characteristic equation (equation 28) and is called the “pole”. The stability of the system is dictated by the sign of  $p$ , i.e., its location in the complex  $p$  plane.

**NOTE**

A **Stable System** is a system where if a disturbance is applied, then the system returns to its initial state.

Figure 2.14: Stability of a first order system for different  $p$  values.

#### 2.5.4 Stability of 1<sup>st</sup>-order Systems Under Forced Response

The total solution of equation 19 is given as:

$$x(t) = x_c(t) + x_p(t) \quad (31)$$

or as:

$$x(t) = ce^{-t/\tau} + x_p(t) \quad (32)$$

The constant  $c$  and the function  $x_p(t)$  have to be determined depending on initial conditions and the forms of  $f(t)$ . In control theory, the initial conditions are usually assumed to be zero.

This is incomplete and needs more context.

#### 2.6 2<sup>nd</sup> Order Systems

Consider the system:

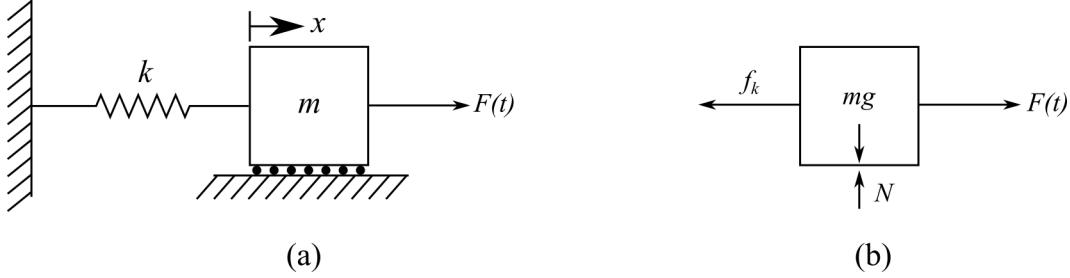


Figure 2.15: 1-DOF system with an external force ( $F(t)$ ) applied, showing: (a) the system configuration; and (b) the free body diagram

Newton's second law of motion states:

$$m\ddot{x} = -kx + f^* \quad (33)$$

this can be rearranged to the standard equation of motion for a system with forced vibrations:

$$m\ddot{x}(t) + kx(t) = f^* \quad (34)$$

It is an 2<sup>nd</sup>-order inhomogeneous ordinary differential equation in time  $t$ . Similar to the solutions to the EOM for a 1<sup>st</sup>-order system, the solution for a 2<sup>nd</sup>-order system are made of two parts, the “complementary solution” and the “particular solution”. Again, this is written as:

$$x(t) = x_c(t) + x_p(t) \quad (35)$$

where  $x_c(t)$  satisfies the homogeneous equation ( $m\ddot{x}(t) + kx(t) = 0$ ) and represent the “free” response of the system.  $x_p(t)$  satisfies the complete equation  $m\ddot{x}(t) + kx(t) = f^*$  and represents the “forced” response of the system.

### 2.6.1 Homogeneous Equation (Free Vibration Response)

The homogeneous equation  $x_c(t)$  is obtained by setting the forcing function of the EOM to zero:

$$m\ddot{x}(t) + kx(t) = 0 \quad (36)$$

normalizing by the mass returns:

$$\ddot{x}(t) + \frac{k}{m}x(t) = 0 \quad (37)$$

denoting the natural frequency of the system:

$$\omega_n^2 = \frac{k}{m} \quad (38)$$

or more simply (and traditionally)

$$\omega_n = \sqrt{\frac{k}{m}} \quad (39)$$

Inserting the natural frequency of the system back into the EOM yields the homogeneous equation in standard form:

$$\ddot{x} + \omega_n^2 x = 0 \quad (40)$$

To solve the homogeneous equation, assume the form:

$$x(t) = ce^{pt} \quad (41)$$

hence

$$\begin{aligned}\dot{x}(t) &= pce^{pt} = px \\ \ddot{x}(t) &= p^2ce^{pt} = p^2x\end{aligned} \quad (42)$$

inserting these back into the EOM for the 2<sup>nd</sup>-order system results in:

$$p^2x + \omega_n^2x = 0 \quad (43)$$

this simplifies to the characteristic equation:

$$p^2 + \omega_n^2 = 0 \quad (44)$$

Rearranging this equations lead to

$$p^2 = -\omega_n^2 \quad (45)$$

and considering that  $i = \sqrt{-1}$  to the solution

$$p_{1,2} = \pm i\omega_n \quad (46)$$

inserting these back into the EOM for the 2<sup>nd</sup>-order system results in:

$$x_c(t) = c_1 e^{i\omega_n t} + c_2 e^{-i\omega_n t} \quad (47)$$

which is the the complementary solution where  $c_1$  and  $c_2$  are constants to be determine. The complementary solution can also be written as:

$$x_c(t) = c \sin(\omega_n t + \phi) \quad (48)$$

where  $c$  is the amplitude of oscillation and  $\phi$  is the phase angle.

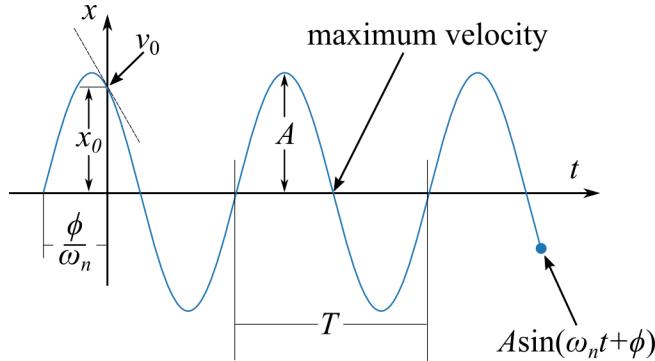


Figure 2.16: Summary of the temporal response for a 1-DOF system.

where  $x_0$  and  $v_0$  are the initial displacement and velocity at  $t=0$  (i.e. the initial conditions).

**Review 2.3** Euler's (pronounced oy-ler) formula, named after Swiss engineer and mathematician Leonhard Euler (1707-1783), is a mathematical formula in complex analysis that establishes the fundamental relationship between the trigonometric functions and the complex exponential function. Euler's formula states that for any real number  $x$ ,

$$e^{j\psi} = \cos(\psi) + j\sin(\psi) \quad (49)$$

where  $j = \sqrt{-1}$ . This equation can also be expressed as:

$$e^{-j\psi} = \cos(\psi) - j\sin(\psi) \quad (50)$$

This can be expressed in terms of polar coordinates as:

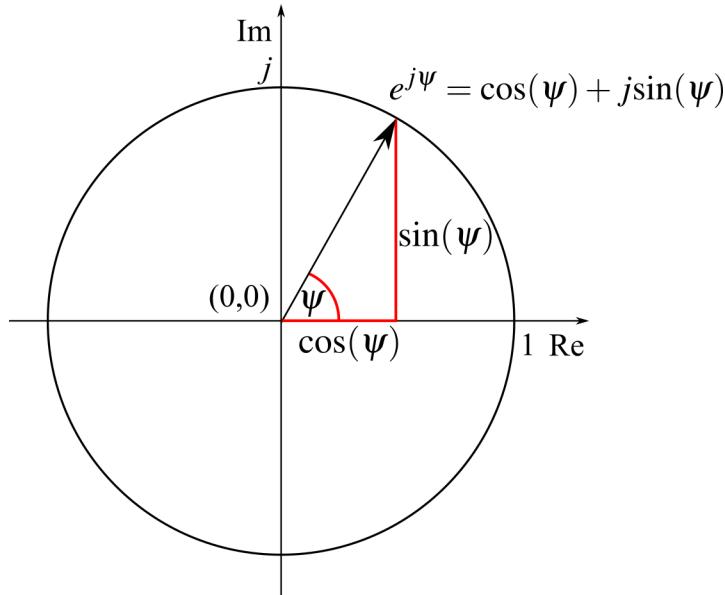


Figure 2.17: Euler's formula illustrated on the unit circle in the complex plane.

### Proof

Showing that  $x_c(t) = c_1 e^{i\omega_n t} + c_2 e^{-i\omega_n t}$  is the same as  $x_c(t) = C \sin(\omega_n t + \phi)$ , consider Euler's identity,

$$e^{i\alpha} = \cos(\alpha) + i\sin(\alpha) \quad (51)$$

therefore:

$$\begin{aligned} c_1 e^{i\omega_n t} &= c_1 \cos(\omega_n t) + c_1 i \sin(\omega_n t) \\ c_2 e^{-i\omega_n t} &= c_2 \cos(\omega_n t) + c_2 i \sin(\omega_n t) \end{aligned} \quad (52)$$

rearranging terms yields:

$$c_1 e^{i\omega_n t} + c_2 e^{-i\omega_n t} = (c_1 + c_2) \cos(\omega_n t) + (c_1 + c_2) i \sin(\omega_n t) \quad (53)$$

which is equivalent to:

$$c_1 e^{i\omega_n t} + c_2 e^{-i\omega_n t} = A \cos(\omega_n t) + B \sin(\omega_n t) \quad (54)$$

where:

$$A = c_1 + c_2; \quad B = i(c_1 - c_2) \quad (55)$$

Recall that  $\sin(\alpha + \beta) = \sin(\alpha)\cos(\beta) + \cos(\alpha)\sin(\beta)$ . Therefore,  $x_c(t) = c \sin(\omega_n t + \phi)$  can be written as (with a little rearranging):

$$x_c(t) = c \cos(\phi) \sin(\omega_n t) + c \sin(\phi) \cos(\omega_n t) \quad (56)$$

Therefore we can show that:

$$A \cos(\omega_n t) + B \sin(\omega_n t) = c \sin(\omega_n t + \phi) \quad (57)$$

as

$$B = c \cos(\phi); \quad A = c \sin(\phi) \quad (58)$$

Moreover,

$$A^2 + B^2 = C^2 \sin^2(\phi) + C^2 \cos^2(\phi) = C^2 (\sin^2(\phi) + \cos^2(\phi)) = C^2 \quad (59)$$

and

$$C = \sqrt{A^2 + B^2}, \quad \frac{A}{B} = \frac{c \sin(\phi)}{c \cos(\phi)} = c \tan(\phi), \quad \phi = \tan^{-1} \frac{A}{B} \quad (60)$$

The Standard form of Inhomogeneous Equation can be found, starting at the EOM:

$$m\ddot{x} + kx = f^* \quad (61)$$

divide by  $m$ , therefore:

$$\ddot{x} + \frac{k}{m}x = \frac{1}{m}f^* \quad (62)$$

Recall that  $\frac{k}{m} = \omega_n^2$  and define the normalized forcing function as  $f(t) = \frac{1}{k}f^*$ , this results in:

$$\frac{1}{m}f^*(t) = \frac{1}{m}kf(t) = \omega_n^2 f(t) \quad (63)$$

inserting this back into the the EOM yields a 2<sup>nd</sup>-order inhomogeneous ODE in standard form:

$$\ddot{x} + \frac{k}{m}x = \omega_n^2 f(t) \quad (64)$$

## 2.7 Forced Vibration Response

The total vibration  $x(t)$  of the system can be explained as:

$$x(t) = x_c(t) + x_p(t) = C \sin(\omega_n t + \phi) + x_p(t) \quad (65)$$

inserting the first part of this expression for  $x(t)$  into equation 64 yields:

$$\ddot{x}_c + \ddot{x}_p + \omega_n^2(x_c + x_p) = \omega_n^2 f(t) \quad (66)$$

assuming that the complementary solution equals zero, this expression simplifies to:

$$\ddot{x}_p + \omega_n^2 x_p = \omega_n^2 f(t) \quad (67)$$

This expression requires that we find  $c$ ,  $\phi$ , and  $x_p(t)$ ; this is not easy.

## 2.8 Spring-mass-damper oscillator

Consider the system:

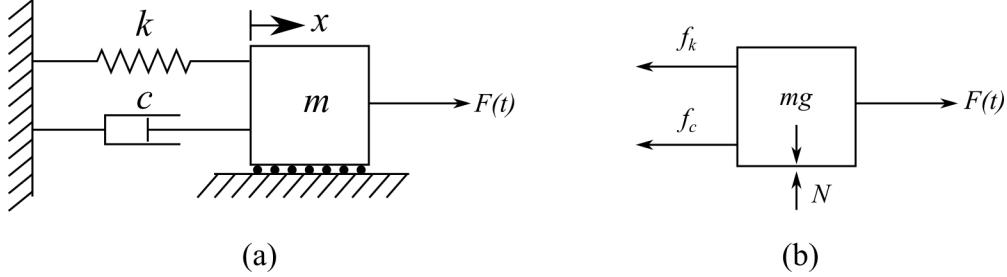


Figure 2.18: Damped 1-DOF system with an external force ( $F(t)$ ) applied, showing: (a) the system configuration; and (b) the free body diagram

Newton's second law of motion tells us

$$m\ddot{x} = -kx - cx + f^* \quad (68)$$

therefore, we can write the EOM as:

$$m\ddot{x}(t) + c\dot{x}(t) + kx(t) = f^*(t) \quad (69)$$

for the considered damped forced vibration problem. This is a 2<sup>nd</sup> order inhomogeneous ODE. As before, the solution of  $x(t)$  for equation 69 is the sum of the complementary and particular solution where the complementary solution satisfies the homogeneous equation while the particular solution satisfies the inhomogeneous equation.

To solve the homogeneous equation, set the right hand side of equation 69 to zero to get the equation for damped free vibration:

$$m\ddot{x}(t) + c\dot{x}(t) + kx(t) = 0 \quad (70)$$

Again, assume  $x(t) = Ce^{pt}$ , therefore,

$$\begin{aligned} \dot{x}(t) &= pce^{pt} = px \\ \ddot{x}(t) &= p^2ce^{pt} = p^2x \end{aligned} \quad (71)$$

Putting these expressions back into the EOM yields:

$$mp^2x + cpx + kx = 0 \quad (72)$$

or the characteristic equation

$$mp^2 + cp + k = 0 \quad (73)$$

solving this expression leads to:

$$p_{1,2} = \frac{-c \pm \sqrt{c^2 - 4mk}}{2m} = -\frac{c}{2m} \pm \sqrt{\frac{c}{2m} - \frac{k}{m}} \quad (74)$$

or:

$$p_{1,2} = -\frac{c}{2m} \pm i\sqrt{\frac{k}{m} - \left(\frac{c}{2m}\right)^2} \quad (75)$$

The critical damping value  $c_{cr}$  for the system is defined as the value of  $c$  that results in a 0 for the radicand (number under the square root). Therefore, we need:

$$\frac{k}{m} - \left(\frac{c_{cr}}{2m}\right)^2 = 0 \quad (76)$$

$$c_{cr} = \frac{km^2k}{m} = 4mk \quad (77)$$

this results in:

$$c_{cr} = 2\sqrt{mk} \quad (78)$$

The damping ratio  $\zeta$  is the ratio between the damping  $c$  and the critical damping ratio  $c_{cr}$ , i.e.,

$$\zeta = \frac{c}{c_{cr}} = \frac{c}{2\sqrt{mk}} = \frac{c}{2m\omega_n} \quad (79)$$

Recall that the natural frequency of a system is  $\omega_n = \sqrt{k/m}$ . Damping slightly reduces the natural frequency of the system, resulting in the damped natural frequency  $\omega_d$

$$\omega_d = \omega_n \sqrt{1 - \zeta^2} \quad (80)$$

Using this expression for the damped natural frequency, the poles of the system can be identified as:

$$\begin{aligned} p_{1,2} &= -\zeta\omega_n \pm i\sqrt{\omega_n^2 - \zeta^2\omega_n^2} \\ &= -\zeta\omega_n \pm i\sqrt{1 - \zeta^2} \\ &= -\zeta\omega_n \pm i\omega_d \end{aligned} \quad (81)$$

The damped free vibration response is:

$$x(t) = c_1 e^{(-\zeta\omega_n + i\omega_d)t} + c_2 e^{(-\zeta\omega_n - i\omega_d)t} \quad (82)$$

This simplifies to:

$$x(t) = e^{-\zeta \omega_n t} (c_1 e^{i \omega_d t} + c_2 e^{-i \omega_d t}) \quad (83)$$

or using Euler's equation,

$$x_c(t) = C e^{-\zeta \omega_n t} \sin(\omega_d t + \phi) \quad (84)$$

This equation breaks down into two parts, one for the decay of the system and one for the oscillations as shown below:

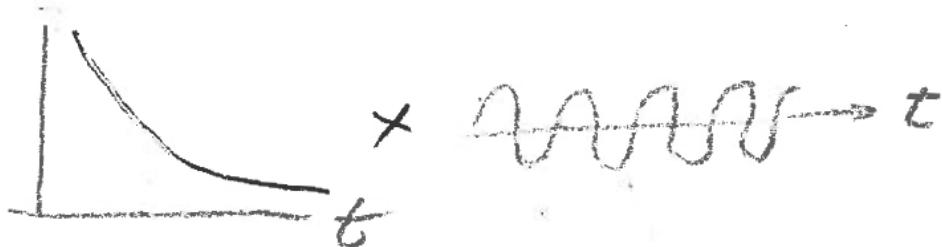


Figure 2.19: Decay and oscillations of the system.

Together, these make the free damped vibration response, as computed by the complementary solution  $x_c(t)$

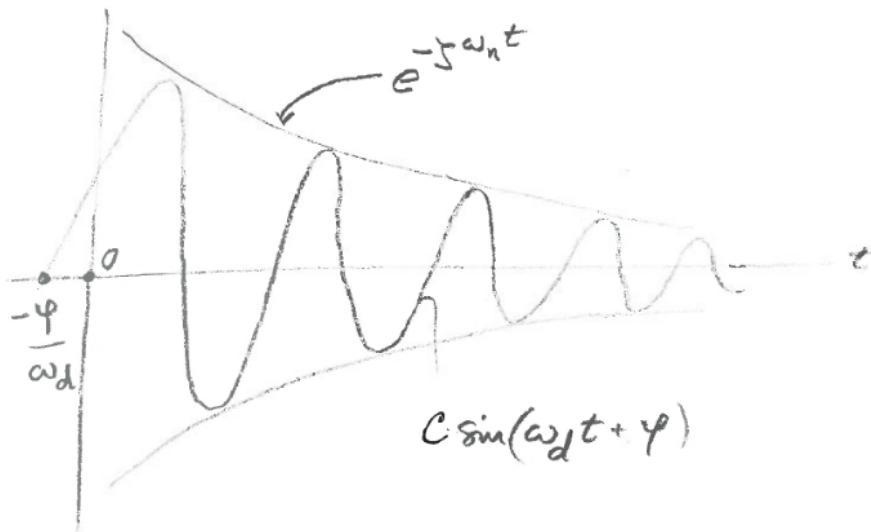


Figure 2.20: Free decay with oscillations.

Generating the standard form of the damped forced vibration problem is done by dividing the EOM by what leads the first term, mainly:

$$\ddot{x} + \frac{c}{m} \dot{x} + \frac{k}{m} x = \frac{1}{m} f^*(t) \quad (85)$$

The expression  $c/m$  can be rewritten as:

$$\frac{c}{m} = \frac{\zeta c_{cr}}{m} = \frac{\zeta 2\sqrt{mk}}{m} = 2\zeta \omega_n \quad (86)$$

while the previously defined normalized forcing function is  $f(t) = \frac{1}{k}f^*(t)$ , this results in a right hand side of EOM written as:

$$\frac{1}{m}f^*(t) = \frac{1}{m}kf(t) = \omega_n^2 f(t) \quad (87)$$

therefore, the standard form of the EOM for the 1-DOF damped forced vibration problem is written as a 2<sub>nd</sub> order ODE in standard form:

$$\ddot{x} + 2\zeta\omega_n\dot{x} + \omega_n^2 x = \omega_n^2 f(t) \quad (88)$$

The timed resolved solution for the forced vibration problem is:

$$\begin{aligned} x(t) &= x_c(t) + x_p(t) \\ &= e^{-\zeta\omega_n t} C \sin(\omega_d t + \phi) + x_p(t) \end{aligned} \quad (89)$$

Putting equation 89 back into equation 88 results in the expression:

$$(\ddot{x}_c + 2\zeta\omega_n\dot{x}_c + \omega_n^2 x_c) + \ddot{x}_p + 2\zeta\omega_n\dot{x}_p + \omega_n^2 x_p = \omega_n^2 f(t) \quad (90)$$

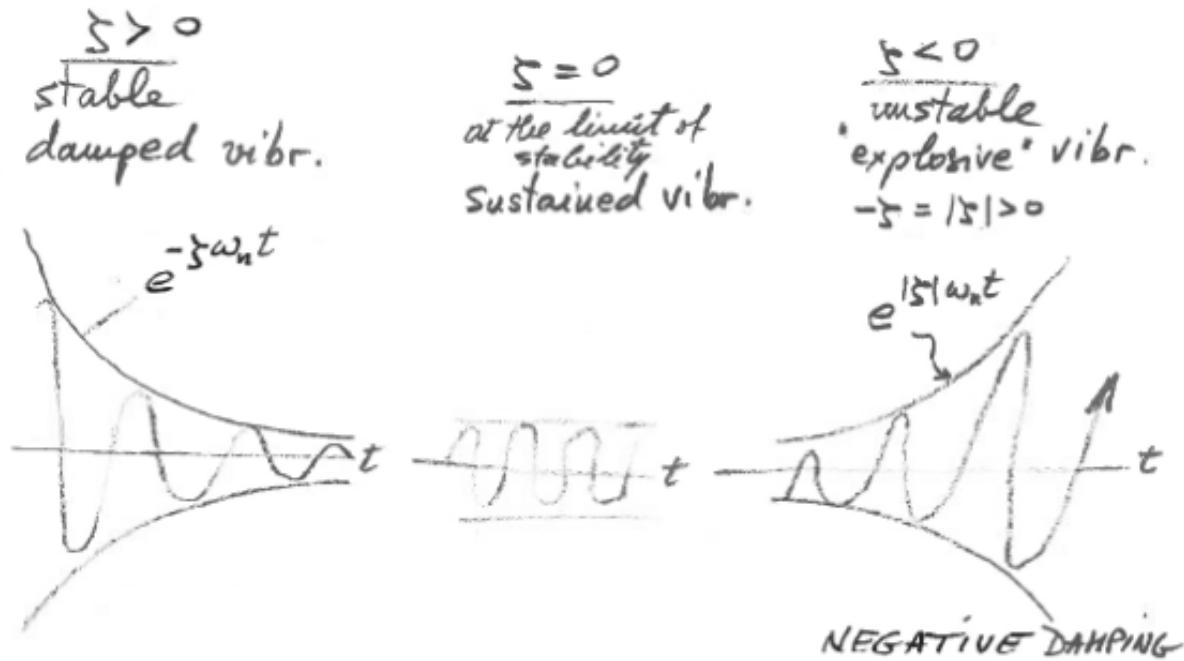
where we need to find  $C$ ,  $\phi$ ,  $x_p(t)$  based on the initial conditions of the system. This is not an easy task.

I don't like the rest of this, why not just use the method of undetermined coefficients?

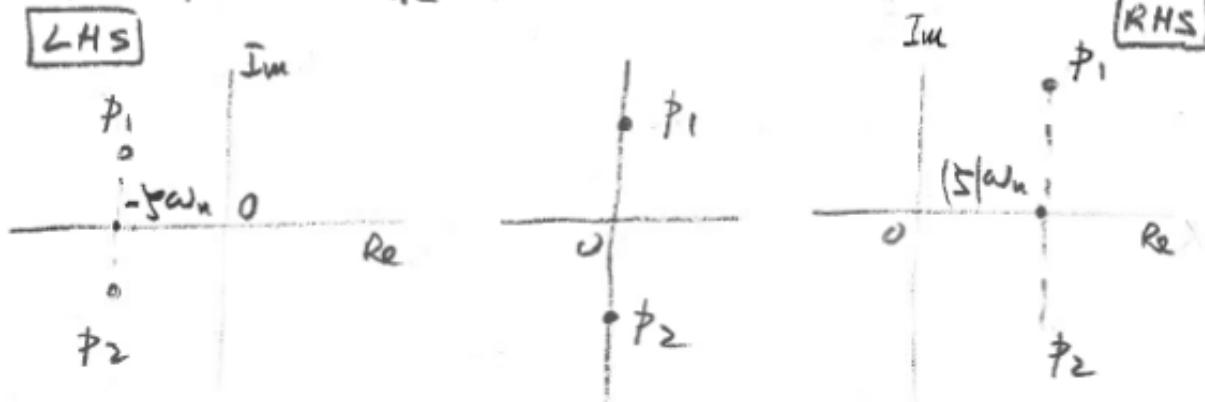
## 2.9 Stability of 2nd order systems

Recall that the damped free vibration response:

$$x_c(t) = Ce^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \quad (91)$$



Complex poles:  $p_{1,2} = -\zeta \omega_n \pm i \omega_d$  (poles)



$p_{1,2}$  in LHS

of complex  
 $\lambda$  plane

$p_{1,2}$  in RHS

UNSTABLE!

Figure 2.21: Stability of 2<sup>nd</sup> order systems.

Underdamped, critical-damped, overdamped 2<sup>nd</sup> order systems and vibration response.

$$p_{1,2} = -\zeta \omega_n \pm i \sqrt{1 - \zeta^2} \quad (92)$$

$0 < \zeta < 1$  underdamped vibration resp.

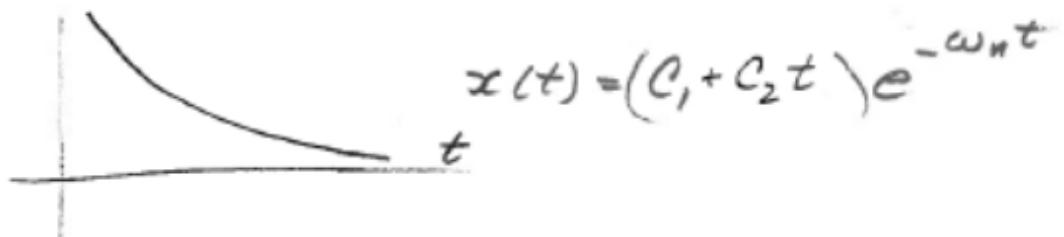
$$x(t) = C e^{-\zeta \omega_n t} \sin(\omega_d t + \varphi),$$

$$\omega_d = \omega_n \sqrt{1 - \zeta^2}$$



$\zeta = 1$  critically damped response

$$\tau_1 = \tau_2 = -\omega_n$$



$\zeta > 1$  overdamped response

$$\tau_1 = -\zeta \omega_n + \omega_n \sqrt{\zeta^2 - 1}$$

$$= (-\zeta + \sqrt{\zeta^2 - 1}) \omega_n < 0$$

$$\tau_2 = (-\zeta - \sqrt{\zeta^2 - 1}) \omega_n < 0$$

$$x(t) = C_1 e^{-\lambda_1 t} + C_2 e^{-\lambda_2 t}$$

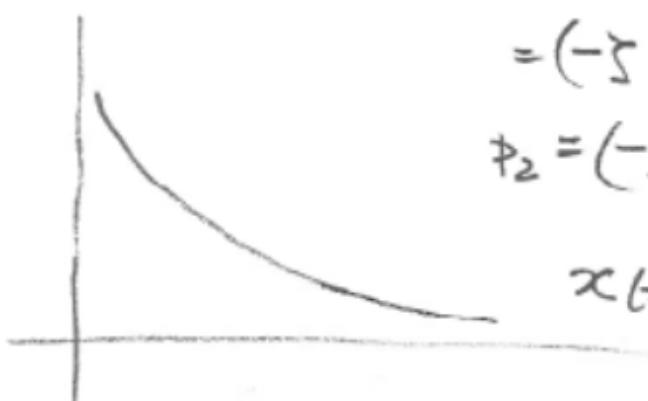


Figure 2.22: vibration responses.

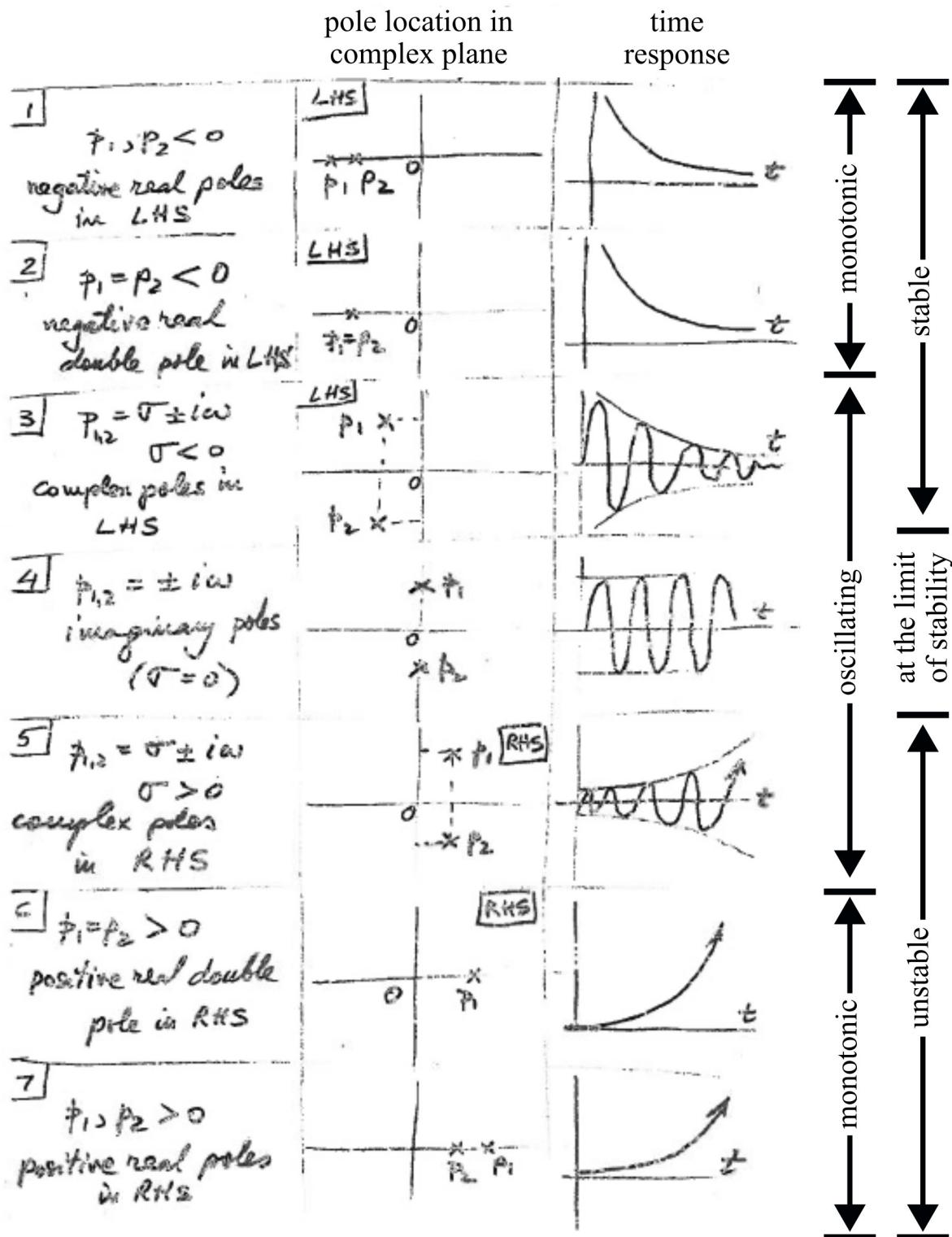


Figure 2.23: 2nd order poles.

**Example 2.3** Do example on stability of 2<sup>nd</sup> order systems

### 3 Transfer Functions

Thus-far, this text has only considered forced vibrations for 1-DOF systems excited with forcing functions that can be easily expressed using either sin or cos examples. Therefore, the previously developed solutions are only acceptable for systems with known and simple excitations. This chapter will introduce the concept of transfer functions for solving vibration related problems. The transfer function, in particular the Laplace transfer function, is an important tool in the study of vibrations as it allows the practitioner to solve for the temporal response of a system for a variety of inputs using a single approach. Examples of force excitation that can be calculated include using this method include:

- sinusoidal
- base excitation
- impulse
- arbitrary input
- arbitrary periodic input

Consider the following system

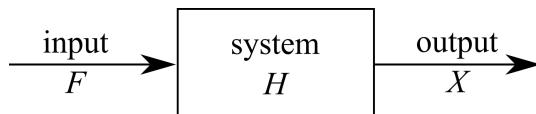


Figure 3.1: Generic system  $H$  subjected to an input  $F$  and its corresponding output  $X$ .

where  $F$  is the input,  $H$  is the system, and  $X$  is the output from the system. This formulation is called the transfer-function approach and is commonly used for the formulation and solution of dynamic problems in the control literature. It can also be used for solving the various forced-vibration problems including those from complex or stochastic inputs.

#### 3.1 Laplace Transforms

**Review 3.1** Laplace transforms, or more broadly integral transform, are a procedure for integrating the time ( $t$ ) dependence of a function into a function of position or space ( $s$ ). By transforming the whole differential equation from the time domain into a lower order function of space the problem becomes easier to solve as the function can often be manipulated algebraically.



Figure 3.2: Portrait of Pierre-Simon Laplace by Johann Ernst Heinsius (1775).<sup>1</sup>

The Laplace transform is named after mathematician and astronomer Pierre-Simon Laplace (23 March 1749 - 5 March 1827). Pierre-Simon Laplace was one of the greatest scientists of all time and is often considered the French Newton. He taught Napoleon at the École Militaire in 1784, became a count of the empire in 1806, and a marquis in 1817 after the restoration of the monarchy. He is credited with advancements in engineering, mathematics, statistics, physics, astronomy, and philosophy; however, maybe his greatest achievement is not only surviving but benefiting from the change from the Ancien Régime → Bonaparte → Bourbon Restoration.

Of interest to this class is the Laplace transform ( $\mathcal{L}[ ]$ ) of the function  $f(t)$ , expressed as  $\mathcal{L}[f(t)]$ . Here, a Laplace transform is used as a method of solving the differential equations of motion by reducing the computation needed to that of integration and algebraic manipulation.

The definition of the Laplace transform of the function  $f(t)$  is:

$$\mathcal{L}[f(t)] = F(s) = \int_0^{\infty} f(t)e^{-st} dt \quad (93)$$

where  $s$  represents a variable in the complex plane (also called the  $s$ -plane) and  $f(t) = 0$  for all values of  $t < 0$ . Here, the  $s$  is a complex value. Lastly, the term  $F(s)$  is a generic term that represents the input to a system. As this class needs the derivatives of the base function, we will calculate these next:

$$\mathcal{L}[\dot{f}(t)] = \int_0^{\infty} \dot{f}(t)e^{-st} dt = \int_0^{\infty} e^{-st} \frac{d[f(t)]}{dt} dt \quad (94)$$

integration by parts yields:

$$\mathcal{L}[\dot{f}(t)] = e^{-st} f(t) \Big|_0^\infty + s \int_0^\infty e^{-st} f(t) dt \quad (95)$$

Astutely, it can be noticed that the second term  $s \int_0^\infty e^{-st} f(t) dt$  is the input to the system  $F(s)$ . Therefor, with a little rearranging this becomes:

$$\mathcal{L}[\dot{f}(t)] = sF(s) - f(0) \quad (96)$$

Taking the derivative of again yields:

$$\mathcal{L}[\ddot{f}(t)] = s^2 F(s) - sf(0) - \dot{f}(0) \quad (97)$$

A few key points of the Laplace transforms are:

- The domain of the problem changes from the real number line ( $t$ ) to the complex plane ( $s$ -plane).
- The integration of the Laplace transform changes differentiation into multiplication.
- The transform procedure is linear. Therefore, the transform of the linear combination of two transforms is the same as the linear transformation of these functions.
- To move from the time domain to the complex number plane we typically use tables of pre-solved integral.
- The function  $x(t)$  can be obtained by taking the inverse Laplace transform defined as  $x(t) = \mathcal{L}[X(s)]^{-1}$

The Laplace transform can be calculated in symbolic form. In particular interest to this text is the Laplace form of the system input  $F(s)$  and output  $X(s)$ . To expand the symbolic form of the Laplace transform for the system inputs are and for system outputs:

$$\mathcal{L}[f(t)] = F(s) \quad (98)$$

$$\mathcal{L}[\dot{f}(t)] = sF(s) - f(0) \quad (99)$$

$$\mathcal{L}[\ddot{f}(t)] = s^2 F(s) - sf(0) - \dot{f}(0) \quad (100)$$

here,  $f(0)$  and  $\dot{f}(0)$  are the initial values of the function  $f(t)$ . Furthermore, the for system outputs are:

$$\mathcal{L}[x(t)] = X(s) \quad (101)$$

$$\mathcal{L}[\dot{x}(t)] = sX(s) - x(0) \quad (102)$$

$$\mathcal{L}[\ddot{x}(t)] = s^2X(s) - sx(0) - \dot{x}(0) \quad (103)$$

here,  $x(0)$  and  $\dot{x}(0)$  are the initial values of the function  $x(t)$ .

<sup>a</sup>Johann Ernst Heinsius, CC BY-SA 4.0 <<https://creativecommons.org/licenses/by-sa/4.0/>>, via Wikimedia Commons

## 3.2 Transfer Function Method for Solving Vibrating Systems

As mentioned in the introduction to this chapter, a variety of systems can be solved for using the transfer function method. The procedure for using the Laplace transform to solve equations of motion expressed as an inhomogeneous ordinary differential equation is:

1. Take the Laplace transform of both sides of the EOM while treating the time derivatives symbolically.
2. Solve for  $X(s)$  in the obtained equation.
3. Apply the inverse transform  $x(t) = \mathcal{L}[X(s)]^{-1}$

### 3.2.1 Free Vibration for Undamped Systems

Consider the undamped single-DOF system:

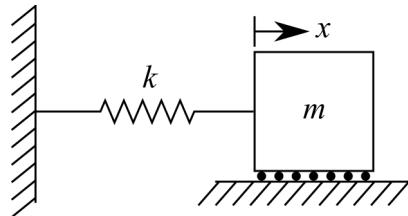


Figure 3.3: A spring mass model of a 1-DOF system.

The EOM for this system is a homogeneous differential equation because the right-hand side is equal to zero:

$$m\ddot{x}(t) + kx(t) = 0 \quad (104)$$

Here we will leave the “( $t$ )” for clarity to differentiate the time domain solution from Laplace solution “( $s$ )” in the  $s$ -plane, as discussed in review 3.1. The EOM can be rewritten in standard form as:

$$\ddot{x}(t) + \omega_n^2 x(t) = 0 \quad (105)$$

where the initial conditions at  $t = 0$  are  $x(0) = x_0$  and  $\dot{x}(0) = v_0$ . Taking the Laplace transforms, in symbolic form using equations 101 - 103, of both sides of the EOM yields:

$$[s^2X(s) - sx_0 - v_0] + [\omega_n^2 X(s)] = 0 \quad (106)$$

using equations 101 and 103 from section 3.1. Solving for the output of the system  $X(s)$  yields:

$$X(s) = \frac{sx_0 + v_0}{s^2 + \omega_n^2} \quad (107)$$

We can expand this form of  $X(s)$  to obtain equations listed in our Laplace Transform table:

$$X(s) = \frac{sx_0}{s^2 + \omega_n^2} + \frac{v_0}{s^2 + \omega_n^2} \cdot \frac{\omega_n}{\omega_n} \quad (108)$$

This becomes:

$$X(s) = x_0 \frac{s}{s^2 + \omega_n^2} + \left( \frac{v_0}{\omega_n} \right) \cdot \frac{\omega_n}{s^2 + \omega_n^2} \quad (109)$$

Next, using the inverse Laplace transform  $x(t) = \mathcal{L}[X(s)]^{-1}$  and the two following Laplace transforms (#5 and #6):

$$f(t) \text{ is } \cos(\omega t) \text{ when } F(s) \text{ is } \frac{s}{s^2 + \omega^2} \quad (110)$$

$$f(t) \text{ is } \sin(\omega t) \text{ when } F(s) \text{ is } \frac{\omega}{s^2 + \omega^2} \quad (111)$$

Therefore, we can obtain the solution for the system output  $X(s)$  as:

$$x(t) = x_0 \cos(\omega_n t) + \frac{v_0}{\omega_n} \sin(\omega_n t) \quad (112)$$

The same procedure can be used to calculate the under damped and forced responses. However, when calculating these responses the algebraic solution for  $X(s)$ ,  $s$  often contains quotients of polynomials. These Polynomial ratios may not be found in simple Laplace tables and must be solved using the method of partial fractions. An example of this procedure can be found in Appendix B of Inman.

### 3.2.2 Forced Vibration (Impulse) for damped Systems

Shock loads on mechanical systems represent a very common source of vibration. These short-duration forces are also called an impulse. An impulse excitation is defined as a force that is applied for a very short, or infinitesimal, length of time. An impulse is a nonperiodic force that is represented by the symbol  $\delta$ . The response of a system to an impulse load is the same as the system's free response provided that the correct initial conditions are applied. This is illustrated in the following where the applied force  $F(t)$  is impulsive in nature (i.e., large magnitude over a very short time).

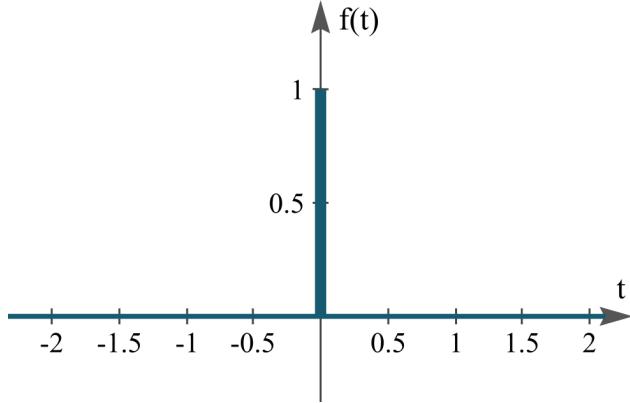


Figure 3.4: An impulse function with the impulse at  $t = 0$ .

The impulse response function can be solved for analytically, however, we will solve it using the transfer function approach. Here we will consider the under-damped spring-mass system. First, assume that the system is at rest (no initial conditions). Next, we write the EOM as:

$$m\ddot{x} + c\dot{x} + kx = \delta(t) \quad (113)$$

Taking the Laplace transform of both sides of the equation yields

$$m(s^2X(s) - sx(0) - \dot{x}(0)) + c(sX(s) - x(0)) + kX(s) = 1 \quad (114)$$

note that the  $\mathcal{L}[\delta] = 1$  per #1 in the transform table. However, if we assume zero initial conditions (a system at rest when the impulse happens), the equation simplifies to.

$$ms^2X(s) + csX(s) + kX(s) = 1 \quad (115)$$

or

$$(ms^2 + cs + k)X(s) = 1 \quad (116)$$

Solving this equation for  $X(s)$ :

$$X(s) = \frac{1}{m} \cdot \frac{1}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (117)$$

Again, the mass is extracted to develop a formulation that can be found in the Laplace tables. Setting the constraint that  $\zeta < 1$  and consulting #10 in the table for Laplace transforms results in:

$$x(t) = \frac{1}{m\omega_d} e^{-\zeta\omega_n t} \sin(\omega_d t) \quad (118)$$

where this is the general solution for a damped system subjected to an impulse loading function. For the undamped case a solution can be obtained by setting  $\zeta = 0$ . This Results in the following form for the undamped case:

$$x(t) = \frac{1}{m\omega_n} \sin(\omega_n t) \quad (119)$$

Below is a typical response for both a undamped and underdamped 1-DOF system subject to an impulse response at  $t = 0$  seconds.

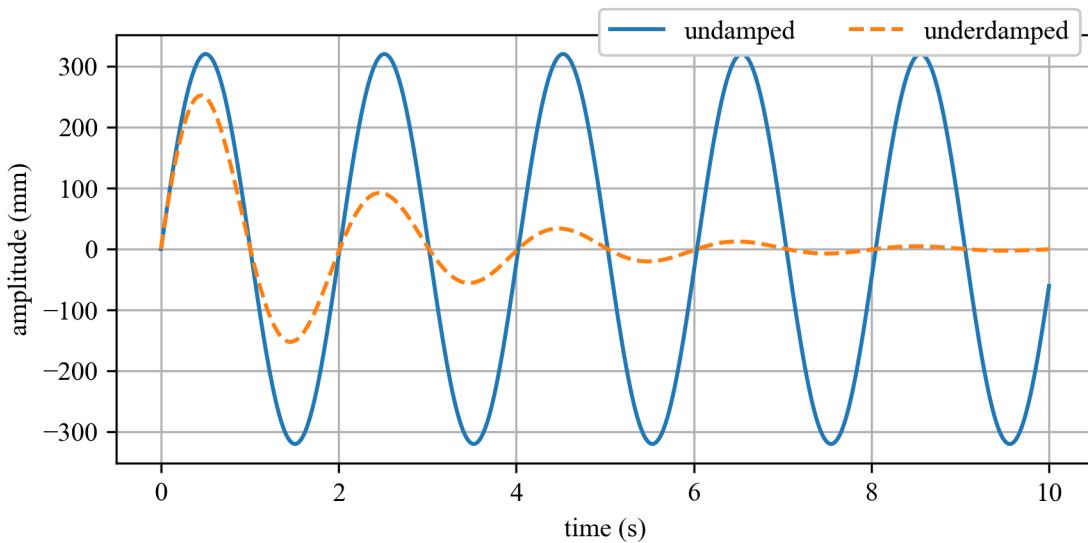


Figure 3.5: Temporal responses from a underdamped and undamped 1-DOF systems to a impulse response function.

### 3.2.3 Arbitrary Inputs to a System

The time domain response of a system to an arbitrary input force in time can be calculated using a series of impulses as shown in figure 3.6. This method allows the practitioner to easily calculate the response of an arbitrary input to a system using a single expression executed in a “for loop”. This type of analysis is often more efficient in terms of programming than more direct methods such as the transfer functions sown in this text.

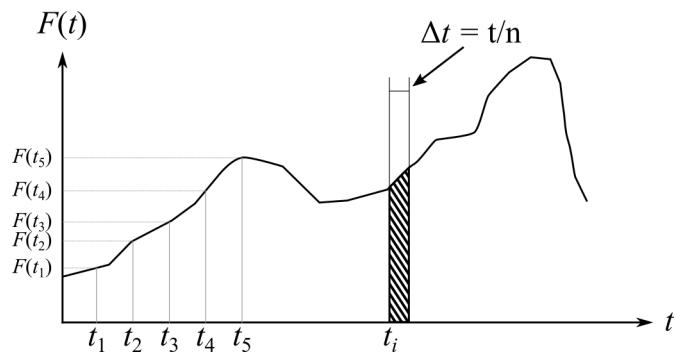


Figure 3.6: Generalized response showing that any signal can be represented as a series of impulse signals.

**Example 3.1** In testing, an hammer is used to excite a 1-DOF system with an impact (i.e. impulse), however, the hammer ascendantly impacts the system twice. The first impact has a force of 0.2 N, while the second has a force of 0.1 N and happens 0.1 seconds after the first impact. Plot the response for the double impact. The system has the parameters  $m = 1 \text{ kg}$ ,  $c = 0.5 \text{ kg/s}$ ,  $k = 4 \text{ N/m}$ .

**Solution:** First, we can define the forcing function as:

$$F(t) = 0.2\delta(t) + 0.1\delta(t - \tau) \quad (120)$$

where  $\tau$  is the offset between the first and second impacts. Next, considering that the unit impulse has a magnitude of 1 we can obtain solutions for the first impact by first writing it's EOM:

$$m\ddot{x}(t) + c\dot{x}(t) + kx(t) = 0.2\delta(t) \quad (121)$$

Taking the Laplace transform of both sides of the equation yields

$$m(s^2X(s) - sx(0) - \dot{x}(0)) + c(sX(s) - x(0)) + kX(s) = 0.2 \quad (122)$$

However, assuming zero initial conditions, the equation simplifies to.

$$(ms^2 + cs + k)X(s) = 0.2 \quad (123)$$

Solving this equation for  $X(s)$ :

$$X(s) = \frac{0.2}{m} \cdot \frac{1}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (124)$$

Again, consulting #10 in the table for Laplace transforms results in:

$$x_1(t) = \frac{0.2}{m\omega_d} e^{-\zeta\omega_n t} \sin(\omega_d t) \quad (125)$$

where this is the general solution for a damped system subjected to an impulse loading function. The second impact can now be solved for using the same method. However, now the time ( $t$ ) must be offset by ( $\tau$ ) to allow the impact to still be located at  $t = 0$  in terms of the second impact. This results in:

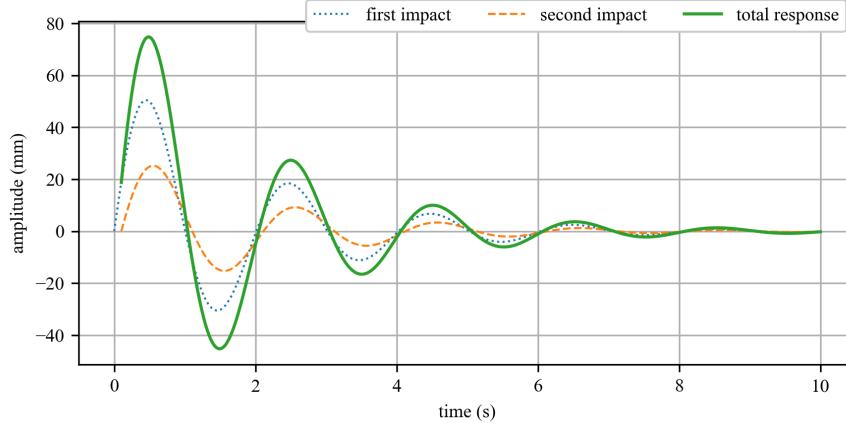
$$x_1(t) = \frac{0.2}{m\omega_d} e^{-\zeta\omega_n t} \sin(\omega_d t) \quad (126)$$

$$x_2(t) = \frac{0.1}{m\omega_d} e^{-\zeta\omega_n(t-\tau)} \sin(\omega_d(t-\tau)) \quad (127)$$

Next, using the knowledge that the systems are linear and that the Laplace transform of a linear combination of two transforms is the same as the linear transformation of these functions we can build the piecewise function:

$$x(t) = \begin{cases} \frac{0.2}{m\omega_d} e^{-\zeta\omega_n t} \sin(\omega_d t) & \text{if } t < \tau \\ \frac{0.2}{m\omega_d} e^{-\zeta\omega_n t} \sin(\omega_d t) + \frac{0.1}{m\omega_d} e^{-\zeta\omega_n(t-\tau)} \sin(\omega_d(t-\tau)) & \text{if } \tau \leq t \end{cases}$$

For the mass, damping, and stiffness values given above this can be plotted as:



### 3.3 Laplace method in Controls

Again, consider the Laplace system:

$$\mathcal{L}[f(t)] = F(s) = \int_0^\infty f(t)e^{-st} dt \quad (128)$$

where:

$$\int_0^\infty = \lim_{p \rightarrow \infty} \int_0^p \quad (129)$$

exists. For now,  $s \in \mathbb{R}$  (real). Later, we will let  $s \in \mathbb{C}$  (complex). For any given  $F(s)$ , we can find the original  $f(t)$  using the pairing method.

$$f(t) \xrightleftharpoons[ILT]{LT} F(s) \quad (130)$$

this can also be written as:

$$\text{LT pair} = \begin{cases} F(s) & \mathcal{L}[f(t)] \\ f(t) & \mathcal{L}[F(s)]^{-1} \end{cases} \quad (131)$$

#### 3.3.1 Laplace transform of exponential function

Consider the exponential function,

$$\mathcal{L}[e^{p_0 t}] = \frac{1}{s - p_0} \quad (132)$$

where  $f(t) = e^{p_0 t}$  and  $F(s) = \frac{1}{s - p_0}$ . As  $F(s) \rightarrow \infty$ ,  $s \rightarrow p_0$ , therefore  $p_0$  is a pole of the system. This is shown in Fig. 3.7.

$$\text{LT pair} = \begin{cases} f(t) & e^{p_0 t} \\ F(s) & \frac{1}{s-p_0} \end{cases} \quad (133)$$

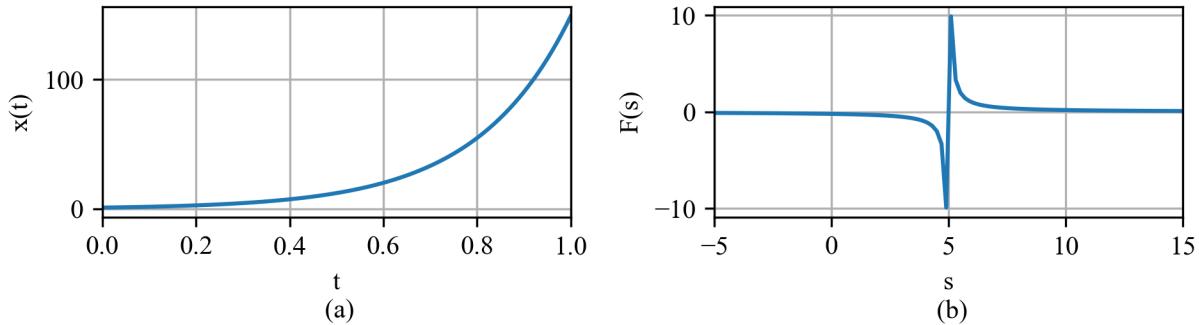


Figure 3.7: Function  $e^{p_0 t}$  with  $p_0 = 5$ ; showing the (a) time domain and; (c) the s-space.

### Proof

Show that

$$\mathcal{L}[e^{p_0 t}] = \frac{1}{s - p_0} \quad (134)$$

Again, consider the Laplace system:

$$\mathcal{L}[f(t)] = F(s) = \int_0^{\infty} f(t) e^{-st} dt \quad (135)$$

therefore,

$$\int_0^{\infty} f(t) e^{-st} dt = \int_0^{\infty} e^{p_0 t} e^{-st} dt = \int_0^{\infty} e^{-(s-p_0)t} dt \quad (136)$$

Change of variables can be used such that:

$$\begin{aligned} (s - p_0)t &= t^* \\ (s - p_0)dt &= dt^* \\ dt &= \frac{1}{s - p_0} dt^* \end{aligned} \quad (137)$$

therefore

$$\begin{aligned}
 \int_0^\infty e^{-(s-p_0)t} dt &= \int_0^\infty e^{-t^*} \frac{1}{s-p_0} dt^* \\
 &= \frac{1}{s-p_0} \int_0^\infty e^{-t^*} dt^* \\
 &= \frac{1}{s-p_0} (-e^{-t^*}) \Big|_0^\infty \\
 &= \frac{-1}{s-p_0} (e^{-t^*}) \Big|_0^\infty \\
 &= \frac{-1}{s-p_0} [e^{-\infty} - e^0] \\
 &= \frac{-1}{s-p_0} (-1) \\
 &= \frac{1}{s-p_0}
 \end{aligned} \tag{138}$$

Therefore, it can be shown that:

$$\mathcal{L}[e^{p_0 t}] = \frac{1}{s-p_0} \tag{139}$$

### 3.3.2 Laplace transform of step function

$$\text{LT pair} = \begin{cases} f(t) & 1(t), \quad t > 0 \\ F(s) & \frac{1}{s} \end{cases} \tag{140}$$

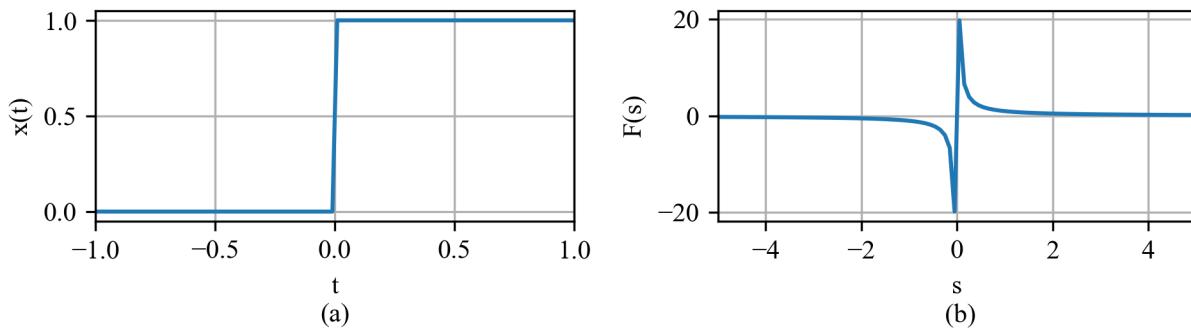


Figure 3.8: Step function; showing the (a) time domain and; (c) the s-space.

#### Proof

Show that

$$\mathcal{L}[1(t)] = \frac{1}{s} \tag{141}$$

This can be done as:

$$\begin{aligned}
 \mathcal{L}[1(t)] &= \int_0^\infty 1 \cdot e^{-st} dt \\
 &= \int_0^\infty e^{-st} dt \\
 &= \left. \frac{-1}{s} e^{-st} \right|_0^\infty \\
 &= \frac{1}{s}
 \end{aligned} \tag{142}$$

### 3.3.3 Laplace transform of ramp function

$$\text{LT pair} = \begin{cases} f(t) & t, \quad t > 0 \\ F(s) & \frac{1}{s^2} \end{cases} \tag{143}$$

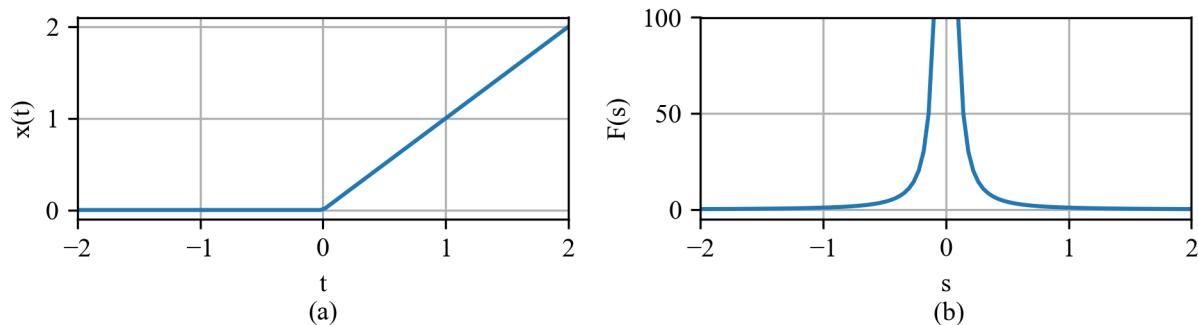


Figure 3.9: Ramp function; showing the (a) time domain and; (c) the s-space.

#### Proof

Show that

$$\mathcal{L}[t] = \frac{1}{s^2} \tag{144}$$

This can be done as:

$$\mathcal{L}[t] = \int_0^\infty t \cdot e^{-st} dt \tag{145}$$

Integration by parts leads to:

$$d[uv] = u dv + v du \rightarrow v du = d[uv] - u dv \tag{146}$$

where

$$\int_a^b v du = uc \Big|_a^b - \int_a^b u dv \tag{147}$$

which leads to:

$$du = e^{-st}; \quad u = \frac{1}{-s}e^{-st} \quad (148)$$

or more simply,

$$du = 1; \quad u = t \quad (149)$$

Using these expressions, we get

$$\begin{aligned} \int_0^\infty t \cdot e^{-st} dt &= \frac{t}{-s} e^{-st} \Big|_0^\infty - \int_0^\infty \frac{-1}{-s} e^{-st} dt \\ &= \frac{1}{s} \int_0^\infty e^{-st} dt \\ &= \left[ \frac{1}{s} \right] \left[ \frac{1}{s} \right], \text{ as } \mathcal{L}[1(t)] = \int_0^\infty e^{-st} dt = \frac{1}{s} \\ &= \frac{1}{s^2} \end{aligned} \quad (150)$$

### 3.3.4 Laplace transform shifted step function

$$\text{LT pair} = \begin{cases} f(t) & 1(t - \tau) \\ F(s) & e^{-\tau s} \frac{1}{s} \end{cases} \quad (151)$$

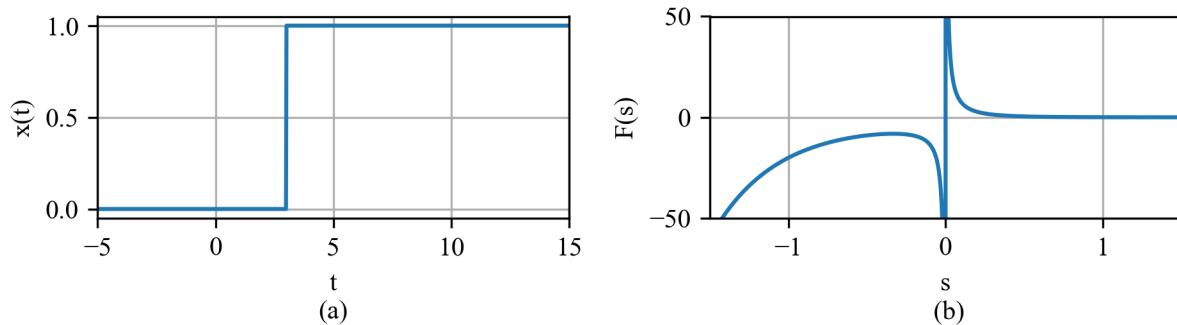


Figure 3.10: Shifted step function; showing the (a) time domain and; (c) the s-space.

#### Proof

Show that

$$\mathcal{L}[1 \cdot (t - \tau)] = e^{-\tau s} \frac{1}{s} \quad (152)$$

This can be done as:

$$(153)$$

$$\begin{aligned}
 \mathcal{L}[1 \cdot (t - \tau)] &= \int_0^\infty 1(t - \tau) e^{-st} dt \\
 &= \int_0^\tau 0 \cdot e^{-st} dt + \int_\tau^\infty e^{-st} dt \\
 &= \int_\tau^\infty e^{-st} dt
 \end{aligned} \tag{154}$$

using a change of variable substitution,

$$t^* = t - \tau; \quad t = t^* + \tau; \quad dt^* = dt \tag{155}$$

Therefore

$$\begin{aligned}
 \int_\tau^\infty e^{-st} dt &= \int_0^\infty e^{-s(t^* + \tau)} dt^* \\
 &= e^{-s\tau} \int_0^\infty e^{-s(t + \tau)} dt^* \\
 &= e^{-s\tau} \frac{1}{s}
 \end{aligned} \tag{156}$$

connecting the start of proof to the end shows:

$$\mathcal{L}[1 \cdot (t - \tau)] = e^{-\tau s} \frac{1}{s} \tag{157}$$

### 3.3.5 Laplace transform pulse function

$$\text{LT pair} = \begin{cases} f(t) & p(t; \tau) \\ F(s) & \frac{1 - e^{-s\tau}}{s\tau} \end{cases} \tag{158}$$

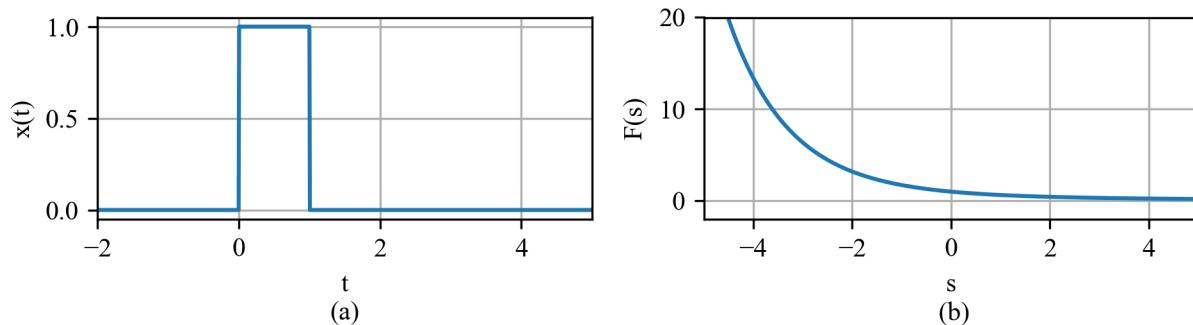


Figure 3.11: Pulse function; showing the (a) time domain and; (c) the s-space.

**Proof**

Show that

$$\mathcal{L}[p(t; \tau)] = \frac{1 - e^{-st\tau}}{s\tau} \quad (159)$$

This can be done by writing the pulse as a step up followed by a step down at  $t = \tau$  and scaled to be  $\frac{1}{\tau}$ , i.e.,:

$$p(t; \tau) = \frac{1}{\tau} 1(t) - \frac{1}{\tau} 1(t - \tau) \quad (160)$$

given that

$$\mathcal{L}[1(t)] = \frac{1}{s}, \quad \mathcal{L}[1(t - \tau)] = e^{-s\tau} \frac{1}{s} \quad (161)$$

the  $s$  domain of  $p(t; \tau)$  is expressed as

$$F(s) = \frac{1}{\tau} \frac{1}{s} - \frac{1}{\tau} e^{-s\tau} \frac{1}{s} \quad (162)$$

$$\begin{aligned} F(s) &= \frac{1}{\tau} \frac{1}{s} - \frac{1}{\tau} e^{-s\tau} \frac{1}{s} \\ &= \frac{1 - e^{-s\tau}}{s\tau} \end{aligned} \quad (163)$$

connecting the start of proof to the end shows:

$$\mathcal{L}[p(t; \tau)] = \frac{1 - e^{-s\tau}}{s\tau} \quad (164)$$

### 3.3.6 Laplace transform impulse function

$$\text{LT pair} = \begin{cases} f(t) & \delta(t) \\ F(s) & 1 \end{cases} \quad (165)$$

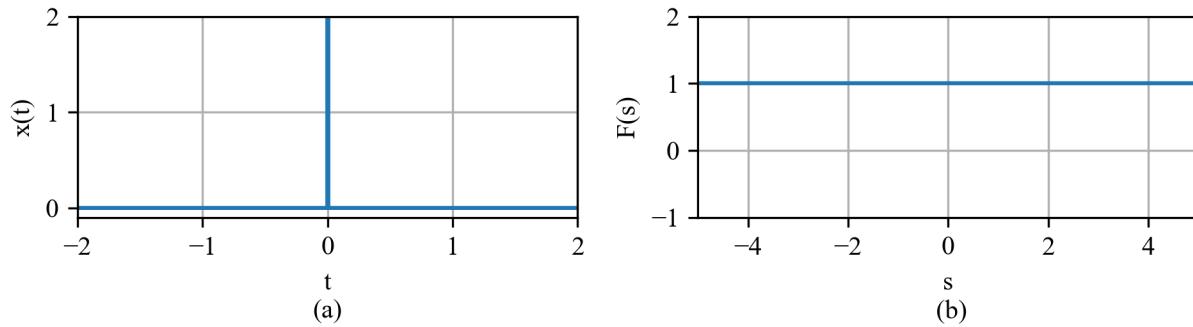


Figure 3.12: Impulse function; showing the (a) time domain and; (c) the s-space.

### Proof

Show that

$$\mathcal{L}[\delta(t)] = 1 \quad (166)$$

Consider  $\delta(t)$  as the limit of  $p(t; \tau)$  as  $\tau \rightarrow 0$  and take the Laplace Transform.

$$\delta(t) = \lim_{\tau \rightarrow 0} p(t; \tau) \quad (167)$$

taking the Laplace transform

$$\begin{aligned} \mathcal{L}[\delta(t)] &= \lim_{\tau \rightarrow 0} \mathcal{L}[p(t; \tau)] \\ &= \lim_{\tau \rightarrow 0} \frac{1 - e^{-st}}{s\tau} \end{aligned} \quad (168)$$

taking the limit of this in  $1 - 1/0 = 0/0$ . Applying L'Hospital's (loh-peey-TAHL) rule,

$$\begin{aligned} \lim_{\tau \rightarrow 0} \frac{\frac{\partial}{\partial \tau}(1 - e^{-st})}{\frac{\partial}{\partial \tau}(s\tau)} &= \lim_{\tau \rightarrow 0} \frac{-(-se^{-s\tau})}{s} \\ &= \lim_{\tau \rightarrow 0} e^{-s\tau} \\ &= 1 \end{aligned} \quad (169)$$

connecting the start of proof to the end shows:

$$\mathcal{L}[\delta(t)] = 1 \quad (170)$$

Moreover, this can also be done using the localization property of the delta function. Recall that

$$\int_0^\infty \delta(t)g(t)dt = g(0) \quad (171)$$

then

$$\begin{aligned}
 \mathcal{L}[\delta(t)] &= \int_0^\infty \delta(t)e^{-st}dt; \text{ where } g(t) = e^{-st} \\
 &= e^{-st} \Big|_{t=0} \\
 &= e^0 \\
 &= 1
 \end{aligned} \tag{172}$$

Again, connecting the start of proof to the end shows:

$$\mathcal{L}[\delta(t)] = 1 \tag{173}$$

### 3.4 Properties of Laplace Transforms

There are a variety of Laplace transforms that can assist in moving between  $f(t)$  and  $F(s)$ . Importantly, the differentiation, integration, time shift, and final value theorems are useful as they allow you to expand a Laplace table by taking the Laplace transform of functions without using the Laplace transform definition. Moreover, these properties are used extensively during control system design so being fluent in them makes you a better designer. A subset of these are shown in Table 1 while some of the more important ones are derived in the following sub sections.

The goal of studying the Laplace transform properties is to be able to Laplace Transform a differential equation with initial conditions and to relate differentiation and integration in the time domain to their counterparts in the s-domain domain.

Table 1: A list of important Laplace Transform theorems used in control.

number	theorem	name
1	$\mathcal{L}[f(t)] = F(s) = \int_0^\infty f(t)e^{-st}dt$	definition
2	$\mathcal{L}[kf(t)] = kF(s)$	linearity theorem
3	$\mathcal{L}[f_1(t) + f_2(t)] = F_1(s) + F_2(s)$	linearity theorem
4	$\mathcal{L}[e^{-at}f(t)] = F(s+a)$	frequency shift theorem
5	$\mathcal{L}[f(t-t_0)] = e^{-t_0s}F(s)$	time shift theorem
6	$\mathcal{L}[e^{s_0t}f(t)] = F(s-s_0)$	time shift theorem
7	$\mathcal{L}[f(at)] = \frac{1}{a}F\left(\frac{s}{a}\right)$	scaling theorem
8	$\mathcal{L}[f'(t)] = sF(s)$ , if $f(0) = 0$	differentiation theorem
9	$\mathcal{L}[f''(t)] = s^2F(s)$ , if $f'(0) = f(0) = 0$	differentiation theorem
10	$\mathcal{L}[f^n(t)] = s^nF(s)$ , if $f^{n-1}(0) = \dots = f'(0) = f(0) = 0$	differentiation theorem
11	$\mathcal{L}\left[\int_0^t f(t^*)dt^*\right] = \frac{1}{s}F(s)$	integration theorem
12	$f(\infty) = \lim_{s \rightarrow 0} [sF(s)]$	final value theorem <sup>1</sup>
13	$f(0+) = \lim_{s \rightarrow \infty} [sF(s)]$	initial value theorem <sup>2</sup>

<sup>1</sup>This requires all roots of the denominator of  $F(s)$  to have negative real parts, and no more than one can be on the origin of the real axis.

<sup>2</sup>This requires that  $f(t)$  be continuous or have a step discontinuity at  $t = 0$ . This does not allow for impulses at  $t = 0$ .

### 3.4.1 Differentiation Property

The Laplace transforms can be simplified if we assume  $f(0) = 0$ , where these are the terms associated with the initial conditions. This is shown as:

$$\begin{aligned}\mathcal{L}[f'(t)] &= sF(s), \text{ if } f(0) = 0 \\ \mathcal{L}[f''(t)] &= s^2F(s), \text{ if } f'(0) = f(0) = 0 \\ &\vdots \\ \mathcal{L}[f^{(n)}(t)] &= s^{(n)}F(s), \text{ if } f^{(n-1)}(0) = \dots = f'(0) = f(0) = 0\end{aligned}\quad (174)$$

Recall that differentiation in the time domain becomes multiplication by  $s$  in the  $s$ -domain.

#### Proof

**Method 1:** Show that  $\mathcal{L}[f'(t)] = sF(s)$  if  $f(0) = 0$ ;

$$\mathcal{L}[f'(t)] = \int_0^\infty f'(t)e^{-st}dt \quad (175)$$

integration by parts  $udv = d(uv) - vdu$  where

$$\begin{aligned}e^{-st} &= u \\ f'(t)dt &= dv \\ f(t) &= v \\ -se^{-st} &= du\end{aligned}\quad (176)$$

leads to

$$\mathcal{L}[f'(t)] = f(t)e^{-st}|_0^\infty - \int_0^\infty f(t)(-se^{-st})dt \quad (177)$$

further simplification results in:

$$\begin{aligned}\mathcal{L}[f'(t)] &= f(\infty)e^{-\infty} - f(0) + s \int_0^\infty fe^{-st}dt \\ &= sF(s) - f(0) \\ &= sF(s)\end{aligned}\quad (178)$$

where  $F(s) = \int_0^\infty fe^{-st}dt$ ,  $e^{-\infty} = 0$ , and  $f(0) = 0$ . Therefore,

$$\mathcal{L}[f'(t)] = sF(s) \quad (179)$$

**Method 2:** Show that  $\mathcal{L}[f'(t)] = sF(s)$  if  $f(0) = 0$ . Denote  $f(t) = f'(t)$ , therefore:

$$\begin{aligned} G(s) &= \mathcal{L}[g(t)] = \mathcal{L}[f'(t)] = sF(s) \\ \mathcal{L}[f''(t)] &= \mathcal{L}[g'(t)] = s\mathcal{L}[G(s)] = s^2F(s) \end{aligned} \quad (180)$$

by induction it can be shown that

$$\mathcal{L}[f'(t)] = sF(s) \quad (181)$$

Bottom line: “to differentiate, multiply by  $s$ ”, provided  $f(0) = 0$ ,  $f'(0) = 0$ , etc. Else, need to subtract them  $\mathcal{L}[f'(t)] = sF(s) - f(0)$  etc.

### 3.4.2 Integration Property

The integral of  $f(t)$  is

$$\mathcal{L}\left[\int_0^t f(t^*)dt^*\right] = \frac{1}{s}F(s) \quad (182)$$

where  $t^*$  is the dummy variable to integrate over as  $t$  is used for the limit. The main take away here is that integration in the time domain becomes division by  $s$  in the  $s$ -domain.

#### Proof

Show that  $\mathcal{L}\left[\int_0^t f(t^*)dt^*\right] = \frac{1}{s}F(s)$ . Denote  $g(t) = \int_0^t f(t^*)dt^*$ , then:

$$g'(t) = f(t); g(0) = 0 \quad (183)$$

taking the Laplace of the left hand side of this expression yields:

$$\mathcal{L}[g'(t)] = s\mathcal{L}[g(t)] = s\mathcal{L}\left[\int_0^t f(t^*)dt^*\right] \quad (184)$$

the Laplace of the right hand side of this expression yields:

$$\mathcal{L}[f(t)] = F(s) \quad (185)$$

therefore, by equation 183

$$s\mathcal{L}\left[\int_0^t f(t^*)dt^*\right] = F(s) \quad (186)$$

divide by  $s$  to get

$$\mathcal{L}\left[\int_0^t f(t^*)dt^*\right] = \frac{1}{s}F(s) \quad (187)$$

Bottom line: “to integrate, divide by  $s$ ”.

### 3.4.3 Time Shift Property

To shift in  $t$

$$\mathcal{L}[f(t - t_0)] = e^{-t_0 s}F(s) \quad (188)$$

To shift in  $s$

$$\mathcal{L}[e^{s_0 t} f(t)] = F(s - s_0) \quad (189)$$

A shift in the  $t$  domain

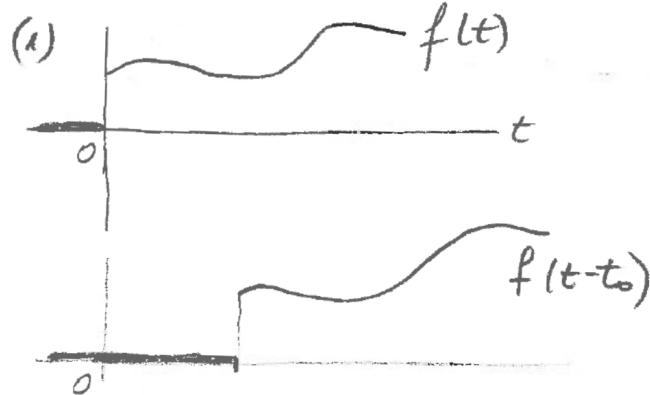


Figure 3.13: Shift of a signal in the  $t$  domain.

Note that:

- Function  $f(t)$  is zero for  $t < 0$  arrangement
- Function  $f(t - t_0)$  is zero for  $t < t_0$  arrangement

### Proof

#### t-domain

For a shift in  $t$ , show that  $\mathcal{L}[f(t - t_0)] = e^{-t_0 s} F(s)$

$$\begin{aligned} \mathcal{L}[f(t - t_0)] &= \int_0^\infty f(t - t_0) e^{-st} dt \\ &= \int_0^{t_0} 0 \cdot e^{-st} dt + \int_{t_0}^\infty f(t - t_0) e^{-st} dt \\ &= \int_{t_0}^\infty f(t - t_0) e^{-st} dt \end{aligned} \quad (190)$$

using a change of variables:

$$\begin{aligned} t^* &= t - t_0 \rightarrow t = t^* + t_0 \\ dt^* &= dt \end{aligned} \quad (191)$$

therefore,

$$\begin{aligned} \mathcal{L}[f(t - t_0)] &= f(t^*) e^{-s(t^* + t_0)} dt^* \\ &= e^{-st_0} \int_0^\infty f(t^*) e^{-st^*} dt^* \\ &= e^{-st_0} F(s) \end{aligned} \quad (192)$$

lastly, we can show that:

$$\mathcal{L}[f(t - t_0)] = e^{-t_0 s} F(s) \quad (193)$$

### s-domain

For a shift in  $s$ , show that  $\mathcal{L}[e^{s_0 t} f(t)] = F(s - s_0)$

$$\begin{aligned} \mathcal{L}[e^{s_0 t} f(t)] &= \int_0^\infty e^{-s_0 t} f(t) e^{-st} dt \\ &= \int_0^\infty f(t) e^{-(s-s_0)t} dt \\ &= F(s - s_0) \end{aligned} \quad (194)$$

### 3.4.4 Final Value Theorem

This applies to the steady state response  $x(\infty)$  which can also be denoted with a subscript  $ss$  as  $x_{ss}$  where

$$x_{ss} = x(\infty) = \lim_{t \rightarrow \infty} [x(t)] = \lim_{s \rightarrow 0} [sX(s)] \quad (195)$$

therefore,

$$x_{ss} = \lim_{s \rightarrow 0} [sX(s)] \quad (196)$$

### Proof

Prove that  $x_{ss} = \lim_{s \rightarrow 0} sX(s)$ . Starting with,

$$\mathcal{L}[\dot{x}(t)] = sX(s), x(0) = 0 \quad (197)$$

by definition

$$\mathcal{L}[\dot{x}(t)] = \int_0^\infty \dot{x}(t) e^{-st} dt \quad (198)$$

combing the last two equations yields

$$\int_0^\infty \dot{x}(t) e^{-st} dt = sX(s) \quad (199)$$

taking the limit as  $s \rightarrow 0$  results in

$$\lim_{s \rightarrow 0} \left[ \int_0^\infty \dot{x}(t) e^{-st} dt \right] = \lim_{s \rightarrow 0} [sX(s)] \quad (200)$$

but

$$\lim_{s \rightarrow 0} [e^{-st}] = e^0 = 1 \quad (201)$$

therefore, the left hand side of equation 200 simplifies to

$$\int_0^\infty \dot{x}(t)dt = x(t)|_0^\infty = x(\infty) = x_{ss} \quad (202)$$

inserting this term back into equation 200 yields the desired proof

$$x_{ss} = \lim_{s \rightarrow 0} [sX(s)] \quad (203)$$

## 3.5 Convolutional property of Laplace Transforms

$$\begin{aligned} \mathcal{L}[G(s) \cdot F(s)]^{-1} &= \mathcal{L}[F(s) \cdot G(s)]^{-1} \\ &= \int_0^t f(\tau)g(t-\tau)d\tau \\ &= (f * g)(t) \\ &= x(t) \end{aligned} \quad (204)$$

where  $f(\tau)$  is the excitation function and  $g(t - \tau)$  is the impulse response of the function. Convolution expresses the system response  $x(t)$  to a complicated excitation  $f(\tau)$  as an integral using the impulse response  $g(t)$  shifted by  $\tau$  to  $t - \tau$ . The integral in equation 204 is not easily computed in the time domain, however, Laplace transforms make it easy using a three step process:

1. Calculate  $F(s) = \mathcal{L}[f(t)]$ ,  $G(s) = \mathcal{L}[g(t)]$
2. Multiply  $F(s)G(s)$
3. Take the inverse Laplace transform to get  $x(t)$

## 3.6 Transfer Function for Response to Random Inputs

Consider the following system

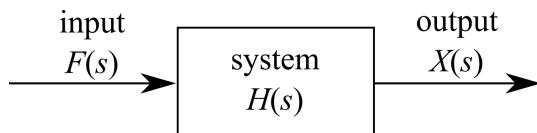


Figure 3.14: Generic block diagram of a system  $H(s)$  subjected to an input  $F(s)$  and its corresponding output  $X(s)$  where the  $(s)$  denotes that the considered system is in the  $s$ -plane.

where  $F(s)$  is the input,  $H(s)$  is the system, and  $X(s)$  is the output from the system. This formulation is called the transfer-function approach and is commonly used for the formulation and solution of dynamic problems in the control literature. It can also be used for solving the various forced-vibration problems including those from complex or stochastic inputs.

### 3.6.1 Defining the transfer function $H(s)$

Again, consider the generic system represented in figure 3.14. For this system representation,  $F(s)$  is the Laplace of the transform of the driving force and  $H(s)$  is the Laplace transform of the response of the system  $h(t)$ .

We need to define transfer function  $H(s)$  for a generic system. To do this let us show the reasoning behind the transfer function. Here we will show that the output of any system ( $x(t)$ ) can be related to the input of the system ( $f(t)$ ) through a series of polynomial coefficients ( $a$  and  $b$ ). Consider the general  $n^{th}$ -order linear, time-invariant differential equation that governs the behavior of the dynamic system.

$$a_n \frac{d^n x(t)}{dt^n} + a_{n-1} \frac{d^{n-1} x(t)}{dt^{n-1}} + \dots + a_0 x(t) = b_m \frac{d^m f(t)}{dt^m} + b_{m-1} \frac{d^{m-1} f(t)}{dt^{m-1}} + \dots + b_0 f(t) \quad (205)$$

where  $x(t)$  is the output and  $f(t)$  is the input. Note that this is similar to the formulation we have had before for the EOM. Taking the Laplace transformation of both side of the above equation yields

$$\begin{aligned} a_n s^n X(s) + a_{n-1} s^{n-1} X(s) + \dots + a_0 X(s) &+ \text{initial condition for } x(t) = \\ b_m s^m F(s) + b_{m-1} s^{m-1} F(s) + \dots + b_0 F(s) &+ \text{initial condition for } f(t) \end{aligned} \quad (206)$$

It can be seen that this equation is a purely algebraic expression. If we assume the initial conditions to be zero, the equation reduces to the following:

$$(a_n s^n + a_{n-1} s^{n-1} + \dots + a_0) X(s) = (b_m s^m + b_{m-1} s^{m-1} + \dots + b_0) F(s) \quad (207)$$

if we rearrange equation 207 to solve for the relationship between the Laplace variables ( $X(s)$  and  $F(s)$ ) and the algebraic expressions we get:

$$\frac{X(s)}{F(s)} = \frac{b_m s^m + b_{m-1} s^{m-1} + \dots + b_0}{a_n s^n + a_{n-1} s^{n-1} + \dots + a_0} \quad (208)$$

this shows that the ratio of the input algebraic expressions over the output algebraic expressions is equal to the ratio of the output Laplace variable over the input Laplace variable. This shows that we can relate the Laplace variables to the algebraic expressions. Therefore, we can define the transfer function  $H(s)$  as:

$$G(s) = H(s) = \frac{X(s)}{F(s)} \quad (209)$$

In a more formal term, the transfer function that is defined as: “The ratio of the Laplace transforms of the output or response function to the Laplace transform of the input or forcing function assuming zero initial conditions”. Note that many texts define  $H(s)$  as  $G(s)$  and this is simply a matter of syntax.

Equation 209 can be rearranged to show that the output of the system  $X(s)$ , can be obtained if we know the input  $F(s)$  and the transfer function  $H(s)$ :

$$X(s) = H(s)F(s) \quad (210)$$

### 3.6.2 Transfer Function method (Steady-State solution)

Considering the forced system:

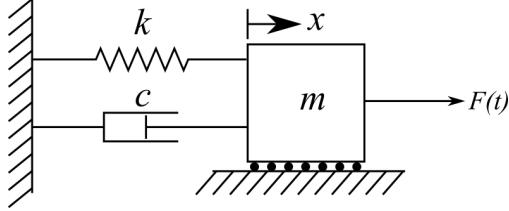


Figure 3.15: A spring-dashpot-mass model of a 1-DOF system with external excitation.

that can be expressed as the equation of motion

$$m\ddot{x}(t) + c\dot{x}(t) + kx(t) = F_0 \cos(\omega t) \quad (211)$$

Here  $F_0 \cos(\omega t)$ , is used at the input but any input will develop the same transfer function as the transfer function is bounded to the system and not the input. From the #6 in the table for Laplace Transforms, we know that

$$\mathcal{L}[\cos(\omega t)] = \frac{s}{s^2 + \omega^2} \quad (212)$$

Therefore,

$$F(s) = \frac{F_0 s}{s^2 + \omega^2} \quad (213)$$

Ignoring the initial conditions, and therefore considering only the particular solution, and taking the Laplace transform of the EOM equation yields:

$$(ms^2 + cs + k)X(s) = \frac{F_0 s}{s^2 + \omega^2} \quad (214)$$

where  $X(s)$  denotes the Laplace transform of the unknown function  $x(t)$  and  $s$  is the complex transform variable. Rearranging the above equation for  $X(s)$  yields:

$$X(s) = \frac{F_0 s}{(ms^2 + cs + k)(s^2 + \omega^2)} \quad (215)$$

Now that we have  $F(s)$  and  $X(s)$  we can obtain  $H(s)$  as

$$H(s) = \frac{X(s)}{F(s)} = \frac{F_0 s}{(ms^2 + cs + k)(s^2 + \omega^2)} \cdot \frac{s^2 + \omega^2}{F_0 s} = \frac{1}{ms^2 + cs + k} \quad (216)$$

or

$$H(s) = \frac{1}{ms^2 + cs + k} \quad (217)$$

This ratio is termed the transfer function of a system and is an important tool in vibration analysis.

Sometimes, how the system responds to inputs with certain frequency components is important in understanding the system in general, therefore, we want to solve for the frequency response function of the system. The frequency response function is denoted as  $H(j\omega)$  where the

complex number  $s$  is replaced by the frequency component of the system while considering the imaginary portion in the complex plane (i.e.,  $s = j\omega$ ). Therefor, the frequency response function of the system becomes:

$$H(j\omega) = \frac{1}{m(j\omega)^2 + cj\omega + k} = \frac{1}{-m\omega^2 + cj\omega + k} \quad (218)$$

rearranging into a standard form yields:

$$H(j\omega) = \frac{1}{k - m\omega^2 + c\omega j} \quad (219)$$

recall that  $j^2 = -1$ . This is the frequency response function of the system. Therefore, it can be seen that the frequency response function of the system is the transfer function of the system evaluated along the imaginary axis  $s = j\omega$ . However, this expression contains imaginary values (that help to account for the phase in the system) and therefore can be challenging to work with. As the amplitude  $|H(j\omega)|$  of the response (the real portion of the equation) is useful to the practitioner, it is prudent to consider the special case of amplitude response while neglecting the phase response. Consider that:

$$H(j\omega) = R + Ij \quad (220)$$

so

$$|H(j\omega)| = \sqrt{R^2 + I^2} \quad (221)$$

multiplying  $H(j\omega)$  by 1 that is represented by its unit complex conjugate yields:

$$\begin{aligned} H(j\omega) &= \left( \frac{1}{k - m\omega^2 + c\omega j} \right) \left( \frac{k - m\omega^2 - c\omega j}{k - m\omega^2 - c\omega j} \right) \\ &= \left( \frac{k - m\omega^2}{(k - m\omega^2)^2(c\omega)^2} \right) \left( \frac{-c\omega}{(k - m\omega^2)^2(c\omega)^2} j \right) \end{aligned} \quad (222)$$

therefore,  $R = \frac{k - m\omega^2}{(k - m\omega^2)^2(c\omega)^2}$  and  $I = \frac{-c\omega}{(k - m\omega^2)^2(c\omega)^2}$ . Now, calculating the amplitude of  $H(j\omega)$  we get:

$$\begin{aligned} H(\omega) &= |H(j\omega)| \\ &= \sqrt{R^2 + I^2} \\ &= \sqrt{\frac{(k - m\omega^2)^2 + (-c\omega)^2}{((k - m\omega^2)^2 + (c\omega)^2)^2}} \\ &= \sqrt{\frac{1}{(k - m\omega^2)^2 + c^2\omega^2}} \\ &= \frac{1}{\sqrt{(k - m\omega^2)^2 + c^2\omega^2}} \end{aligned} \quad (223)$$

where  $H(\omega)$  represents only the amplitude of the frequency response function and therefore drops the  $j$  term from the expression.

To recap, for a single DOF damped spring-mass system the transfer function is:

$$H(s) = \frac{1}{ms^2 + cs + k} \quad (224)$$

And the frequency response function is:

$$H(j\omega) = \frac{1}{k - m\omega^2 + c\omega j} \quad (225)$$

While the amplitude of the frequency response is:

$$H(\omega) = |H(j\omega)| = \frac{1}{\sqrt{(k - m\omega^2)^2 + c^2\omega^2}} \quad (226)$$

**Example 3.2** Considering the forced system:

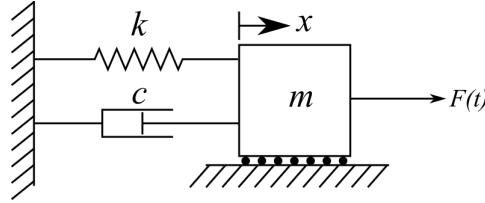


Figure 3.16: A spring-dashpot-mass model of a 1-DOF system with external excitation.

Set the forcing function to be  $F_0 \sin(\omega t)$  and calculate the transfer function.

**Solution:** The equation of motion for the system is:

$$m\ddot{x} + c\dot{x} + kx = F_0 \sin(\omega t) \quad (227)$$

From the #6 in the table for Laplace Transforms, we know that:

$$\mathcal{L}[\sin(\omega t)] = \frac{\omega}{s^2 + \omega^2} \quad (228)$$

Therefore,

$$F(s) = \frac{F_0 \omega}{s^2 + \omega^2} \quad (229)$$

Ignoring the initial conditions and taking the Laplace transform of the EOM equation yields:

$$(ms^2 + cs + k)X(s) = \frac{F_0 \omega}{s^2 + \omega^2} \quad (230)$$

Solving algebraically for the  $X(s)$  yields:

$$X(s) = \frac{F_0 \omega}{(ms^2 + cs + k)(s^2 + \omega^2)} \quad (231)$$

Now that we have  $F(s)$  and  $X(s)$  we can obtain  $H(s)$  as

$$H(s) = \frac{X(s)}{F(s)} = \frac{F_0\omega}{(ms^2 + cs + k)(s^2 + \omega^2)} \cdot \frac{s^2 + \omega^2}{F_0\omega} = \frac{1}{ms^2 + cs + k} \quad (232)$$

or

$$H(s) = \frac{1}{ms^2 + cs + k} \quad (233)$$

This is identical to the solution obtained using  $F_0 \cos(\omega t)$  as would be expected because the transfer function is related to the system and not to the input.

### Review 3.2 Frequency and Time Domains

The frequency domain is a mathematical representation of a signal or data in terms of its frequency components, as opposed to its temporal or time-based representation. The frequency domain provides a different perspective on the signal by decomposing it into its constituent sinusoidal signals at discrete frequencies and their respective magnitudes. A 3D representation of this process is shown in figure 3.17. The transformation between the time domain and the frequency domain is typically achieved using mathematical techniques such as the Fourier Transform or the Fast Fourier Transform (FFT).

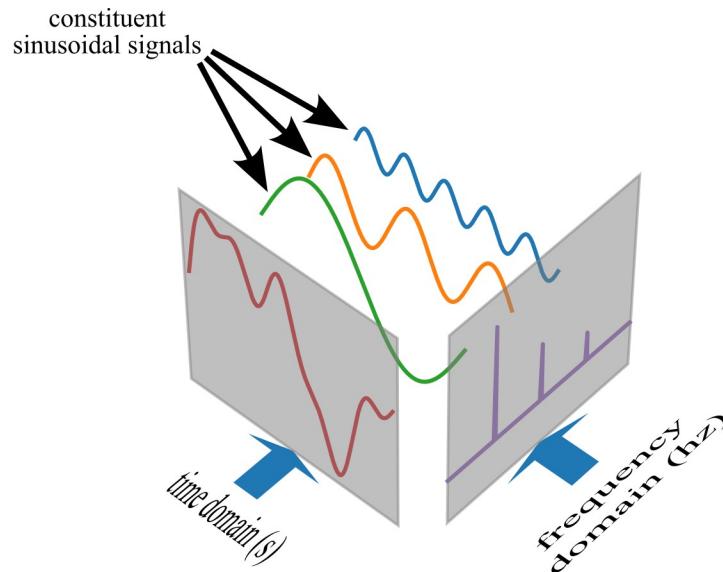


Figure 3.17: 3D visualization of time and frequency domains where a temporal signal is decomposed into constituent sinusoidal signals.

### 3.6.3 Response to Random Inputs

The transfer and frequency response functions can be very useful for determining the system's response to random inputs. Up to this point we have solved for deterministic input.

- **Deterministic**-For a known time  $t$ , the value of the input force  $F(t)$  is precisely known.

- **Random** For a known time  $t$ , the value of the input force  $F(t)$  is known only statistically.

To expand, a random signal is a signal with no obvious pattern. For these types of it is not possible to focus on the details of the input signal, as is done with a deterministic signal, rather the signal is classified and manipulated in terms of its statistical properties.

Randomness in vibration analysis can be thought of as the result of a series of results obtained from testing a system repeatability for various inputs under varying conditions. In these cases, one record or time history is not enough to describe the system. Rather, an ensemble of various tests are used to describe how the system will respond to the various inputs.

First, let us consider two inputs, a deterministic input (typical sin wave), and a random input (white noise). These inputs are shown in figure 3.18.

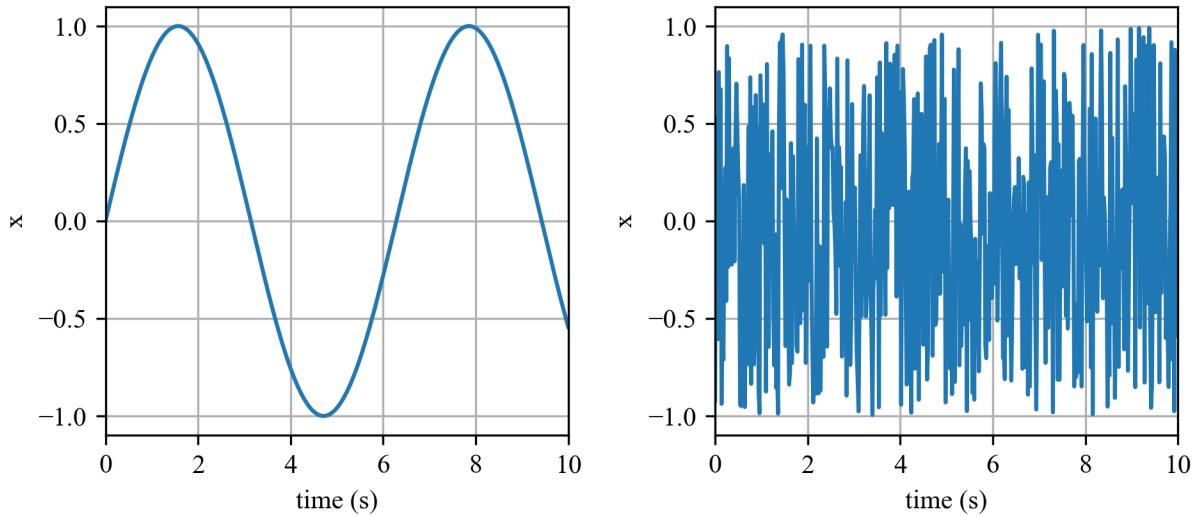


Figure 3.18: Two arbitrary inputs: (a) sinusoidal; and (b) uniform random noise.

One of the first factors to consider is the mean of the random signal  $x(t)$ , defined as:

$$E[x] = \bar{x} = \lim_{T \rightarrow \infty} \frac{1}{T} \int_0^T x(t) dt \quad (234)$$

where  $T$  is the length in time of the data collected. However, for random signals we often want to consider signals with an average mean of zero (i.e.  $\bar{x}(t) = 0$ ). Therefore, for signals not centered around zero we can obtain a zero centered signal if the signal is stationary and we subtract the mean value from  $\bar{x}$  from the signal  $x(t)$ . This can be written as:

$$x'(t) = x(t) - \bar{x} \quad (235)$$

where the  $x'(t)$  is now centered around zero. As mentioned before, it is important to consider whether or not the input signals are stationary. A signal is stationary if its statistical properties (usually expressed by its mean) do not change with time. Here, it can be seen that for our inputs considered the signals are stationary if a long enough time period is considered.

Another important variable is variance (or mean-square value) of the random variable  $x(t)$  defined as:

$$E[(x - \bar{x})^2] = \lim_{T \rightarrow \infty} \frac{1}{T} \int_0^T (x(t) - \bar{x})^2 dt \quad (236)$$

and provides a measure of the magnitude of the fluctuations in the signal  $x(t)$ . If the signal has an expected value of zero, or  $E[x] = 0$ , this simplifies to.

$$E[x^2] = \bar{x}^2 = \lim_{T \rightarrow \infty} \frac{1}{T} \int_0^T x^2(t) dt \quad (237)$$

This expression leads to the calculation of the root-mean-square (RMS) of the signal:

$$x_{\text{rms}} = \sqrt{\bar{x}^2} \quad (238)$$

Considering a nonstationary signal, an important measure of interests is how fast the value of the variables change. This is important to understand as it provides context for how long a signal must be sampled to before a meaningful representation of the signal can be calculated in a statistical sense. One way to quantify how fast the values of signal change is the autocorrelation function:

$$R_{xx}(\tau) = \lim_{T \rightarrow \infty} \frac{1}{T} \int_0^T x(t)x(t + \tau) dt \quad (239)$$

The subscript  $xx$  denotes that this is a measure of the response for the variable  $xx$ ,  $\tau$  is the time difference between the values at which the signal  $x(t)$  is sampled. The auto correlation for the two inputs considered above are expressed in figure 3.19.

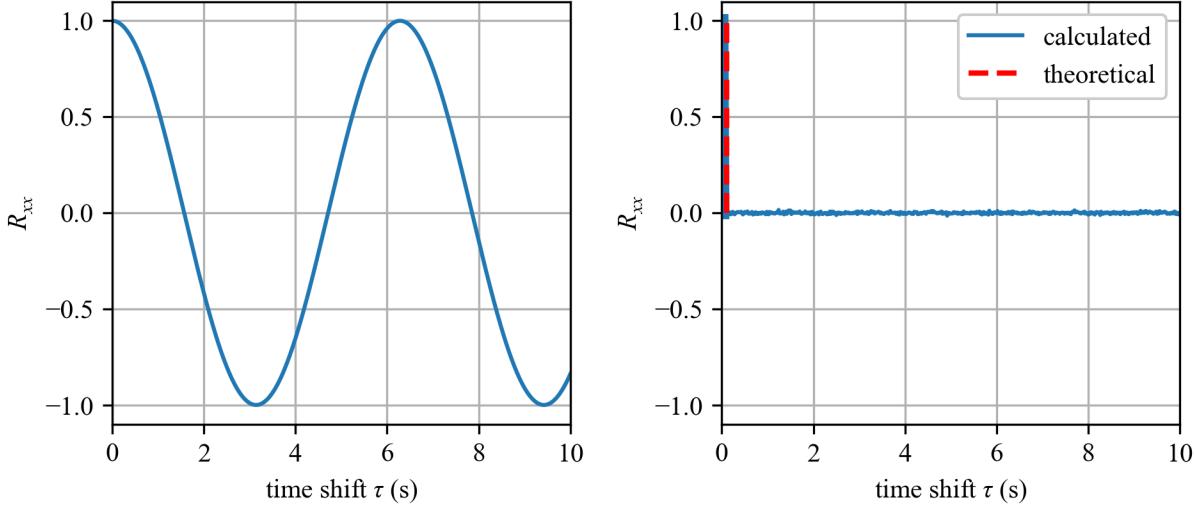


Figure 3.19: Responses from the autocorrelation function for the inputs shown in figure 3.18 showing: (a) a sinusoidal; and (b) uniform random noise.

Note that the value of  $\tau$  selected in the auto correlation function greatly affects its response for the sinusoidal input. This is because the values for the sinusoidal are highly correlated. To expand, the

value at any time  $t$  is greatly effected by the values immediately before and after it. This is not the case for the random input where the signal is not correlated and therefore there is little difference in changing the value of  $\tau$  on the response of the autocorrelation function.

Next, if we take the Fourier transform of the autocorrelation function we obtain the power spectral density (PSD) defined as:

$$S_{xx}(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} R_{xx}(\tau) e^{-j\omega\tau} d\tau \quad (240)$$

where the integral of  $R_{xx}(\tau)$  changes the real number  $\tau$  into the frequency-domain value  $\omega$ . The frequency spectrum is denoted with  $S$  and the subscript of the considered variable (e.g.,  $S_{xx}(\omega)$ ). The frequency spectrum for the two input cases considered are plotted in figure 3.20.

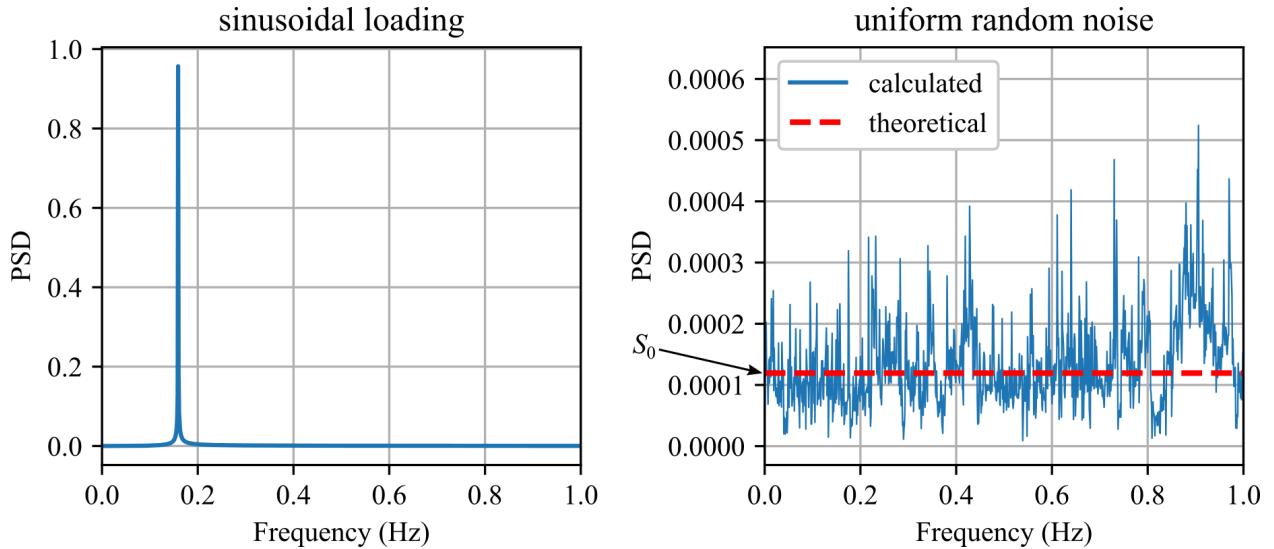


Figure 3.20: Power spectral density plots for the inputs shown in figure 3.18 showing: (a) a sinusoidal; and (b) uniform random noise.

where the flat frequency response for the random input denotes that the random input is white noise input. This flat frequency response in the frequency domain can be denoted  $S_0$ , such that  $S_{ff}(\omega) = S_0$  or  $S_{xx}(\omega) = S_0$ , depending on whether the frequency spectrum of the input ( $ff$ ) or output ( $xx$ ) is being considered. While a true white noise input would be perfectly flat, white noise is really just a theoretical concept as all real-world data will have some variation in the frequency domain as diagrammed in figure 3.20(b).

Recall that  $S_{xx}$  is the spectrum of the response of the system. For the one-DOF system considered here, we can express the arbitrary input as a series of impulse inputs as shown in section 3.2.3. This knowledge, along with the frequency response function can be used to relate the spectrum of the input  $S_{ff}(\omega)$  to the output through the transfer function as:

$$S_{xx}(\omega) = |H(j\omega)|^2 \left[ \frac{1}{2\pi} \int_{-\infty}^{\infty} R_{ff}(\tau) e^{-j\omega\tau} d\tau \right] \quad (241)$$

This can also be expressed in symbolic form as:

$$S_{xx}(\omega) = |H(j\omega)|^2 S_{ff}(\omega) \quad (242)$$

where  $R_{ff}$  denotes the autocorrelation function of  $F(t)$  and  $S_{ff}$  denotes the PSD of the forcing function  $F(t)$ . The notation  $|H(j\omega)|^2$  is the square of the magnitude of the complex frequency response function. A more detailed derivation can be found in [Engineering Vibrations, Inman (2001)], [Random Vibrations, Spectral & Wavelet Analysis, Newland (1993)], but here it is more important to study the results rather than the derivations.

**Example 3.3** Consider the following system

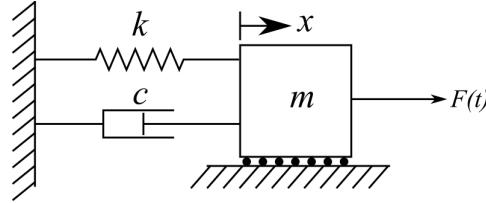


Figure 3.21: A spring-dashpot-mass model of a 1-DOF system with external excitation.

Calculate the PSD of the response  $x(t)$  given that the PSD of the applied force  $S_{ff}(\omega)$  is white noise.

**Solution:** From the system we know that the EOM is

$$m\ddot{x}(t) + c\dot{x}(t) + kx(t) = F(t) \quad (243)$$

The frequency response function for this system is

$$H(j\omega) = \frac{1}{k - m\omega^2 + c\omega j} \quad (244)$$

while the amplitude of the response is:

$$H(\omega) = |H(j\omega)| = \frac{1}{\sqrt{(k - m\omega^2)^2 + c^2\omega^2}} \quad (245)$$

Applying the equation that relates  $S_{ff}(\omega)$  to  $S_{xx}(\omega)$  we get:

$$S_{xx}(\omega) = |H(j\omega)|^2 S_{ff}(\omega) = \left| \frac{1}{k - m\omega^2 + c\omega j} \right|^2 S_{ff}(\omega) \quad (246)$$

White noise means the forcing function  $S_{ff}(\omega)$  is constant across the frequency spectrum, therefore,  $S_{ff}(\omega) = S_0$ . Additionally as:

$$|H(j\omega)|^2 = \left| \frac{1}{k - m\omega^2 + c\omega j} \right|^2 = \frac{1}{(k - m\omega^2)^2 + c^2\omega^2} \quad (247)$$

where the absolute value is the amplitude of the system. Therefore, we obtain:

$$S_{xx}(\omega) = |H(j\omega)|^2 S_0 = \frac{1}{(k - m\omega^2)^2 + c^2\omega^2} S_0 = \frac{S_0}{(k - m\omega^2)^2 + c^2\omega^2} \quad (248)$$

Using various values for the elements in the system, the PSD for the system considered looks like:

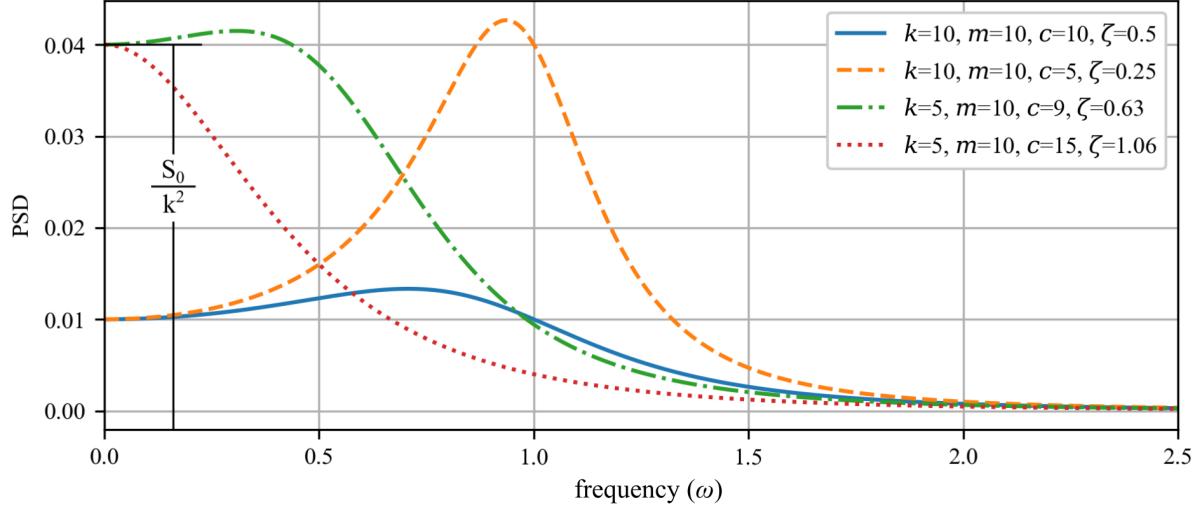


Figure 3.22: Response for considered 1-DOF systems subjected to a white noise input.

Another useful quantity to consider is the expected output, in terms of its mean and variance, for a given input. Working within the constraint that the system will oscillate about zero,  $E[x] = 0$ , the mean-square value can be directly related to the PSD function as:

$$E[x^2] = \bar{x^2} = \int_{-\infty}^{\infty} |H(j\omega)|^2 S_{ff}(\omega) d\omega \quad (249)$$

For a constant input  $S_0$ , as diagrammed in figure 3.20(b), the mean-square value can be expressed as:

$$E[x^2] = \bar{x^2} = S_0 \int_{-\infty}^{\infty} |H(j\omega)|^2 d\omega \quad (250)$$

After inspecting the above equation, it becomes clear that to obtain the square of the expected value, a solution for  $\int_{-\infty}^{\infty} |H(j\omega)|^2 d\omega$  must be obtained. For cases where  $S_{ff}(\omega) = S_0$  and as such  $S_{ff}(\omega)$  can be pulled out of the integral, these integrals have been solved [Random Vibrations, Spectral & Wavelet Analysis, Newland (1993)]. For example, given  $\int_{-\infty}^{\infty} |H(j\omega)|^2 d\omega$ :

$$\int_{-\infty}^{\infty} \left| \frac{B_0}{A_0 + j\omega A_1} \right|^2 d\omega = \frac{\pi B_0^2}{A_0 A_1} \quad (251)$$

and

$$\int_{-\infty}^{\infty} \left| \frac{B_0 + j\omega B_1}{A_0 + j\omega A_1 - \omega^2 A_2} \right|^2 d\omega = \frac{\pi (A_0 B_1^2 + A_2 B_0^2)}{A_0 A_1 A_2} \quad (252)$$

When combined with equation 250, these integrals allow for the easy calculation of the expected values.

**Example 3.4** For system below, calculate the mean-square response of the system given that the spectrum of the input force  $F(t)$  is a perfect theoretical white noise.

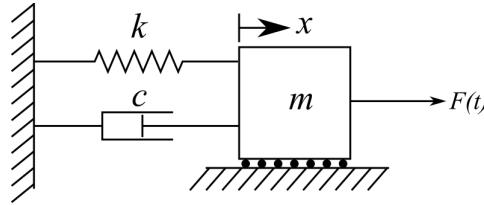


Figure 3.23: A spring-dashpot-mass model of a 1-DOF system with external excitation.

**Solution:** Again, as the forcing function  $S_{ff}(\omega)$  is constant across the frequency spectrum  $S_{ff}(\omega) = S_0$  the mean-square response can be calculated as:

$$E[x^2] = \bar{x^2} = S_0 \int_{-\infty}^{\infty} |H(j\omega)|^2 d\omega \quad (253)$$

Using the already tabulated response:

$$\int_{-\infty}^{\infty} \left| \frac{B_0 + j\omega B_1}{A_0 + j\omega A_1 - \omega^2 A_2} \right|^2 d\omega = \frac{\pi(A_0 B_1^2 + A_2 B_0^2)}{A_0 A_1 A_2} \quad (254)$$

and the frequency response function for the system as derived in equation 219:

$$H(j\omega) = \frac{1}{k - m\omega^2 + c\omega j} \quad (255)$$

when  $B_0 = 1$ ,  $B_1 = 0$ ,  $A_0 = k$ ,  $A_1 = c$ , and  $A_2 = m$ . Therefore, using the tabulated expression we can show that:

$$E[x^2] = S_0 \frac{\pi m}{kcm} = \frac{S_0 \pi}{kc} \quad (256)$$

### 3.7 Inverse Laplace Transform

Consider that  $X(s)$  can be expanded in partial fractions as

$$X(s) = \frac{B(s)}{A(s)} = \frac{r_1}{s - p_1} + \frac{r_2}{s - p_2} + \cdots + \frac{r_n}{s - p_n} = \sum_{i=1}^n \frac{r_i}{s - p_i} \quad (257)$$

recall that  $\mathcal{L}[e^{p_0 t}] = \frac{1}{s - p_0}$  for  $p_0$ . The inverse Laplace Transform of  $X(s)$  is a number:

$$x(t) = r_1 e^{p_1 t} + r_2 e^{p_2 t} + \cdots + r_n e^{p_n t} = \sum_{i=1}^n r_i e^{p_i t} \quad (258)$$

where the values  $p_1, p_2, \dots, p_n$  are the roots of the denominator equation and are the poles of the system and may be complex. The values  $r_1, r_2, \dots, r_n$  are called “residuals” and are defined

as,

$$r_i = \lim_{s \rightarrow p_i} [(s - p_i)X(s)] \quad (259)$$

the MATLAB function `residue` will perform partial fraction expansion (PFD) and can be used to find the values.

When complex poles appear in the system, they are conjugate pairs, i.e.,

$$p_{1,2} = \sigma \pm i\omega_d \quad (260)$$

where  $\sigma$  is the position of the poles on the real axis in the S-domain. This can also be written as  $\sigma = \zeta \omega_n$ . Partial fraction decomposition gives us:

$$X(s) = X_1(s) + X_2(s) = \frac{r_1}{s - p_1} + \frac{r_2}{s - p_2} \quad (261)$$

where  $A(s)$  and  $B(s)$  are polynomial expansions, similar to equation 207. The linearity of the system allows us to define  $X(s) = X_1(s) + X_2(s)$  and therefore,  $x(t) = x_1(t) + X_2(t)$  in the time domain.

From equation 261,  $r_1$  and  $r_2$  represent the residuals of the solution as a whole, but we want to extract the steady state (harmonic component) and transient state (exponential component) of the signals. For that, we need a joint residual value that we will call  $r_0$ . With this in mind, it can be shown that  $r_1 = ir_0$ , and that  $r_2 = ir_0$

$$\begin{aligned} x_1(t) &= \frac{-ir_0}{s - p_1}, \quad p_1 = \sigma + i\omega_d \\ &= \mathcal{L}[X_1(s)]^{-1} \\ &= \mathcal{L}\left[\frac{-ir_0}{s - p_1}\right]^{-1} \\ &= -ir_0 e^{p_1 t} \\ &= -ir_0 e^{(\sigma+i\omega_d)t} \\ &= -ir_0 e^{\sigma t} e^{i\omega_d t} \\ &= -ir_0 e^{\sigma t} (\cos(\omega_d t) + i \sin(\omega_d t)) \\ &= r_0 e^{\sigma t} (-i \cos(\omega_d t) + \sin(\omega_d t)) \end{aligned} \quad (262)$$

similarly,

$$\begin{aligned} x_2(t) &= \frac{ir_0}{i - p_1}, \quad p_2 = \sigma + i\omega_d \\ &= r_0 e^{\sigma t} (i \cos(\omega_d t) + \sin(\omega_d t)) \end{aligned} \quad (263)$$

the time series response can be rebuilt as

$$\begin{aligned} x_1(t) + x_2(t) &= r_0 e^{\sigma t} (-i \cos(\omega_d t) + \sin(\omega_d t)) + r_0 e^{\sigma t} (i \cos(\omega_d t) + \sin(\omega_d t)) \\ &= 2r_0 e^{\sigma t} \sin(\omega_d t) \end{aligned} \quad (264)$$

For stable systems, we expect  $\sigma > 0$  so,  $\sigma = -|\sigma|$  as the poles are on the left hand side for a stable response

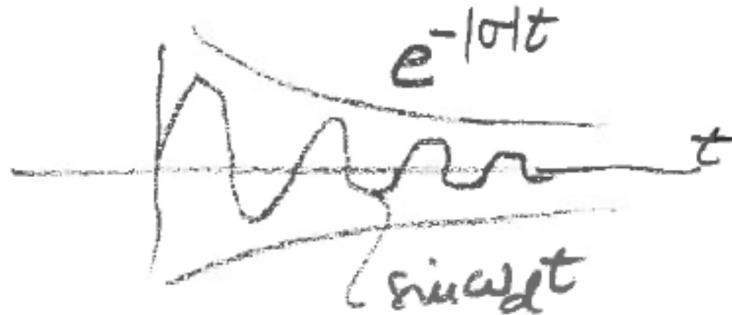


Figure 3.24: Time series response for the signal.

**Example 3.5** For the system in the S-domain expressed as,

$$X(s) = \frac{1}{s+2} \frac{2}{s+1} \quad (265)$$

use MATLAB to compute the roots and residuals, and use the roots and residuals to compute  $X(s)$ .

$$\begin{aligned} X(s) &= \frac{1}{s+2} \frac{2}{s+1} \\ &= \frac{(s+1)+2(s+2)}{(s+1)(s+2)} \\ &= \frac{3s+5}{s^2+3s+2} \\ &= \frac{B(s)}{A(s)} \end{aligned} \quad (266)$$

The `residue` command can be used such as:

Listing 1: MATLAB code to find poles and residuals.

---

```

1 % Define B and A
2 B = [3, 5];
3 A = [1, 3, 2]
4
5 % find the poles and the residues
6 [r, p, k] = residue(B, A)
7
8 % use poles and residuals to calculate the A and B
9 [B_new, A_new] = residue(r, p, k)

```

---

this returns  $r_1 = 1$ ,  $r_2 = 2$ ,  $p_1 = -2$ ,  $p_2 = -1$ ,  $k = \text{void}$ . Inserting this back into the ex-

pression yields

$$\frac{r_1}{s-p_1} + \frac{r_2}{s-p_2} = \frac{1}{s+2} + \frac{2}{s+1} \quad (267)$$

which checks out! Now, converting the partial fraction expansion back to the ratio of two polynomials results in  $A(s) = 1, 3, 2$ , and  $B(s) = 3, 5$ . So this also check outs!

## 3.8 Dominant Poles

Given

$$X(s) = \frac{B(s)}{A(s)} = \frac{r_1}{s-p_1} + \frac{r_2}{s-p_2} + \cdots + \frac{r_k}{s-p_k} + \cdots \quad (268)$$

where  $p_k$  is either a complex number ( $p_k \in \mathbb{C}$ ) and a conjugate pair

$$p_k = \sigma_k \pm i\omega_k \quad (269)$$

or a real number ( $p_k \in \mathbb{R}$ )

$$p_k = \sigma_k \quad (270)$$

Again, knowing that the inverse Laplace of  $X(s)$  results in

$$x(t) = r_1 e^{p_1 t} + r_2 e^{p_2 t} + \cdots + r_k e^{p_k t} + \cdots \quad (271)$$

we can solve for the time series response. For controls, we are interested in the long term behavior of  $x(t)$ , i.e., to find  $x_{ss} = \lim_{t \rightarrow \infty} [x(t)]$  which is the steady state response of the system. Note, that every term in the expansion will have the following form

$$r_k e^{p_k t} = r_k e^{\sigma_k t} \cdot e^{i\omega_k t} \quad (272)$$

where:

- $r_k e^{\sigma_k t}$  is the exponential function
- $e^{i\omega_k t}$  is the harmonic oscillation

we distinguish the following possible cases

### 3.8.1 Case A (Unstable):

If at least one  $\sigma_k$  is positive, then  $e^{\sigma_k t} \rightarrow \infty$ , unstable system.

### 3.8.2 Case B (Stable):

If system is stable,  $\sigma_k < 0$  (all poles on the left hand side) then:

**Case B1 - Dominant Poles on the imaginary axis:** If terms with  $\sigma_k = 0$  exist, than these will be maintain oscillations as the rest of the poles die out. This means the poles placed on the imaginary axis, if they exist, are dominant poles.

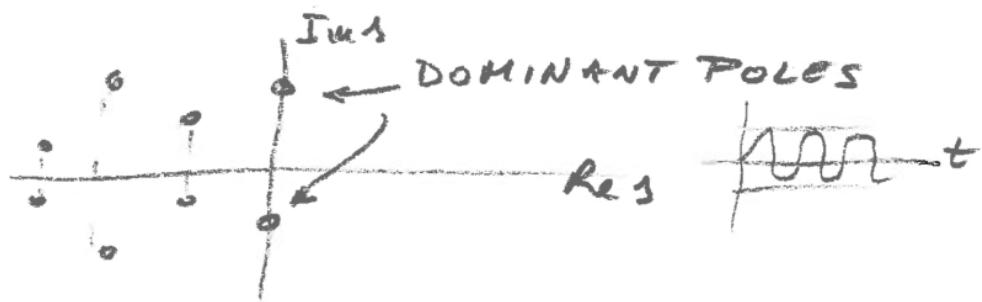


Figure 3.25: System with dominant poles on the imaginary axis.

**Case B2 - Dominant Poles to the left of the imaginary axis:** If terms with  $\sigma_k = 0$  do not exist, than the dominant poles are the poles closest to the imaginary axis because they take the longest to die out, having small  $\sigma_k$  values (small damping).

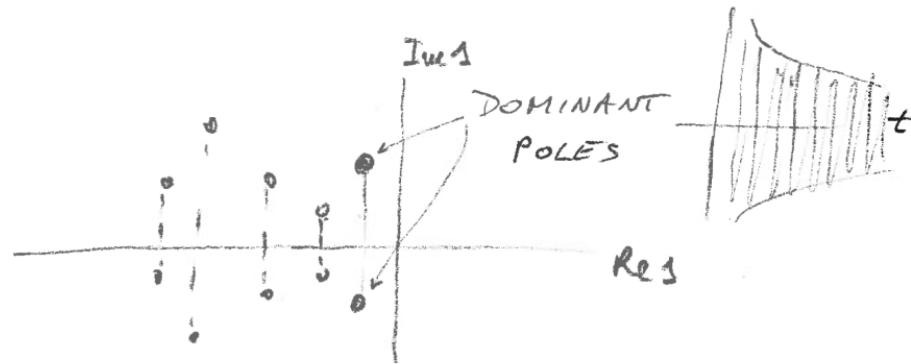


Figure 3.26: System with dominant poles to the left of the imaginary axis.

## Table of Laplace Transforms for Vibrations

This is a partial lists of important Laplace transforms for vibrations that assumes zero initial conditions,  $0 < t$ , and  $\zeta < 1$ .

$f(t)$	$\mathcal{L}[f(t)] = F(s)$		$f(t)$	$\mathcal{L}[f(t)] = F(s)$	
$\delta(t)$	1	(1)		$\frac{1}{\omega^3}(\omega t - \sin(\omega t))$	$\frac{1}{s^2(s^2 + \omega^2)}$ (17)
$\delta(t - t_0)$	$e^{-st_0}$	(2)		$\frac{1}{2\omega^3}(\sin(\omega t) - \omega t \cos(\omega t)) \dots$	
1	$\frac{1}{s}$	(3)			$\frac{1}{(s^2 + \omega^2)^2}$ (18)
$e^{at}$	$\frac{1}{s-a}$	(4)		$\frac{t}{2\omega} \sin(\omega t)$	$\frac{s}{(s^2 + \omega^2)^2}$ (19)
$\sin(\omega t)$	$\frac{\omega}{s^2 + \omega^2}$	(5)		$t \sin(\omega t)$	$\frac{2\omega s}{(s^2 + \omega^2)^2}$ (20)
$\cos(\omega t)$	$\frac{s}{s^2 + \omega^2}$	(6)		$t \cos(\omega t)$	$\frac{s^2 - \omega^2}{(s^2 + \omega^2)^2}$ (21)
$\sinh(\omega t)$	$\frac{\omega}{s^2 - \omega^2}$	(7)		$e^{at} \sin(\omega t)$	$\frac{\omega}{(s-a)^2 + \omega^2}$ (22)
$\cosh(\omega t)$	$\frac{s}{s^2 - \omega^2}$	(8)		$e^{at} \cos(\omega t)$	$\frac{s-a}{(s-a)^2 + \omega^2}$ (23)
$\frac{1}{\omega^2}(1 - \cos(\omega t))$	$\frac{1}{s(s^2 + \omega^2)}$	(9)		$e^{at} \sinh(\omega t)$	$\frac{\omega}{(s-a)^2 - \omega^2}$ (24)
$\frac{1}{\omega_d} e^{-\zeta \omega t} \sin(\omega_d t)$	$\frac{1}{s^2 + 2\zeta \omega s + \omega^2}$	(10)		$e^{at} \cosh(\omega t)$	$\frac{s-a}{(s-a)^2 - \omega^2}$ (25)
$1 - \frac{\omega}{\omega_d} e^{-\zeta \omega t} \sin(\omega_d t + \phi)$ , $\phi = \cos^{-1}(\zeta) \dots$	$\frac{\omega^2}{s(s^2 + 2\zeta \omega s + \omega^2)}$	(11)		$\frac{1}{\omega_2} \sin(\omega_2 t) - \frac{1}{\omega_1} \sin(\omega_1 t) \dots$	$\frac{\omega_1^2 - \omega_2^2}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)}$ (26)
$\frac{t^{n-1}}{(n-1)!}$ , $n = 1, 2, \dots$	$\frac{1}{s^n}$	(12)		$\cos(\omega_2 t) - \cos(\omega_1 t)$	$\frac{s(\omega_1^2 - \omega_2^2)}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)}$ (27)
$t^n$ , $n = 1, 2, \dots$	$\frac{n!}{s^{n+1}}$	(13)		$e^{at} f(t)$	$F(s-a)$ (28)
$t^n e^{\omega t}$ , $n = 1, 2, \dots$	$\frac{n!}{(s-\omega)^{n+1}}$	(14)		$f(t-a) \Phi(t-a)$	$e^{-as} F(s)$ (29)
$\frac{1}{\omega}(1 - e^{-\omega t})$	$\frac{1}{s(s+\omega)}$	(15)		$\Phi(t-a)$	$\frac{e^{-as}}{s}$ (30)
$\frac{1}{\omega^2}(e^{-\omega t} + \omega t - 1)$	$\frac{1}{s^2(s+\omega)}$	(16)		$f'(t)$	$sF(s) - f(0)$ (31)

## 4 Time Series Response

### 4.1 Transfer Functions Block Diagrams

Transfer functions can be created using:

- Polynomial model (numerator / denominator)

$$G(s) = \frac{B(s)}{A(s)} = \frac{b_m s^m + b_{m-1} s^{m-1} + \dots + b_0}{a_n s^n + a_{n-1} s^{n-1} + \dots + a_0} \quad (273)$$

–  $n$  = order of transfer function models where  $m < n$

- Zero-pole-gain model (numerator / denominator)

$$G(s) = k \frac{(s - z_1)(s - z_2) \cdots (s - z_m)}{(s - p_1)(s - p_2) \cdots (s - p_n)} \quad (274)$$

– zeros:  $z_1, z_2, \dots, z_m$  roots of  $B(s) = 0$

– poles:  $p_1, p_2, \dots, p_n$  roots of  $A(s) = 0$

– gain:  $k$

- Time constant model

$$G(s) = \frac{k}{s^N} \cdot \frac{(T_a + 1)(T_b + 1) \cdots}{(T_1 + 1)(T_2 + 1) \cdots} \quad (275)$$

–  $N$  = type of transfer function model

#### 4.1.1 Create transfer function using numerator and denominator coefficients

MATLAB can be used to create continuous-time single-input, single-output (SISO) transfer functions from their numerator and denominator coefficients using `tf`. To use the `tf` function, you must have the Control System Toolbox licensed and installed. To find out if you do, type: `ver control` in your Command Window or a script.

##### Method I

Create the transfer function

$$G(s) = \frac{s}{s^2 + 3s + 2} \quad (276)$$

Listing 2: MATLAB code for Method I.

---

```
1 num = [1 0];
2 dem = [1 3 2];
3 G = tf(num, dem)
4 % To use the tf function, you must have the Control System Toolbox
5 % licensed and installed. To find out if you do, type:
6 % ver control in your Command Window or a script.
```

---

where `num` and `dem` are the numerator and denominator polynomial coefficients in descending powers of  $s$ . For example, `den = [1 3 2]` represents the denominator polynomial  $\frac{s^2 + 3s + 2}{s}$

`G` is a `tf` model object, which is a data container for representing the transfer function in polynomial form.

##### Method II

Alternatively, you can specify the transfer function  $G(s)$  as an expression in  $s$ -domain.

1. Create a transfer Function model for the variable  $s$
2. Specify  $G(s)$  as a ratio of polynomials in  $s$

Listing 3: MATLAB code for Method II.

---

```

1 s = tf('s');
2 G = s/(s^2+3*s+2)

```

---

Therefore, the full expression of  $G(s)$  can be written as

$$G(s) = \frac{B(s)}{A(s)} = \frac{s}{s^2 + 3s + 2} = \frac{b_1s + b_0}{a_1s^2 + a_2s + a_0} \quad (277)$$

where

$$B(s) = b_1s + b_0 \quad (278)$$

resulting in  $b_1 = 1$ , and  $b_0 = 0$ ; or,  $B = [1 \ 0]$ . Moreover,

$$A(s) = a_1s^2 + a_2s + a_0 \quad (279)$$

where  $a_1 = 1$ ,  $a_2 = 3$ , and  $a_0 = 2$ ; or,  $A = [1 \ 3 \ 2]$ .

#### 4.1.2 Create transfer function using Zeros, Poles, and Gain

MATLAB can be used to create continuous-time single-input, single-output (SISO) transfer functions in factored form using `zpk`. Create the factored transfer function

$$G(s) = 5 \frac{s}{(s - 1 - i)(s - 1 + i)(s - 2)} \quad (280)$$

Listing 4: MATLAB code to create a transfer function using Zeros, Poles, and Gain.

---

```

1 Z = 0;
2 P = [-1-1i -1+1i -2];
3 K = 5;
4 G = zpk(Z,P,K)

```

---

where  $Z$  and  $P$  are zeros and poles (the roots of the numerator and denominator respectively).  $K$  is the gain of the factored from. Solving fore the poles  $p_1$ ,  $p_2$ , and  $p_3$  of  $G(s)$ ;

$$G(s) = 5 \frac{s}{(s - 1 - i)(s - 1 + i)(s - 2)} \quad (281)$$

$$= 5 \frac{s - 0}{[s - (-1 - i)][(s - (-1 - i)][(s - (-2)]]} \quad (282)$$

where  $K = 5$ ,  $s - 0 = s - z_1 \rightarrow z_1 = 0$ , and  $Z = [0]$ . Therefore,

$$[s - (-1 - i)][(s - (-1 - i)][(s - (-2)] = (s - p_1)(s - p_2)(s - p_3) \quad (283)$$

this leads to

$$G(s) = k \frac{s - z_1}{(s - p_1)(s - p_2)(s - p_3)} \quad (284)$$

therefore,  $G(s)$  has a real pole at  $s = -2$  and a pair of complex poles as  $s = -1 \pm i$ . The vector  $P = [-1-1i -1+1i -2]$  specifies these pole locations.

## 4.2 Order versus Type

A system has a “Type” and an “Order”, which have different meanings.

- Order =  $n$ , highest exponent of  $s$  in the denominator.  $n$  is the number of poles.
- Type =  $N$ , exponent of the factored out  $s$  in the denominator.  $N$  is the number of poles in origin ( $p = 0$ ).

Consider the 1<sup>st</sup>-order mass-damper system (no stiffness) as shown in figure X with the transfer function

[make a figure](#)

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s} \quad (285)$$

The transfer function can easily be written in the basic form

$$G(s) = \frac{b_0}{a_2 s^2 + a_1 s} \quad (286)$$

where  $a_2 = 1$ ,  $a_1 = 2\zeta\omega_n$ ,  $b_0 = \omega_n^2$ . There the presence of  $a_2$  means its a 2<sup>nd</sup> order system. To solve for the type of the system,  $s$  must be factored out of the denominator, leading to:

$$\begin{aligned} G(s) &= \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s} \\ &= \frac{\omega_n^2}{s(s + 2\zeta\omega_n)} \\ &= \frac{\frac{\omega_n}{2\zeta}}{s} \cdot \frac{1}{\frac{1}{2\zeta\omega_n}s + 1} \\ &= \frac{K}{s^N} \cdot \frac{1}{T_1 s + 1} \end{aligned} \quad (287)$$

where  $K = \frac{\omega_n}{2\zeta}$ ,  $N = 1$ , and  $T_1 = \frac{1}{2\zeta\omega_n}$ .  $N = 1$  means that is a Type 1 system. Therefore, this is a 2<sup>nd</sup> order system of Type 1, “Type” and “Order” have different meanings. Table ?? reports the types and orders for different transfer functions.

need to find some real-world examples of orders and types.

Table 2: Examples of types and orders for different transfer functions.

transfer function	Type	Order
$G(s) = \frac{1}{Ts+1}$	0	1
$G(s) = \frac{1}{cs} = \frac{1/c}{s}$	1	1
$G(s) = \frac{1}{Js^2} = \frac{1/J}{s^2}$	2	2
$G(s) = \frac{1}{Js^2+cs} = \frac{K/c}{s} \cdot \frac{1}{s+J/c}$	1	2
$G(s) = \frac{K}{J^4} \frac{T_d s + 1}{(T_1 s + 1)(T_2 s + 1)(T_3 s + 1)}$ $= \frac{b_1 s + b_0}{s^4(a_3 s^3 + a_2 s^2 + a_1 s + a_0)}$ $= \frac{b_1 s + b_0}{(a_3 s^7 + a_2 s^6 + a_1 s^5 + a_0 s^4)}$	4	7

## 4.3 Time Response

Time response calculations are obtained using the Laplace transforms where the Laplace transform is

$$X(s) = G(s)F(s) \quad (288)$$

and the time response is

$$x(t) = \mathcal{L}[X(s)]^{-1} \quad (289)$$

### 4.3.1 Time Series Response for a Step Function

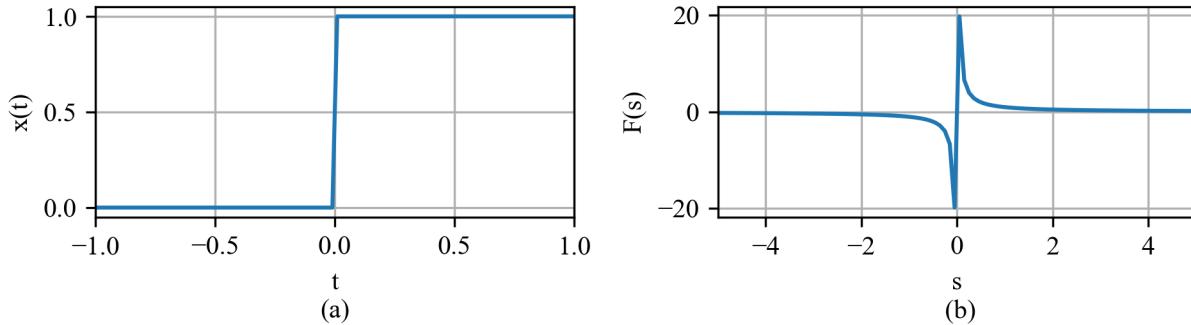


Figure 4.1: Step function; showing the (a) time domain and; (c) the s-space.

A step function is expressed as the following Laplace pair:

$$\text{LT pair} = \begin{cases} f(t) & 1(t), \quad t > 0 \\ F(s) & \frac{1}{s} \end{cases} \quad (290)$$

therefore, the time response of the systems is expressed as

$$x(t) = \mathcal{L} \left[ G(s) \frac{1}{s} \right]^{-1} \quad (291)$$

In MATLAB, this is expressed as:

**Listing 5:** MATLAB code for the time-series response of a step function.

---

```
1 x_t = step(G)
```

---

### 4.3.2 Time Series Response for a Impulse Function

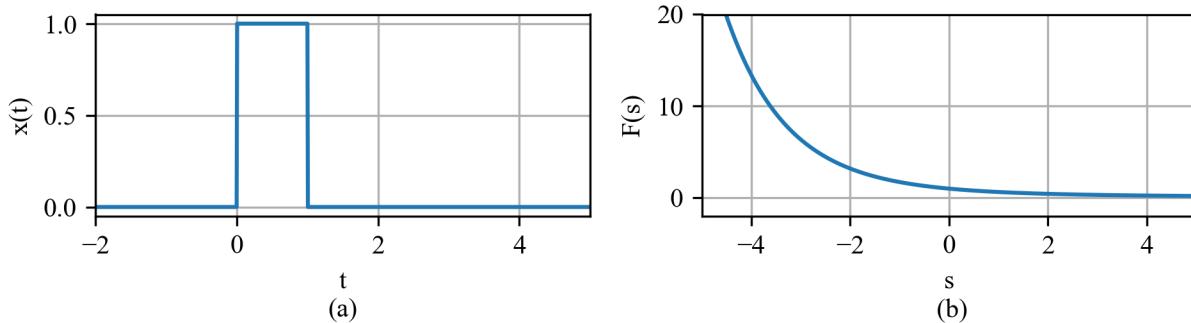


Figure 4.2: Pulse function; showing the (a) time domain and; (c) the s-space.

An impulse function is expressed as the following Laplace pair:

$$\text{LT pair} = \begin{cases} f(t) & p(t; \tau) \\ F(s) & \frac{1-e^{-st\tau}}{s\tau} \end{cases} \quad (292)$$

therefore, the time response of the systems is expressed as

$$x(t) = \mathcal{L} \left[ G(s) \frac{1-e^{-st\tau}}{s\tau} \right]^{-1} \quad (293)$$

In MATLAB, this is expressed as:

**Listing 6:** MATLAB code for the time-series response of a step function.

---

```
1 x_t = impulse(G)
```

---

### 4.3.3 Time Series Response for a Ramp Function

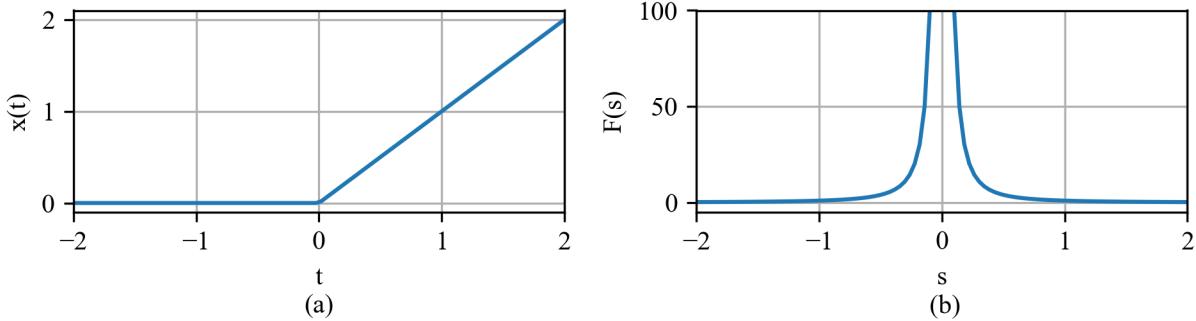


Figure 4.3: Ramp function; showing the (a) time domain and; (c) the s-space.

An ramp function is expressed as the following Laplace pair:

$$\text{LT pair} = \begin{cases} f(t) & t, \quad t > 0 \\ F(s) & \frac{1}{s^2} \end{cases} \quad (294)$$

therefore, the time response of the systems is expressed as

$$x(t) = \mathcal{L}^{-1} \left[ G(s) \frac{1}{s^2} \right] \quad (295)$$

In MATLAB, this is expressed as:

Listing 7: MATLAB code for the time-series resposne of a step function.

---

```
1 x_t = impulse(G/(s^2))
```

---

Note that for the MATLAB code, we used the property:

$$X(s) = G(s) \frac{1}{s^2} = \left( \frac{G(s)}{s^2} \right) \cdot 1 \quad (296)$$

where 1 is the Laplace transform of an impulse. Note that `ramp` is not an option in MATLAB as this command is already used to generate a time-series ramp signal.

## 4.4 1<sup>st</sup> Order System Time Response

The first order equation of motion is

$$T\dot{x}(t) + x(t) = f(t) \quad (297)$$

where  $x(0) = 0$  is the initial condition and  $T$  is a time constant for the first order system. The Laplace transform gives us

$$\begin{aligned} x &\rightarrow X(s) \\ \dot{x} &\rightarrow sX(s) \\ f(t) &\rightarrow F(s) \end{aligned} \quad (298)$$

therefore, the s-domain equation is:

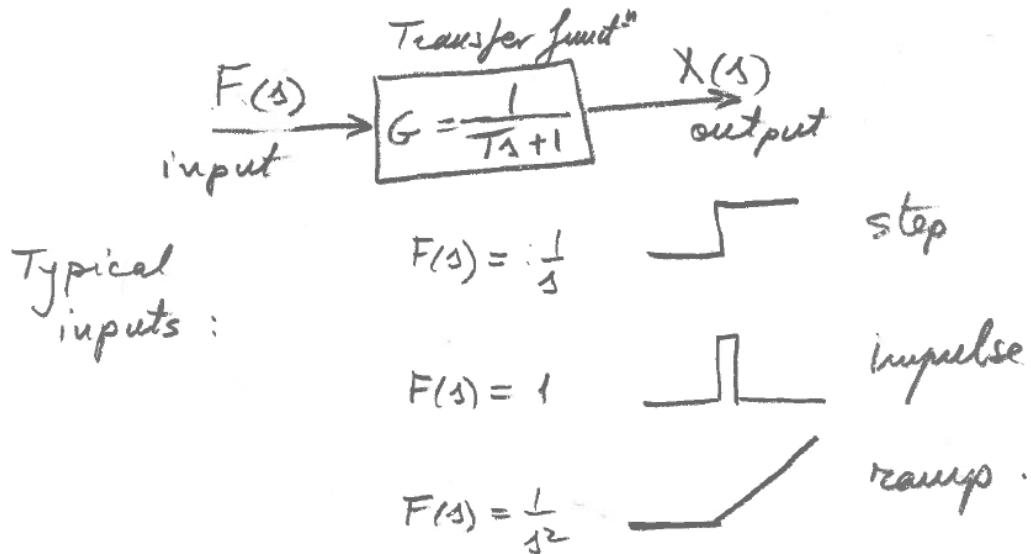
$$TsX(s) + X(s) = F(s) \quad (299)$$

where:

$$X(s) = \frac{F(s)}{Ts+1} = \frac{1}{Ts+1}F(s) = G(s)F(s) \quad (300)$$

therefore, the transfer function is:

$$G(s) = \frac{1}{Ts+1} \quad (301)$$



#### 4.4.1 Step response of a 1<sup>st</sup> Order System

$$X(s) = G(s)F(s) \quad (302)$$

$$\begin{aligned} &= \frac{1}{Ts+1} \cdot \frac{1}{s} \\ &= \frac{1}{s(Ts+1)} \end{aligned} \quad (303)$$

Therefore, solving for  $\mathcal{L}[X(s)]^{-1}$  yields

$$x(t) = 1 - e^{-t/T} \quad (304)$$

or



#### 4.4.2 Impulse response of a 1<sup>st</sup> Order System

$$X(s) = G(s)F(s) \quad (305)$$

$$\begin{aligned} &= \frac{1}{Ts+1} \cdot 1 \\ &= \frac{1}{Ts+1} \end{aligned} \quad (306)$$

Therefore, solving for  $\mathcal{L}[X(s)]^{-1}$  yields

$$x(t) = \frac{1}{T} e^{-t/T} \quad (307)$$

or



#### 4.4.3 Ramp response of a 1<sup>st</sup> Order System

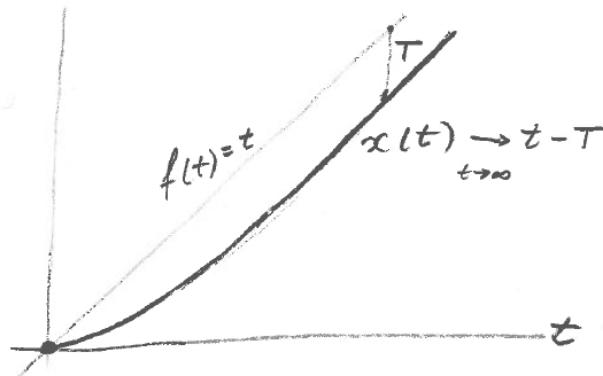
$$X(s) = G(s)F(s) \quad (308)$$

$$\begin{aligned} &= \frac{1}{Ts+1} \cdot \frac{1}{s^2} \\ &= \frac{1}{s^2(Ts+1)} \end{aligned} \quad (309)$$

Therefore, solving for  $\mathcal{L}[X(s)]^{-1}$  yields

$$\begin{aligned} x(t) &= t - T + Te^{-t/T} \\ &= t - T(1 - e^{-t/T}) \end{aligned} \quad (310)$$

or



Moreover,

$$x(t) = t - T + Te^{-t/T} \xrightarrow[t \rightarrow \infty]{} t - T \quad (311)$$

#### 4.4.4 Summary of the First Order System Responses

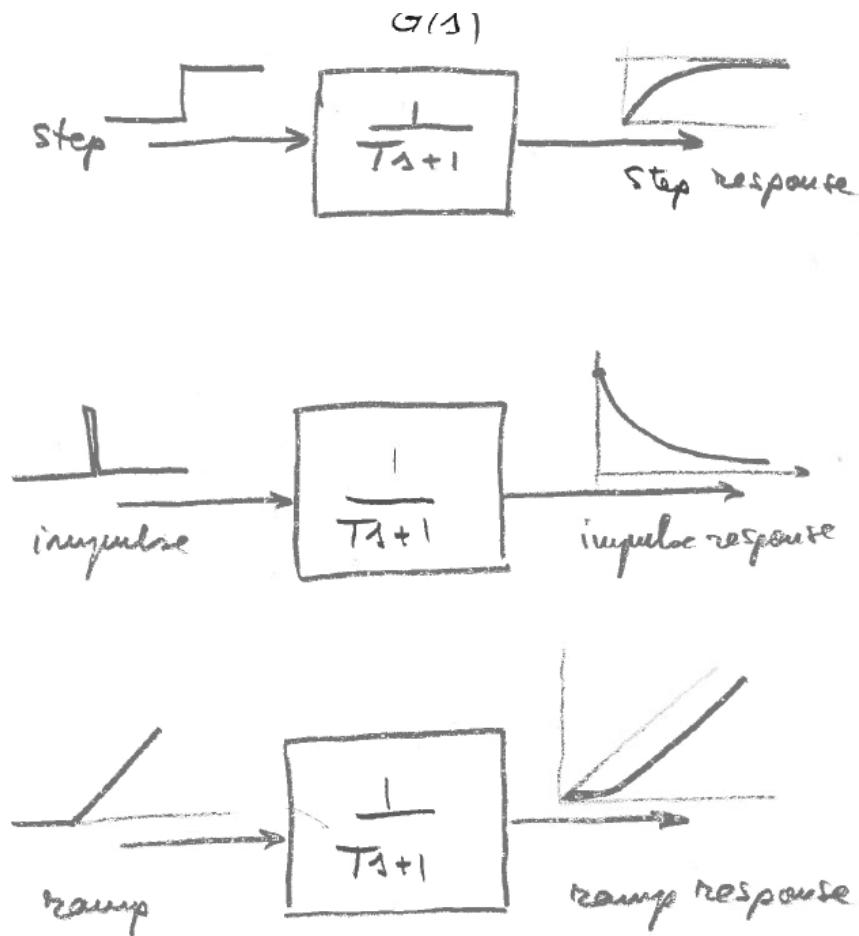


Figure 4.4: A summary of the first order system responses.

Listing 8: MATLAB code for time series responses of 1<sup>st</sup> order system.

```
1 %{
2 This program studies time response of 1st order systems
3 %}
4 clc
5 clear
6 close all
7
8 format compact
9
10 %% Given data
11 T=2.5; % time response for the 1D system
12
13 %% time range setup
14 T_max = 10; % run the test to 10 seconds
15 dt = T_max*1e-4; % find the delta-t value
16 t = 0:dt:T_max; % build the time vector
17
18 %% Define system
19 B = [1];
20 A = [T 1];
21 G = tf(B,A);
22
23 %% create the figure enviornment
24 figure(1)
25
26 %% step response
27 subplot(3,1,1)
28 hold on
29 step(G,t)
30 ylim([0 1.2])
31
32 %% Impulse response
33 subplot(3,1,2)
34 impulse(G,t)
35
36 %% Ramp response
37 F_ramp = tf([1],[1 0 0])
38 subplot(3,1,3)
39 impulse(G*F_ramp,t)
40 title('Ramp Response') % need to set manually.
```

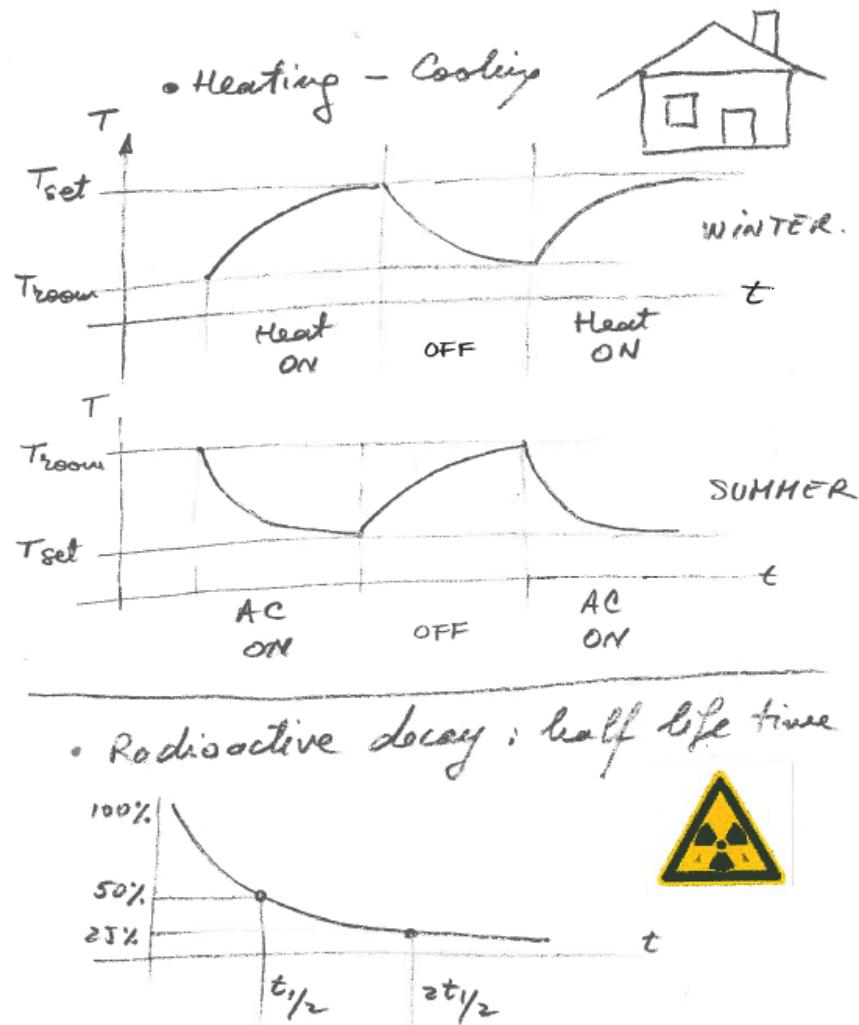


Figure 4.5: Examples of first order systems in the real world.

## 4.5 2<sup>nd</sup> Order System Time Response

The ordinary differential equation for the equation of motion of a 2<sup>nd</sup> order system can be expressed as

$$\ddot{x} + 2\zeta\omega_n\dot{x} + \omega_n^2x = \omega_n^2f(t) \quad (312)$$

with the initial conditions  $\ddot{x}(0) = 0$ , and  $\dot{x}(0) = 0$ . The Laplace transform gives us

$$\begin{aligned} x &\rightarrow X(s) \\ \dot{x} &\rightarrow sX(s) \\ \ddot{x} &\rightarrow s^2X(s) \\ f(t) &\rightarrow F(s) \end{aligned} \quad (313)$$

Taking the Laplace transform of the equation of motion yields

$$s^2X(s) + 2\zeta\omega_n sX(s) + \omega_n^2X(s) = \omega_n^2F(s) \quad (314)$$

Pulling  $X(s)$  out of the first equation results in

$$(s^2 2\zeta \omega_n s + \omega_n^2) X(s) = \omega_n^2 F(s) \quad (315)$$

next, we can solve for  $X(s)$

$$X(s) = \frac{\omega_n^2}{s^2 + 2\zeta \omega_n s + \omega_n^2} F(s) \quad (316)$$

As  $X(s) = G(s)F(s)$ , it we can show that

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta \omega_n s + \omega_n^2} \quad (317)$$

#### 4.5.1 Step Response for a 2<sup>nd</sup> Order System

$$X(s) = G(s)F(s) \quad (318)$$

$$\begin{aligned} &= \frac{\omega_n^2}{s^2 + 2\zeta \omega_n s + \omega_n^2} \cdot \frac{1}{s} \\ &= \frac{\omega_n^2}{s(s^2 + 2\zeta \omega_n s + \omega_n^2)} \end{aligned} \quad (319)$$

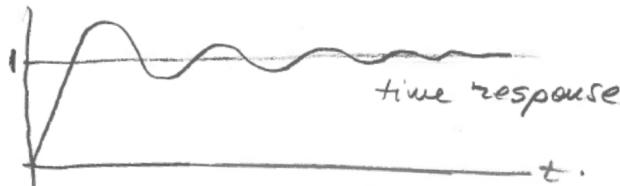
Therefore, solving for  $\mathcal{L}[X(s)]^{-1}$  yields

$$x(t) = 1 - \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta \omega_n t} \sin(\omega_d t + \phi) \quad (320)$$

where

$$\begin{aligned} \phi &= \tan^{-1} \frac{\sqrt{1-\zeta^2}}{\zeta} \\ &= \sin^{-1} \sqrt{1-\zeta^2} \end{aligned} \quad (321)$$

or



Proof

$$7] x_c(t) = e^{-\zeta \omega_n t} (A \cos \omega_d t + B \sin \omega_d t) ; A, B \quad \text{ODE derivation}$$

$$\begin{aligned} &= e^{-\zeta \omega_n t} C \sin(\omega_d t + \varphi) \\ \ddot{x} + 2\zeta \omega_n \dot{x} + \omega_n^2 x &= \omega_n^2 f(t) ; \omega_n^2 f = \frac{f^*}{m} \rightarrow f = \frac{f^*}{k} \\ \text{Step input, } f(t) &= 1(t) \end{aligned}$$

$f = 1$  

$$\begin{aligned} x_p(t) &= D, \quad \dot{x}_p - \ddot{x}_p = 0 \\ \omega_n^2 D &= \omega_n^2 \rightarrow D = 1 \end{aligned}$$

$$x_p(t) = 1$$

$$x(t) = e^{-\zeta \omega_n t} C \sin(\omega_d t + \varphi) + 1$$

$$\dot{x}(t) = (-\alpha \sin \varphi + \omega_d \cos \varphi) C e^{-\zeta t}$$

$$x(0) = C \sin \varphi + 1 = 0$$

$$\dot{x}(0) = (-\alpha \sin \varphi + \omega_d \cos \varphi) = 0 \rightarrow \tan \varphi = \frac{\omega_d}{\alpha}$$

$$\tan \varphi = \frac{\omega_n \sqrt{1-\zeta^2}}{\zeta \omega_n} = \frac{\sqrt{1-\zeta^2}}{\zeta} \rightarrow \varphi = \tan^{-1} \frac{\sqrt{1-\zeta^2}}{\zeta}$$

$$\sin \varphi = \frac{\tan \varphi}{1 + \tan^2 \varphi} = \frac{1 - \zeta^2}{\zeta^2 + 1 - \zeta^2} = 1 - \zeta^2$$

$$\sin \varphi = \sqrt{1 - \zeta^2} \quad \varphi = \sin^{-1} \sqrt{1 - \zeta^2}$$

$$C = -\frac{1}{\sin \varphi} = -\frac{1}{\sqrt{1 - \zeta^2}}$$

$$x(t) = 1 - \frac{1}{\sqrt{1 - \zeta^2}} e^{-\zeta \omega_n t} \sin(\omega_d t + \varphi) \quad \text{unit response}$$

for  $f^* = k$

For  $f^* = F_0$

$$x(t) = \frac{F_0}{k} \left[ 1 - \frac{1}{\sqrt{1 - \zeta^2}} e^{-\zeta \omega_n t} \sin(\omega_d t + \varphi) \right] = x_{sf} \cdot x(t)$$

$\hookrightarrow x_{sf} = F_0/k$  static displacement

89/523

14) Step response  $f(t) = 1(t)$ ,  $F(s) = \frac{1}{s}$

$$X(s) = \frac{\omega_n^2}{(\zeta + \alpha)^2 + \omega_d^2} \cdot \frac{1}{s} = \frac{A}{s} + \frac{Ds + E}{(s + \zeta)^2 + \omega_d^2}$$

ILT by  
partial  
fraction  
expansion

$$A(s^2 + 2\zeta s + \omega_d^2) + Ds^2 + Es = \omega_n^2 s$$

$$s^2: A + D = 0$$

$$s^1: 2\zeta A + E = 0$$

$$s^0: A(\zeta^2 + \omega_d^2) = \omega_n^2 \quad A \left( \zeta^2 \omega_n^2 + (1 - \zeta^2) \omega_n^2 \right) = \omega_n^2$$

$$A = 1$$

$$D = -A = -1$$

$$E = -2\zeta A = -2\zeta$$

$$X(s) = \left[ \frac{1}{s} - \frac{s + 2\zeta}{(s + \zeta)^2 + \omega_d^2} \right] = \left[ \frac{1}{s} - \frac{\zeta + (\zeta + \alpha)}{(s + \zeta)^2 + \omega_d^2} \right]$$

$$x(t) = 1 - e^{-\alpha t} \left[ \frac{\zeta}{\omega_d} \sin \omega_d t + \cos \omega_d t \right]$$

$$x(t) = 1 - \bar{e}^{-\alpha t} \left( \frac{\zeta}{\sqrt{1-\zeta^2}} \sin \omega_d t + \cos \omega_d t \right)$$

$$x(t) = 1 - \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta \omega_n t} \sin(\omega_d t + \varphi) \quad \varphi = \tan^{-1} \frac{\sqrt{1-\zeta^2}}{\zeta}$$



90/523

### 4.5.2 Impulse response of a 2<sup>nd</sup> Order System

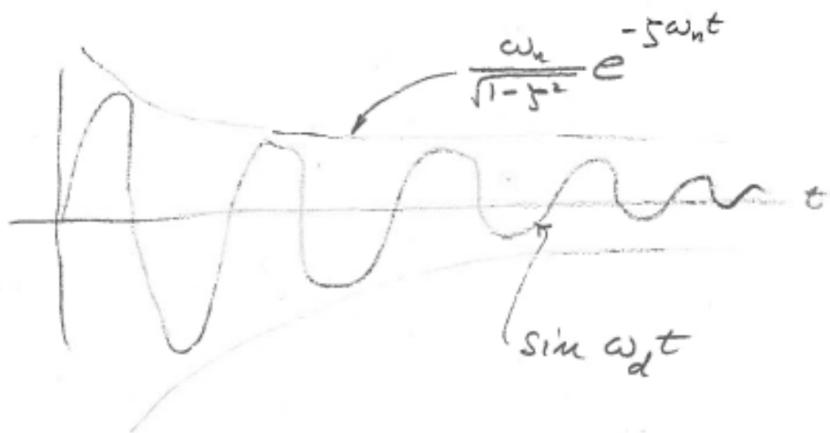
$$X(s) = G(s)F(s) \quad (322)$$

$$\begin{aligned} &= \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot 1 \\ &= \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \end{aligned} \quad (323)$$

Therefore, solving for  $\mathcal{L}[X(s)]^{-1}$  yields

$$x(t) = \frac{\omega_n}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t) \quad (324)$$

or



Proof

13a) *Long hand solution*

$$\frac{X}{\omega_n^2(s+p_1)(s+p_2)} = \frac{A}{s+p_1} + \frac{B}{s+p_2}$$

ILT by partial fraction expansion

$$As + Ap_2 + Bs + Bp_1 = 1$$

$$s^1: A + B = 0 \rightarrow B = -A$$

$$s^0: Ap_2 + Bp_1 = 1 \rightarrow A = \frac{1}{p_2 - p_1}, B = \frac{1}{p_1 - p_2}$$

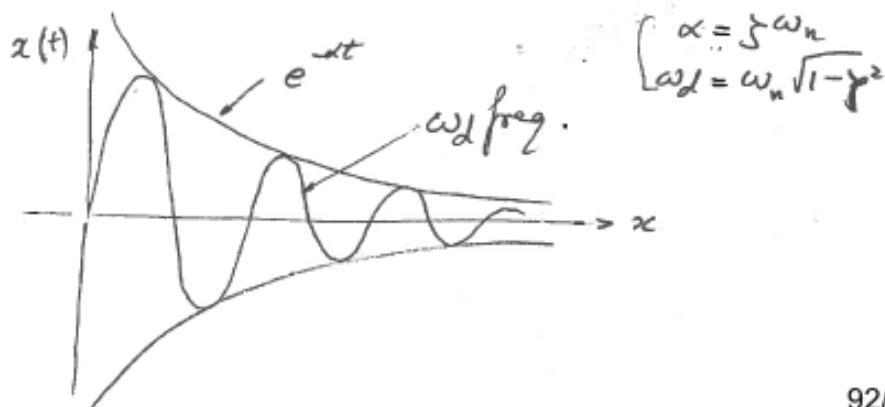
$$\frac{1}{(s+p_1)(s+p_2)} = \frac{1}{p_2 - p_1} \left( \frac{1}{s+p_1} - \frac{1}{s+p_2} \right)$$

$$x(t) = \frac{1}{p_2 - p_1} \left( e^{-p_1 t} - e^{-p_2 t} \right)$$

$$p_{1,2} = \pm \omega_n \sqrt{1 - \zeta^2} = \alpha \mp i\omega_d$$

$$p_2 - p_1 = \pm 2\omega_d$$

$$\begin{aligned} \frac{x(t)}{\omega_n^2} &= \frac{-1}{2i\omega_d} \left[ e^{-(\alpha - i\omega_d)t} - e^{-(\alpha + i\omega_d)t} \right] \\ &= \frac{1}{2i\omega_d} e^{-\alpha t} \left[ e^{i\omega_d t} - e^{-i\omega_d t} \right] \\ &= \frac{1}{\omega_d} e^{-\alpha t} \frac{e^{i\omega_d t} - e^{-i\omega_d t}}{2i} = \frac{e^{-\alpha t}}{\omega_d} \sin \omega_d t \end{aligned}$$



92/523

Alternative Way

$$\frac{1}{(s+p_1)(s+p_2)} \xrightarrow{\mathcal{L}^{-1}} \frac{1}{p_2-p_1} (e^{-p_1 t} - e^{-p_2 t})$$

Another way

Residue Theorem

$$\frac{1}{(s+p_1)(s+p_2)} = \frac{a_1}{s+p_1} + \frac{a_2}{s+p_2}$$

$$a_1 = \left[ \frac{1}{(s+p_1)} \right]_{s=-p_1} = \frac{1}{-p_1 + p_2}$$

$$a_2 = \left[ \frac{1}{(s+p_2)} \right]_{s=-p_2} = \frac{1}{p_1 - p_2} = -a_1$$

#### 4.5.3 Ramp response of a 2<sup>nd</sup> Order System

$$X(s) = G(s)F(s) \quad (325)$$

$$\begin{aligned} &= \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s^2} \\ &= \frac{\omega_n^2}{s^2(s^2 + 2\zeta\omega_n s + \omega_n^2)} \end{aligned} \quad (326)$$

Therefore, solving for  $\mathcal{L}[X(s)]^{-1}$  yields

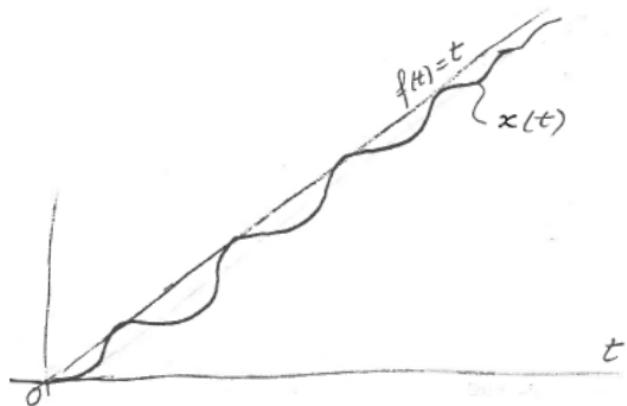
$$x(t) = t - \frac{2\zeta}{\omega_n} \left( 1 + \frac{1}{\sin \gamma_1} e^{-\zeta\omega_n t} \sin(\omega_d t - \gamma_1) \right) \quad (327)$$

where

$$\gamma_1 = \tan^{-1} \frac{2\zeta \sqrt{1-\zeta^2}}{1-2\zeta^2} \quad (328)$$

$$= \sin^{-1} 2\zeta \sqrt{1-\zeta^2} \quad (329)$$

or



Moreover,

$$\begin{aligned} \sin^2 \gamma_1 &= \frac{\tan^2 \gamma_1}{1 + \tan^2 \gamma_1} \\ &= \frac{4s^2(1-\zeta^2)}{(1-2\zeta^2)+4\zeta^2(1-\zeta^2)} \\ &= \frac{4s^2(1-\zeta^2)}{1-4s^2+4s^4+4s^2-4s^4} \\ &= 4s^2(1-\zeta^2) \end{aligned} \quad (330)$$

and

$$\sin \gamma_1 = 2\zeta \sqrt{1-\zeta^2} \quad (331)$$

Proof

PROOF of 2<sup>nd</sup> order system Ramp response

$$\mathcal{L}_p(t) = Dt + E \quad \text{ODE solution}$$

$$\dot{x}_p = D; \ddot{x}_p = 0$$

$$\ddot{x}_p + 2\zeta\omega_n \dot{x}_p + \omega_n^2 x_p = \omega_n^2 f(t)$$

$$2\zeta\omega_n D + \omega_n^2(Dt + E) = \omega_n^2 t$$

$$t: 2\zeta\omega_n D + \omega_n^2 E = 0 \rightarrow E = -\frac{2\zeta}{\omega_n} D$$

$$t': \omega_n^2 D = \omega_n^2 \rightarrow D = 1 \rightarrow E = \frac{2\zeta}{\omega_n}$$

$$x_p(t) = t - 2\zeta/\omega_n$$

$$x(t) = e^{-\zeta\omega_n t} C \sin(\omega_d t + \varphi) + t - \frac{2\zeta}{\omega_n}$$

$$\dot{x}(t) = (-\zeta\omega_n \sin \varphi + \omega_d \cos \varphi) C e^{-\zeta\omega_n t} + 1$$

$$x(0) = 0 \Rightarrow C \sin \varphi - \frac{2\zeta}{\omega_n} = 0 \rightarrow C \omega_n \sin \varphi = 2\zeta \quad (a)$$

$$\dot{x}(0) = 0 \Rightarrow (-\zeta\omega_n \sin \varphi + \omega_d \cos \varphi) C + 1 = 0$$

$$C (\zeta \sin \varphi - \omega_d \cos \varphi) = 1$$

$$C \zeta \omega_n \sin \varphi - C \omega_d \cos \varphi = 1$$

$$(a): 2\zeta^2 - C \omega_d \cos \varphi = 1$$

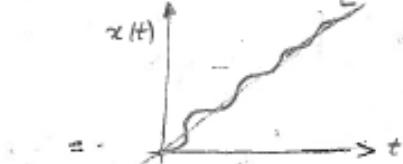
$$C \omega_n \sqrt{1-\zeta^2} \cos \varphi_1 = 2\zeta^2 - 1$$

$$C \omega_n \cos \varphi_1 = -\frac{1-2\zeta^2}{\sqrt{1-\zeta^2}} \quad (b)$$

$$\frac{(a)}{(b)} = \tan \varphi_1 = -\frac{2\zeta \sqrt{1-\zeta^2}}{1-2\zeta^2}$$

$$\varphi_1 = -\tan^{-1} \frac{2\zeta \sqrt{1-\zeta^2}}{1-2\zeta^2}, \quad C = \frac{2\zeta}{\omega_n} \cdot \frac{1}{\sin \varphi_1}$$

$$x(t) = t - \frac{2\zeta}{\omega_n} \left[ 1 - \frac{1}{\sin \varphi_1} e^{-\zeta\omega_n t} \sin(\omega_d t + \varphi_1) \right]$$



Note:  $\varphi_1$  is different from  $\varphi = \tan^{-1} \frac{-1}{\sqrt{1-\zeta^2}}$   
of step response

95/523

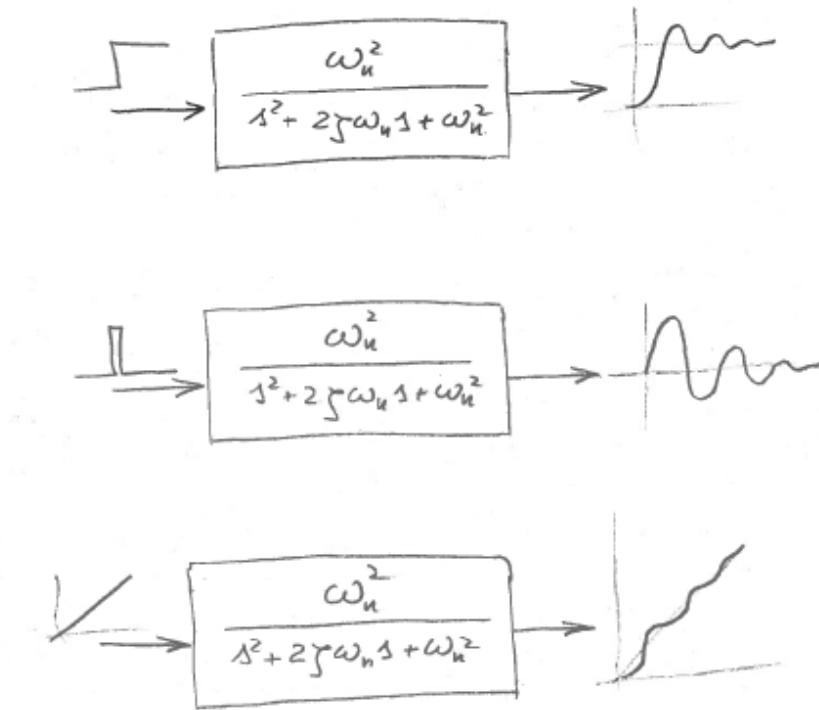
**4.5.4 Summary of the Second Order System Responses**

Figure 4.6: A summary of the second order system responses.

Listing 9: MATLAB code for time series responses of 2<sup>nd</sup> order system.

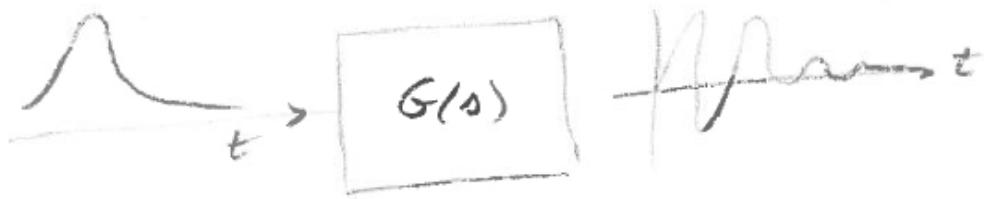
```

1  %{
2  This program studies time response of 2nd order systems
3  %}
4
5  clc
6  clear
7  close all
8
9  format compact
10
11 %% Given data
12 fn = 5; % time response for the 1D system
13 wn = 2*pi()*fn
14 z = 0.035
15
16 %% time range setup
17 T_max = 10; % run the test to 10 seconds
18 dt = T_max*1e-4; % find the delta-t value
19 t = 0:dt:T_max; % build the time vector
20
21 %% Define system
22 B = [wn^2];
23 A = [1 2*z*wn wn^2];
24 G = tf(B,A);
25
26 %% create the figure enviornment
27 figure(1)
28 xlim([0 1])
29
30 %% step response
31 subplot(3,1,1)
32 hold on
33 step(G,t)
34 ylim([0 2])
35
36 %% Impulse response
37 subplot(3,1,2)
38 impulse(G,t)
39 ylim([-35 35])
40
41 %% Ramp response
42 F_ramp = tf([1],[1 0 0])
43 subplot(3,1,3)
44 impulse(G*F_ramp,t)
45 xlim([0 1])
46 ylim([0 1])
47 title('Ramp Response') % need to set manually.

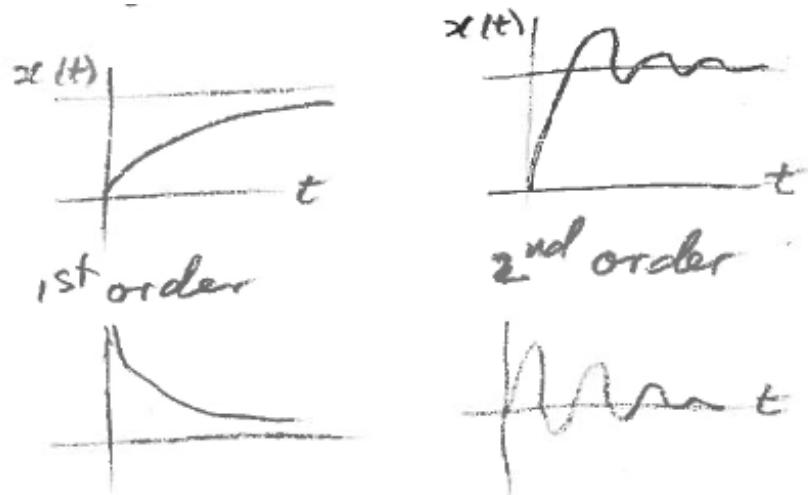
```

## 4.6 Stability of response

A system is stable if any stable input excitation produces a stable output response.



A response is stable if it remains bounded at  $t \rightarrow \infty$



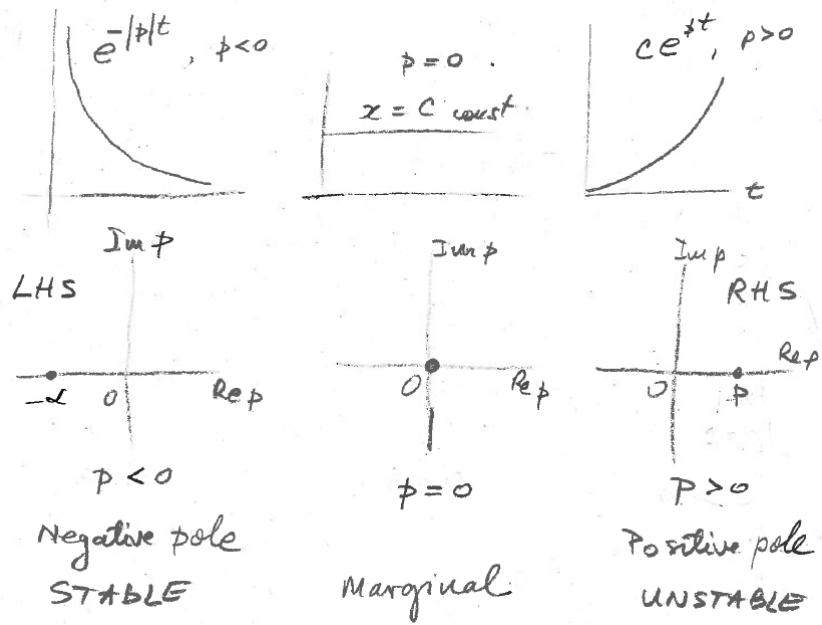
#### 4.6.1 Stability of 1<sup>st</sup>-Order Responses

$$X(s) = \frac{K}{s - p} \quad (332)$$

$$x(t) = Ke^{pt} \quad (333)$$

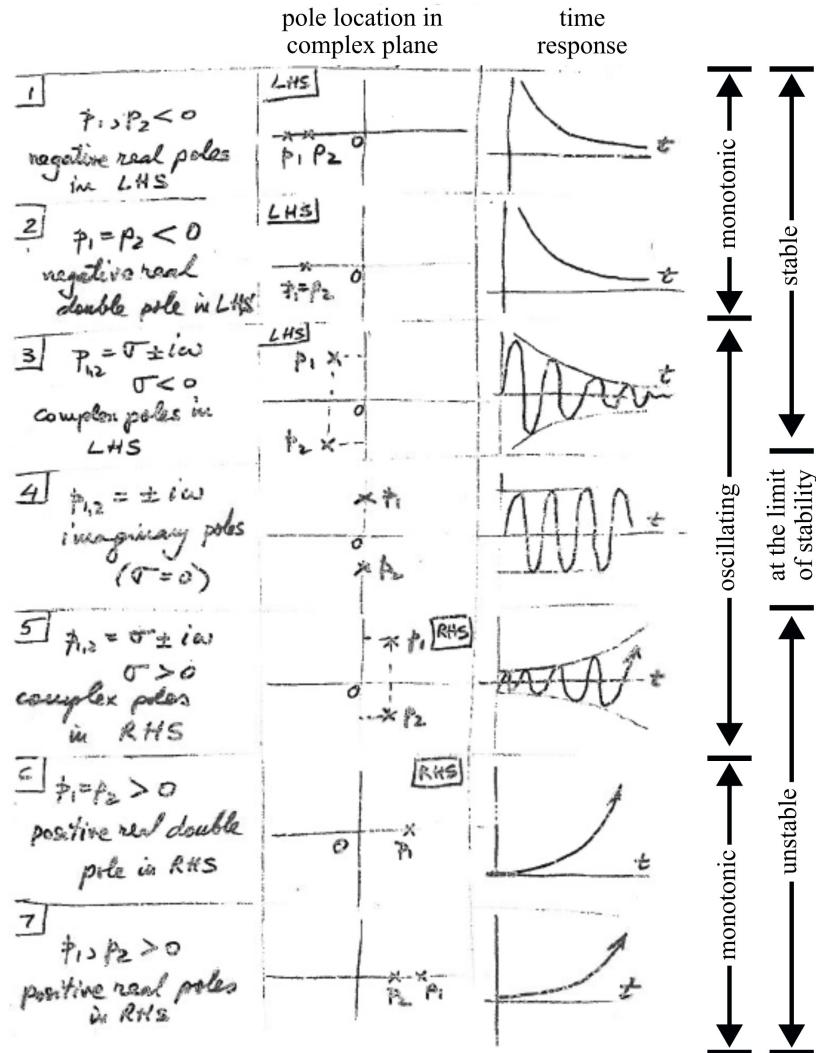
where  $p$  is the pole of  $X(s)$ .

The stability is dictated by the sign of  $p$ , i.e. its location in the complex  $p$ -plane. If  $P < 0$  (or  $p$  is in the left-hand-side), the system is stable. Therefore, if a disturbing force is applied, the system will return to its initial state.



#### 4.6.2 Stability of 2<sup>nd</sup>-Order Responses

$$X(s) = \frac{k(s - z_1)}{(s - p_1)(s - p_2)} \quad (334)$$



### 4.6.3 Stability of Higher-order Responses

Starting with a general expression for the output of a system

$$X(s) = \frac{B(s)}{A(s)} = \frac{b_m s^m + b_{m-1} s^{m-1} + \dots + b_0}{a_n s^n + a_{n-1} s^{n-1} + \dots + a_0} \quad (335)$$

partial fraction expansion results in

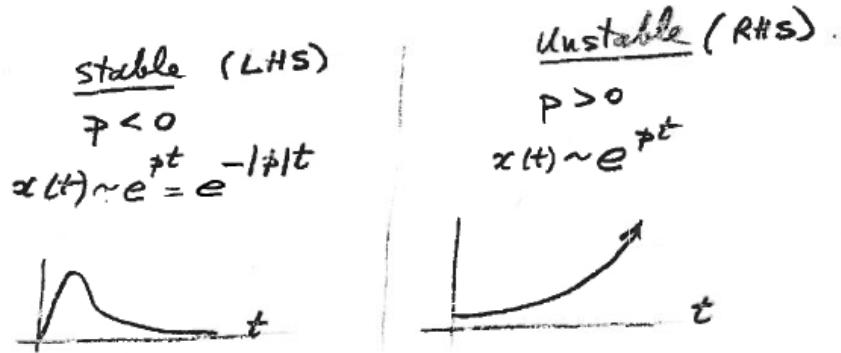
$$X(s) = \frac{r_1}{s - p_1} + \frac{r_2}{s - p_2} + \dots + \frac{r_n}{s - p_n} \quad (336)$$

where  $p_1, p_2, \dots, p_n$  are the poles of the system (i.e. roots of  $A(s) = 0$ ) and  $r_1, r_2, \dots, r_n$  are the residues of the system. Note that the poles can either be real or complex. Again, MATLAB can be used to solve for the roots, poles, and gains of the system using  $[r, p, k] = \text{residue}(B, A)$ . The real poles can be

- single pole:  $\frac{r}{s-p} \xrightarrow{\mathcal{L}[ ]^{-1}} r e^{pt}$

- double poles:  $\frac{r}{(s-p)^2} \xrightarrow{\mathcal{L}^{-1}} rte^{pt}$
- multiple poles:  $\frac{r}{(s-p)^j} \xrightarrow{\mathcal{L}^{-1}} \frac{1}{(j-1)!} t^{j-1} e^{pt}$

where stable and unstable responses are



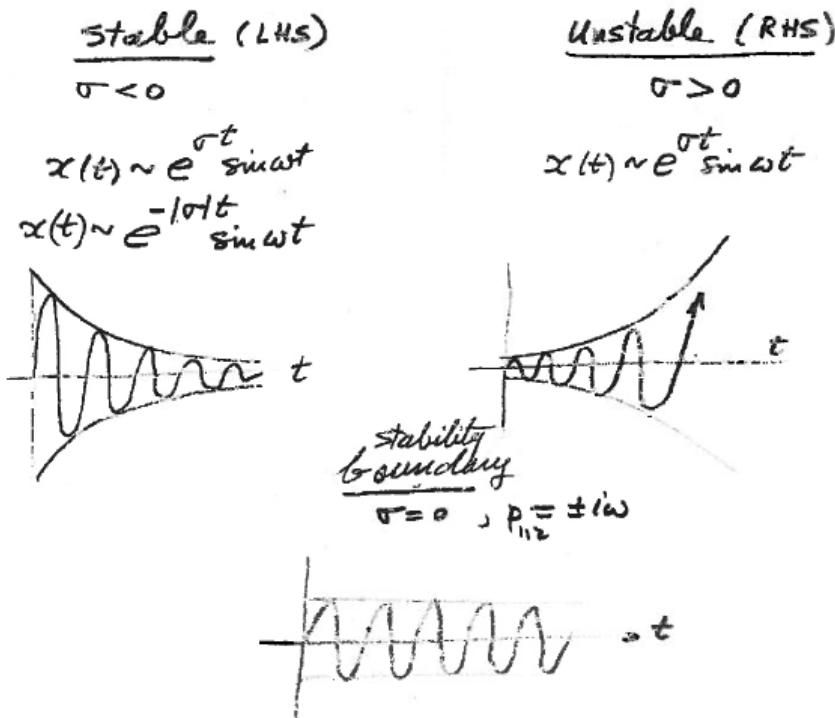
The complex poles are always in conjugate pairs, and are governed by  $p_{1,2} = \sigma \pm i\omega$  where

- first pole:  $\frac{1}{s-p_1} = \frac{1}{s-(\sigma+i\omega)} \xrightarrow{\mathcal{L}^{-1}} e^{(\sigma+i\omega)t} = e^{\sigma t} e^{i\omega t}$
- second pole:  $\frac{1}{s-p_2} = \frac{1}{s-(\sigma-i\omega)} \xrightarrow{\mathcal{L}^{-1}} e^{(\sigma-i\omega)t} = e^{\sigma t} e^{-i\omega t}$

Using Euler's formula, this results in

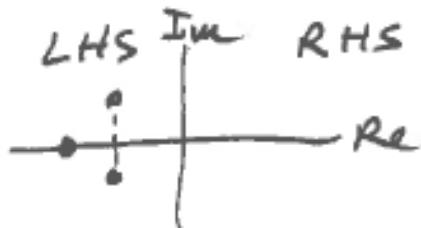
$$x(t) = e^{\sigma t} (e^{i\omega t} - e^{-i\omega t}) = e^{\sigma t} \sin(\omega t) \quad (337)$$

where stable and unstable responses are



#### 4.6.4 Absolute Stability

A necessary and sufficient condition for a system to be stable is that its poles are placed in the Left hand side of the complex plane.



#### 4.6.5 Marginal Stability

If the poles are purely imaginary (i.e. placed on the imaginary axis) then the system has marginal stability.

- bounded impulse response, the system is stable
- unbounded impulse response, or other inputs, the system is not stable.

#### 4.6.6 Relative Stability

Would a stable system still be stable if its parameters are slightly changed? What margin of safety is there?

## 5 Performance Indicators

Performance indicators are used to judge the quality of a control system.

### 5.1 1<sup>st</sup>-order System Generic Performance Indicators

Generic performance indicators are:

- steady-state value  $x_{ss} = \lim_{t \rightarrow \infty} x(t)$
- steady-state error  $e_{ss} = \lim_{t \rightarrow \infty} e(t)$  where  $e(t) = f(t) - x(t)$

To expand on the error definition,  $e(t) = f(t) - x(t)$ , consider a step response where  $x(\infty) = 1$  and  $f(\infty) = 1$ , therefore  $e(\infty) = 0$ .

For a 1<sup>st</sup>-order system,  $x_{ss}$  and  $e_{ss}$  can be found using the final value theorem and exist in both the time domain and the s-domain. Recall the final value theorem for  $x_{ss}$  and  $e_{ss}$  leads to

$$x_{ss} = \lim_{s \rightarrow 0} sX(s) \quad (338)$$

$$e_{ss} = \lim_{s \rightarrow 0} sE(s) \quad (339)$$

Therefore, starting at the transfer function of a 1<sup>st</sup>-order system

$$G(s) = \frac{1}{Ts + 1} \quad (340)$$

we can expand on this to show

$$\begin{aligned} X(s) &= G(s)F(s) \\ &= \frac{1}{Ts + 1}F(s) \end{aligned} \quad (341)$$

next, the error is s-domain is shown to be

$$E(s) = F(s) - X(s) \quad (342)$$

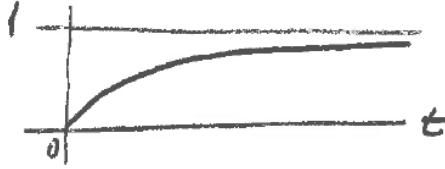
$$= F(s) - G(s)F(s) \quad (343)$$

$$= (1 - G(s))F(s)$$

$$= \frac{Ts}{Ts + 1}F(s)$$

Again, these are general terms for a 1<sup>st</sup>-order system.

### 5.1.1 Step response 1<sup>st</sup>-order system performance indicators



For a 1<sup>st</sup>-order system subjected to a step response, we want to find  $x_{ss}$  and  $e_{ss}$  and we know the transfer function is defined as

$$G(s) = \frac{1}{Ts + 1} \quad (344)$$

while the equation of motion is

$$T\dot{x} + x = F(s) \quad (345)$$

Where the step function is defined as

$$f(t) = 1, t > 0 \quad (346)$$

therefore, the system response is

$$x(t) = 1 - e^{-t/T} \quad (347)$$

while the steady-state response is

$$\begin{aligned} x_{ss} &= \lim_{t \rightarrow \infty} x(t) \\ &= 1 \end{aligned} \quad (348)$$

Next, the error as a function of time is

$$\begin{aligned} e(t) &= 1 - (1 - e^{-t/T}) \\ &= e^{-t/T} \end{aligned} \quad (349)$$

while the steady-state error is

$$\begin{aligned} e_{ss} &= \lim_{t \rightarrow \infty} e(t) \\ &= \lim_{t \rightarrow \infty} e^{t/T} \\ &= 0 \end{aligned} \quad (350)$$

These same solutions can be found in the s-domain where the step function is  $F(s) = \frac{1}{s}$ . Consider the s-domain expression solved for  $X(s)$ ,

$$X(s) = \frac{1}{Tx + 1} \cdot \frac{1}{s} \quad (351)$$

the steady-state error is shown to be

$$\begin{aligned} x_{ss} &= \lim_{s \rightarrow 0} s \frac{1}{Ts + 1} \cdot \frac{1}{s} \\ &= \lim_{s \rightarrow 0} \frac{1}{Ts + 1} \\ &= 1 \end{aligned} \quad (352)$$

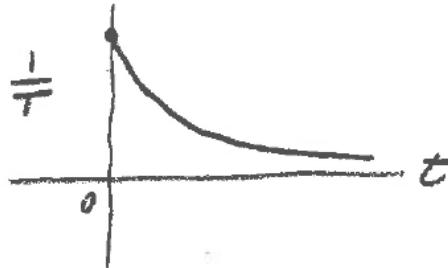
Next, we can build the s-domain representation of the error as

$$\begin{aligned} E(s) &= \frac{Ts}{Ts + 1} \cdot \frac{1}{s} \\ &= \frac{T}{Ts + 1} \end{aligned} \quad (353)$$

solving for the steady-state error results in

$$\begin{aligned} e_{ss} &= \lim_{s \rightarrow 0} s \frac{T}{Ts + 1} \\ &= 0 \end{aligned} \quad (354)$$

### 5.1.2 Impulse response 1<sup>st</sup>-order system performance indicators



For a 1<sup>st</sup>-order system subjected to an impulse response, we want to find  $x_{ss}$  and  $e_{ss}$ . The impulse function is defined as

$$f(t) = \delta(t) \quad (355)$$

therefore, the response is

$$x(t) = \frac{1}{T} e^{-t/T} \quad (356)$$

where the steady-state response is

$$x_{ss} = 0 \quad (357)$$

The error is

$$e(t) = \delta(t) - \frac{1}{T} e^{-t/T} \quad (358)$$

Lastly, the steady-state error is

$$\begin{aligned} e_{ss} &= \lim_{t \rightarrow \infty} e(t) \\ &= 0 \end{aligned} \quad (359)$$

These same solutions can be found in the s-domain where the impulse function is  $F(s) = 1$ . Consider the s-domain expression solved for  $X(s)$ ,

$$X(s) = \frac{1}{Ts + 1} \quad (360)$$

the steady-state error is shown to be

$$\begin{aligned} x_{ss} &= \lim_{s \rightarrow 0} s \frac{1}{Ts + 1} \\ &= 0 \end{aligned} \quad (361)$$

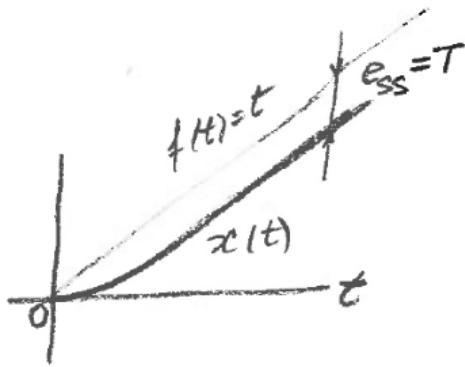
The s-domain error is expressed as

$$E(s) = \frac{T}{Ts + 1} \quad (362)$$

which leads to the steady-state error value

$$\begin{aligned} e_{ss} &= \lim_{s \rightarrow 0} s \frac{T}{Ts + 1} \\ &= 0 \end{aligned} \quad (363)$$

### 5.1.3 Ramp response 1<sup>st</sup>-order system performance indicators



For a 1<sup>st</sup>-order system subjected to a ramp response, we want to find  $x_{ss}$  and  $e_{ss}$ . The impulse function is defined as

$$f(t) = t \quad (364)$$

therefore, the response is

$$x(t) = t - T(1 - e^{-t/T}) \quad (365)$$

which leads to the steady-state response

$$x_{ss} = \infty \quad (366)$$

which means that there is no steady-state value. The error is

$$e(t) = T(1 - e^{-t/T}) \quad (367)$$

Lastly, the steady-state error is

$$\begin{aligned} e_{ss} &= \lim_{t \rightarrow \infty} e(t) \\ &= T - \lim_{t \rightarrow \infty} e^{-t/T} \\ &= T \end{aligned} \quad (368)$$

as  $\lim_{t \rightarrow \infty} e^{-t/T} = 0$ .

These same solutions can be found in the s-domain where the ramp function is  $F(s) = \frac{1}{s^2}$ . Consider the s-domain expression solved for  $X(s)$ ,

$$X(s) = \frac{1}{Ts + 1} \cdot \frac{1}{s^2} \quad (369)$$

the steady-state error is shown to be

$$\begin{aligned} x_{ss} &= \lim_{s \rightarrow 0} s \frac{1}{Ts + 1} \cdot \frac{1}{s^2} \\ &= \lim_{s \rightarrow 0} \frac{1}{Ts + 1} \cdot \frac{1}{s} \\ &= \infty \end{aligned} \quad (370)$$

therefore, there is no steady-state value. The s-domain error is expressed as

$$\begin{aligned} E(s) &= \frac{Ts}{Ts + 1} \cdot \frac{1}{s^2} \\ &= \frac{T}{Ts + 1} \cdot \frac{1}{s} \end{aligned} \quad (371)$$

which leads to the steady-state error value

$$\begin{aligned} e_{ss} &= \lim_{s \rightarrow 0} s \frac{T}{Ts + 1} \cdot \frac{1}{s} \\ &= \lim_{s \rightarrow 0} \frac{T}{Ts + 1} \\ &= T \end{aligned} \quad (372)$$

## 5.2 1<sup>st</sup>-order System Specific Performance Indicators

Specific performance indicators exist. They depend on system order and excitation type. Examples are:

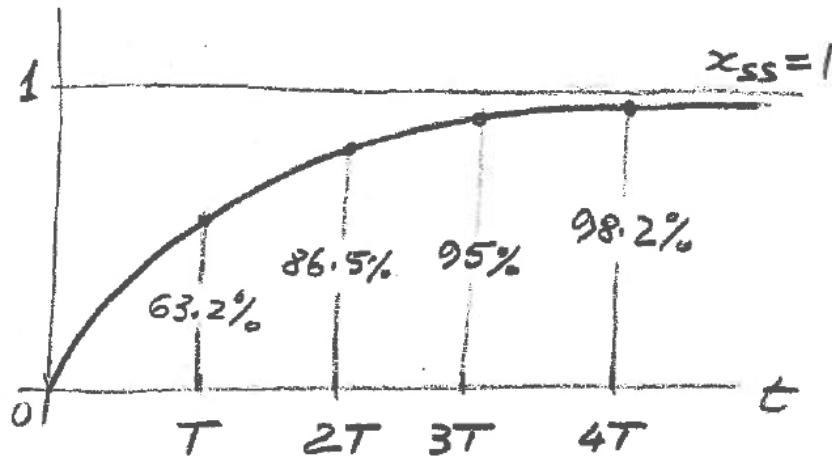
- rise-time  $\rightarrow t_r$
- delay time  $\rightarrow t_d$
- settling time  $\rightarrow t_s$

- decay time / half-time  $\rightarrow t_{1/2}$

For a step response, consider the system displacement

$$x(t) = 1 - e^{-t/T} \quad (373)$$

that is plotted as



The rise-time ( $t_r$ ) of a system

- to rise 63.2% of  $x_{ss}$  is  $t_{r,63.2\%} = T$
- to rise 86.5% of  $x_{ss}$  is  $t_{r,86.5\%} = 2T$
- to rise 95% of  $x_{ss}$  is  $t_{r,95\%} = 3T$
- to rise 98.2% of  $x_{ss}$  is  $t_{r,98.2\%} = 4T$

in general,  $1 - e^{-t/T} = x \rightarrow t = -T \ln(1 - x)$

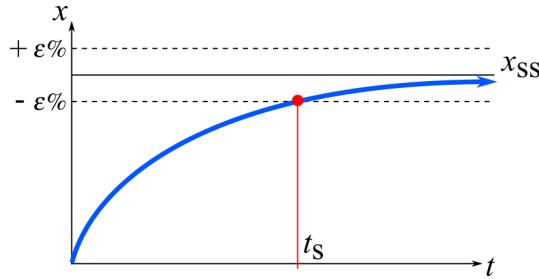
The delay time ( $t_d$ ) is the time it takes to rise to 50% of  $x_{ss}$ ,

$$\begin{aligned} x(t) &= 1 - e^{-t/T} \\ &= 0.5 \end{aligned} \quad (374)$$

therefore, solving for  $t = t_d$ ,

$$\begin{aligned} e^{-t_d/T} &= 0.5 \\ \frac{-t_d}{T} &= \ln(0.5) \\ t_d &= -T \ln(0.5) \\ t_d &= 0.693T \\ t_d &\approx 0.7T \end{aligned} \quad (375)$$

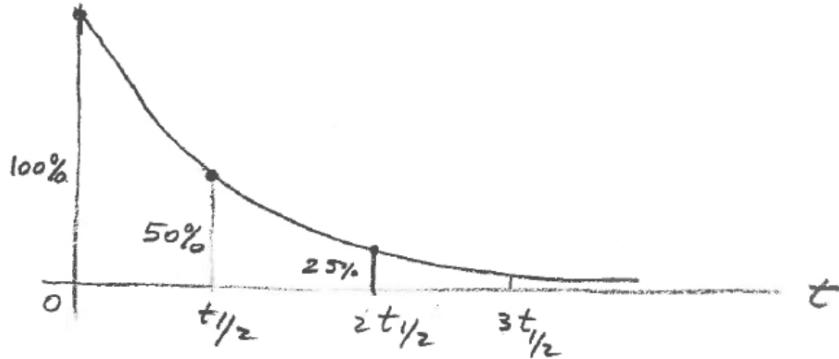
The settling time ( $t_s$ ) is the time it takes  $x$  to get within  $\epsilon\%$  of  $x_{ss}$ .



A more precise estimator is the rise time to any selected  $x$  which can be defined as

$$\begin{aligned} x_{ss}(1 - e^{-t/T}) &= x \\ e^{(-t/T)} &= 1 - \frac{x}{x_{ss}} \\ -t/T &= \log \left[ 1 - \frac{x}{x_{ss}} \right] \\ t &= -T \log \left[ 1 - \frac{x}{x_{ss}} \right] \end{aligned} \quad (376)$$

### 5.2.1 Impulse Response Specific Performance Indicators



The half-cycle decay time ( $t_{1/2}$ ) is an impulse response specific performance indicator. Given that the system response to an impulse is

$$x(t) = \frac{1}{T} e^{-t/T} \quad (377)$$

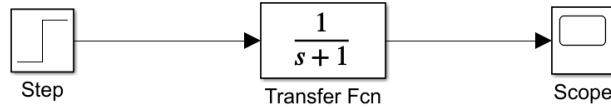
the half-life decay time is where  $e^{-t/T} = \frac{1}{2}$ , therefore

$$\begin{aligned} t_{1/2} &= -\ln\left(\frac{1}{2}\right)T \\ &\approx 0.693T \end{aligned} \quad (378)$$

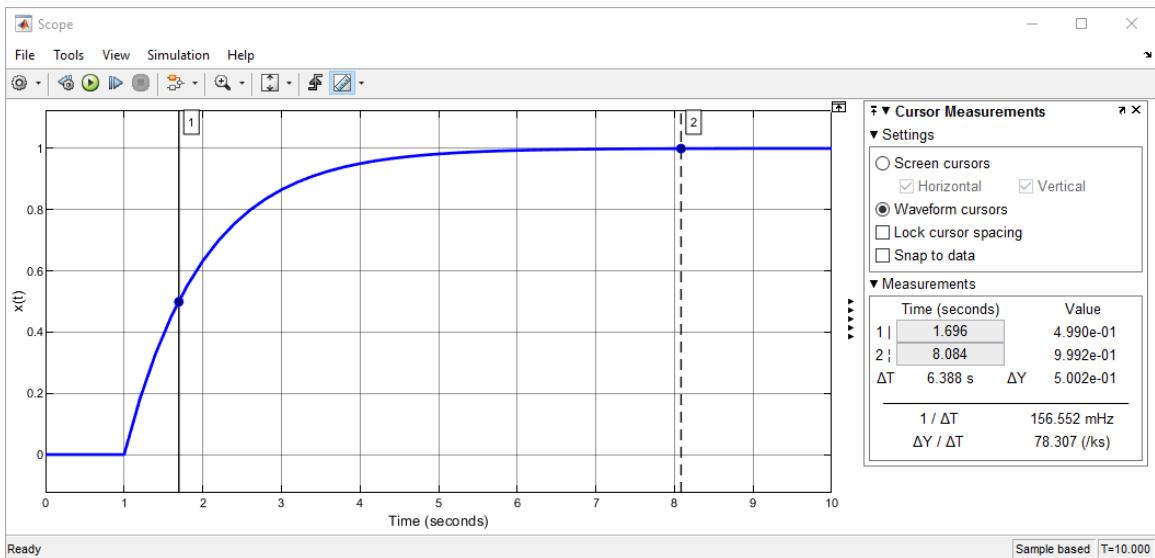
Note that the signal continues to decay by half of its value after each additional  $t_{1/2}$ .

**Example 5.1 SIMULINK Tutorial on Measuring Performance Indicators**

Build the simple 1<sup>st</sup>-order system as shown below



Open the scope and press the ‘Cursor Measurements’ button to activate the cursors.



To measure the delay time  $t_d$ , place the first cursor around the point where ‘Value’ measurement is closest to 0.5.  $x(t) = 0.5$ . Read the ‘Time’ value. This is estimate for  $t_d$ . It gives the value  $t_d = 0.70$  sec, as the step function happens at 1 sec.

The second cursor can be used to get the settling time  $t_s$ . We are going to use the 1% definition of  $t_d$ . This means that the response should be around 0.99, or 9.9e-1. Reading the corresponding time value, we get  $t_s \approx 7.0$  sec.

### 5.3 2<sup>nd</sup>-order System Generic Performance Indicators

Again, starting at the equation of motion

$$\ddot{x} + 2\zeta\omega_n\dot{x} + \omega_n^2x = \omega_n^2f(t) \quad (379)$$

and transfer function of a 2<sup>nd</sup>-order system

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (380)$$

we can expand on this to show

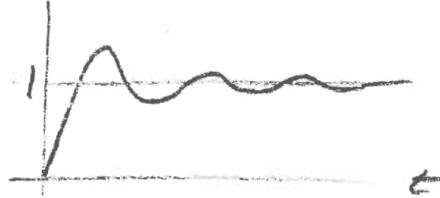
$$\begin{aligned} X(s) &= G(s)F(s) \\ &= \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} F(s) \end{aligned} \quad (381)$$

next, the error in the s-domain is shown to be

$$\begin{aligned} E(s) &= F(s) - X(s) \\ &= F(s) - G(s)F(s) \\ &= (1 - G(s))F(s) \\ &= \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} F(s) \end{aligned} \quad (382)$$

Again, these are general terms for a 2<sup>nd</sup>-order system. Note that the definitions for the steady-state value  $x_{ss}$  and steady-state error  $e_{ss}$  remain largely unchanged from the 1<sup>st</sup>-order system.

### 5.3.1 Step response 2<sup>nd</sup>-order system performance indicators



For a 2<sup>nd</sup>-order system subjected to a step response, we want to find  $x_{ss}$  and  $e_{ss}$ . Using the transfer function method to solve for the response, we know the system response is

$$x(t) = 1 - \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \quad (383)$$

where

$$\omega_d = \omega_n \sqrt{1 - \zeta^2} \quad (384)$$

and

$$\phi = \tan^{-1} \frac{\sqrt{1 - \zeta^2}}{\zeta} = \sin^{-1} \sqrt{1 - \zeta^2} \quad (385)$$

next, the steady-state displacement can be solved for as

$$\begin{aligned} x_{ss} &= \lim_{t \rightarrow \infty} x(t) \\ &= 1 - \frac{\sqrt{1 - \zeta^2}}{\zeta} \lim_{t \rightarrow \infty} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \\ &= 1 \end{aligned} \quad (386)$$

as  $e^{-\zeta\omega_n t}$  goes to 0. Next, the error as a function of time is

$$\begin{aligned} e(t) &= f(t) - x(t) \\ &= \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \end{aligned} \quad (387)$$

while the steady-state error is

$$\begin{aligned} e_{ss} &= \lim_{t \rightarrow \infty} e(t) \\ &= \frac{\sqrt{1-\zeta^2}}{\zeta} \lim_{t \rightarrow \infty} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \\ &= 0 \end{aligned} \quad (388)$$

as  $e^{-\zeta\omega_n t}$  goes to 0.

These same solutions can be found in the s-domain where the step function is  $F(s) = \frac{1}{s}$ . Consider the s-domain expression solved for  $X(s)$ ,

$$X(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s} \quad (389)$$

the steady-state displacement is shown to be

$$\begin{aligned} x_{ss} &= \lim_{s \rightarrow 0} s \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s} \\ &= \frac{\omega_n^2}{\omega_n^2} \\ &= 1 \end{aligned} \quad (390)$$

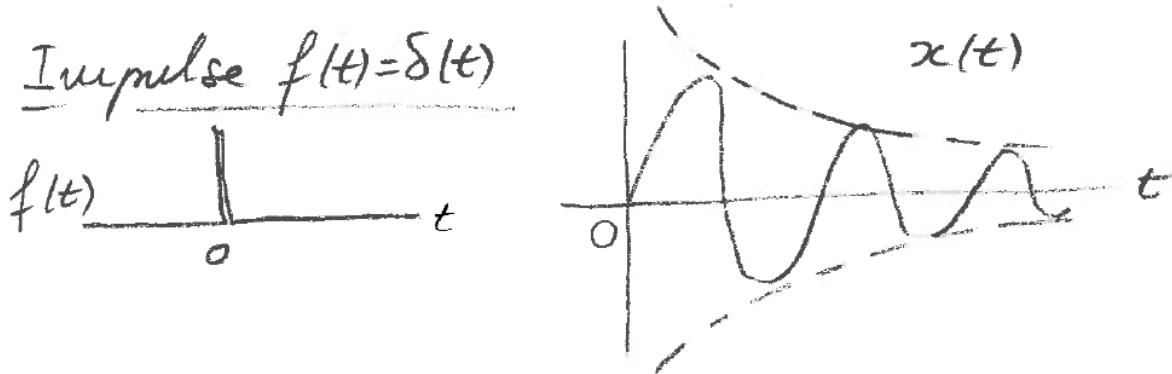
Next, we can build the s-domain representation of the error as

$$E(s) = \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s} \quad (391)$$

solving for the steady-state error results in

$$\begin{aligned} e_{ss} &= \lim_{s \rightarrow 0} s \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s} \\ &= \frac{0}{\omega_n} \\ &= 0 \end{aligned} \quad (392)$$

### 5.3.2 Impulse response 2<sup>nd</sup>-order system performance indicators



For a 2<sup>nd</sup>-order system subjected to an impulse response, we want to find \$x\_{ss}\$ and \$e\_{ss}\$. The impulse function is defined as

$$f(t) = \delta(t) \quad (393)$$

therefore, the response is

$$x(t) = \frac{\omega_n}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t) \quad (394)$$

where the steady-state response is

$$\begin{aligned} x_{ss} &= \lim_{t \rightarrow \infty} x(t) \\ &= \frac{\omega_n}{\sqrt{1-\zeta^2}} \lim_{t \rightarrow \infty} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \\ &= 0 \end{aligned} \quad (395)$$

as \$e^{-\zeta\omega\_n t}\$ goes to 0. Next, the error as a function of time is

$$\begin{aligned} e(t) &= f(t) - x(t) \\ &= \delta(t) - \frac{\omega_n}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \end{aligned} \quad (396)$$

Lastly, the steady-state error is

$$\begin{aligned} e_{ss} &= \lim_{t \rightarrow \infty} e(t) \\ &= \lim_{t \rightarrow \infty} \delta(t) - \frac{\omega_n}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \\ &= 0 \end{aligned} \quad (397)$$

as \$\delta(t)\$ and \$e^{-\zeta\omega\_n t}\$ both go to 0.

These same solutions can be found in the s-domain where the impulse function is \$F(s) = 1\$. Consider the s-domain expression solved for \$X(s)\$,

$$X(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (398)$$

the steady-state error is shown to be

$$x_{ss} = \lim_{s \rightarrow 0} s \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} = 0 \quad (399)$$

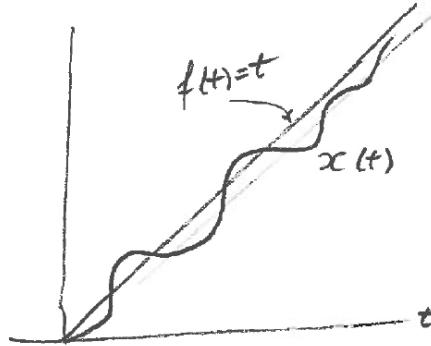
The s-domain error is expressed as

$$E(s) = \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (400)$$

which leads to the steady-state error value

$$e_{ss} = \lim_{s \rightarrow 0} s \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} = 0 \quad (401)$$

### 5.3.3 Ramp response 2<sup>nd</sup>-order system performance indicators



For a 2<sup>nd</sup>-order system subjected to a ramp response, we want to find  $x_{ss}$  and  $e_{ss}$ . The impulse function is defined as

$$f(t) = t \quad (402)$$

therefore, the response can be found through the transfer function approach as

$$x(t) = t - \frac{2\zeta}{\omega_n} \left[ 1 + \frac{1}{\sin \phi_1} e^{-\zeta \omega_n t} \sin(\omega_d t + \phi_1) \right] \quad (403)$$

where

$$\phi_1 = \tan^{-1} \frac{2\zeta \sqrt{1 - \zeta^2}}{1 - 2\zeta^2} \quad (404)$$

which leads to the response

$$\begin{aligned} x_{ss} &= \lim_{t \rightarrow \infty} x(t) \\ &= t - \frac{2\zeta}{\omega_n} \\ &= \infty \end{aligned} \quad (405)$$

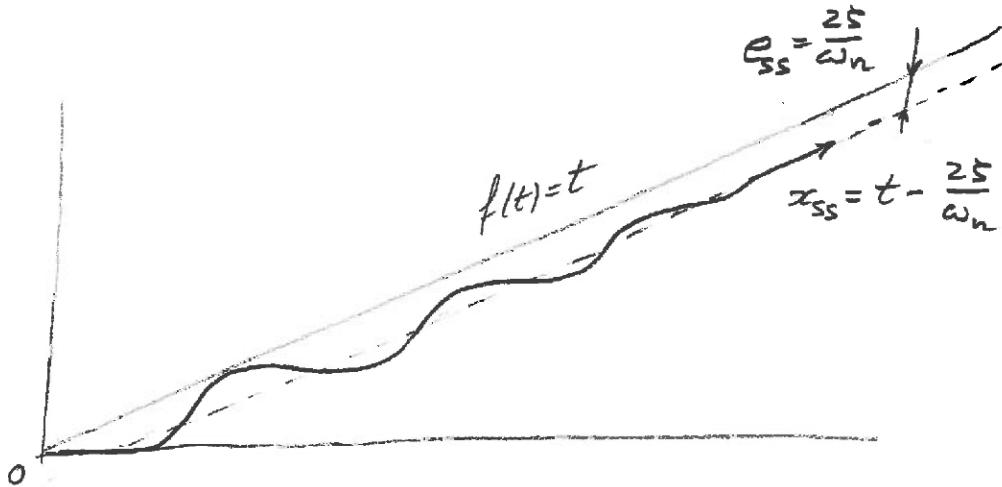
which means that there is no steady-state value. Next, the error as a function of time is

$$\begin{aligned} e(t) &= f(t) - x(t) \\ &= t - \left[ t - \frac{2\zeta}{\omega_n} \left[ 1 + \frac{1}{\sin \phi_1} e^{-\zeta \omega_n t} \sin(\omega_d t + \phi_1) \right] \right] \\ &= \frac{2\zeta}{\omega_n} \left[ 1 + \frac{1}{\sin \phi_1} e^{-\zeta \omega_n t} \sin(\omega_d t + \phi_1) \right] \end{aligned} \quad (406)$$

Lastly, the steady-state error is

$$\begin{aligned} e_{ss} &= \lim_{t \rightarrow \infty} e(t) \\ &= \frac{2\zeta}{\omega_n} \lim_{t \rightarrow \infty} \left[ 1 + \frac{1}{\sin \phi_1} e^{-\zeta \omega_n t} \sin(\omega_d t + \phi_1) \right] \\ &= \frac{2\zeta}{\omega_n} \end{aligned} \quad (407)$$

as the term in the brackets goes to 1.



These same solutions can be found in the s-domain where the ramp function is  $F(s) = \frac{1}{s^2}$ . Consider the s-domain expression solved for  $X(s)$ ,

$$X(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s^2} \quad (408)$$

the steady-state error is shown to be

$$\begin{aligned}
 x_{ss} &= \lim_{s \rightarrow 0} s \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s^2} \\
 &= \lim_{s \rightarrow 0} \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s} \\
 &= \lim_{s \rightarrow 0} \frac{1}{s} \\
 &= \infty
 \end{aligned} \tag{409}$$

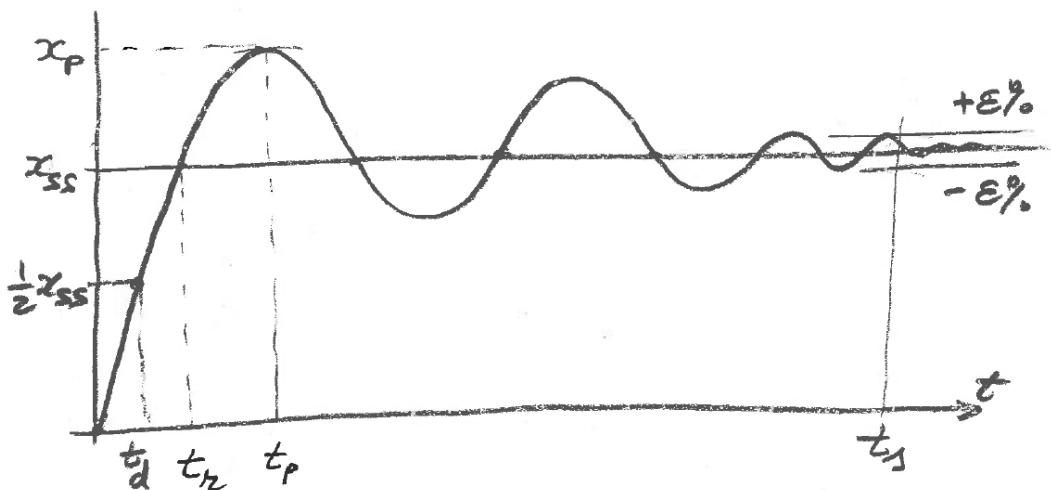
therefore, there is no steady-state value. The s-domain error is expressed as

$$E(s) = \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s^2} \tag{410}$$

which leads to the steady-state error value

$$\begin{aligned}
 e_{ss} &= \lim_{s \rightarrow 0} s \frac{s^2 + 2\zeta\omega_n s}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s^2} \\
 &= \lim_{s \rightarrow 0} \frac{s^2(s + 2\zeta\omega_n)}{s^2 + 2\zeta\omega_n s + \omega_n^2} \cdot \frac{1}{s^2} \\
 &= \frac{2\zeta\omega_n}{\omega_n^2} \\
 &= \frac{2\zeta}{\omega_n}
 \end{aligned} \tag{411}$$

## 5.4 2<sup>nd</sup>-order System Specific Performance Indicators



Specific performance indicators exist. They depend on system order and excitation type. Examples are:

- rise time  $\rightarrow t_r$
- peak time  $\rightarrow t_p$
- peak value  $\rightarrow x_p$
- settling time  $\rightarrow t_s$
- delay time  $\rightarrow t_d$
- max percentage overshoot  $\rightarrow M_p$

Many of these are the same as a 1<sup>st</sup>-order system or taken directly from the system response. The max percentage overshoot ( $M_p$ ) is defined as

$$M_p = \left( \frac{x_p}{x_{ss}} - 1 \right) \cdot 100 \quad (412)$$

### 5.4.1 Procedure for a Step Response

Given the 2<sup>nd</sup>-order system, with the transfer function

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (413)$$

where  $\omega_n$  is the natural frequency and  $\zeta$  is the critical damping ratio. Considering that the system is subjected to a step response, we know the system response is

$$x(t) = 1 - \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \quad (414)$$

where we can calculate:

$$\omega_d = \omega_n \sqrt{1 - \zeta^2} \quad (415)$$

and

$$\phi = \tan^{-1} \frac{\sqrt{1 - \zeta^2}}{\zeta} = \sin^{-1} \sqrt{1 - \zeta^2} \quad (416)$$

Next, compute the values for the performance indicators.

First the rise time is calculated by the fact that  $x(0) = 0$  and  $x(t_r) = 1$ . From equation 414, the mean height of the system is shown to be when  $\sin(\omega_d t + \phi) = 0$ , therefore, as  $\sin(\pi) = 0$ , we can show that  $\omega_d t + \phi = \pi$ , rearranging this yields

$$t_r = \frac{\pi - \phi}{\omega_d} \quad (417)$$

To find the peak time ( $t_p$ ), we need to find the peak, which is defined as  $\frac{dx}{dt}|_{t=t_p} = 0$ , or drawn as



where we consider only the decaying part of the signal caused by the step function, or  $x = e^{-\zeta \omega_n t} \sin(\omega_d t + \phi)$ . Therefore

$$\begin{aligned}
 0 &= \frac{dx}{dt} \\
 &= \frac{d}{dt} \left[ e^{-\zeta \omega_n t} \sin(\omega_d t + \phi) \right] \\
 &= -\zeta \omega_n e^{-\zeta \omega_n t} \sin(\omega_d t + \phi) + \omega_d e^{-\zeta \omega_n t} \cos(\omega_d t + \phi) \\
 \zeta \omega_n \sin(\omega_d t + \phi) &= \omega_d \cos(\omega_d t + \phi) \\
 \frac{\sin(\omega_d t + \phi)}{\cos(\omega_d t + \phi)} &= \frac{\omega_d}{\zeta \omega_n}
 \end{aligned} \tag{418}$$

By converting the left hand side of this equation to  $\tan(\omega_d t_p + \phi)$  yields

$$\begin{aligned}
 \tan(\omega_d t_p + \phi) &= \frac{\omega_d}{\zeta \omega_n} \\
 &= \frac{\sqrt{1 - \zeta^2}}{\zeta} \\
 &= \tan(\phi)
 \end{aligned} \tag{419}$$

when considering the definition of tan provided by equation 416. We must find  $t_p$  such that  $\tan(\omega_d t_p + \phi) = \tan(\phi)$ . Given that the function  $\tan(\alpha)$  repeats itself after  $\pi$ ,  $2\pi$ ,  $3\pi$ , ... as shown in figure 5.1, we know  $\tan(\alpha) = \tan(\alpha + \pi) = \tan(\alpha + 2\pi), \dots$

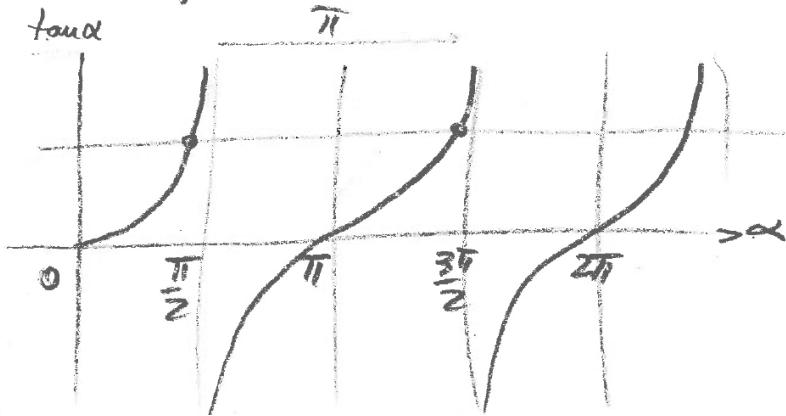


Figure 5.1: Plotting the  $\tan(\alpha)$

Hence, picking back up from equation 419 lead to

$$\begin{aligned}
 \omega_d t_p + \phi &= \phi + \pi \\
 \omega_d t_p &= \pi
 \end{aligned} \tag{420}$$

which results in

$$t_p = \frac{\pi}{\omega_d} \quad (421)$$

The peak value ( $x_p$ ) can be found as

$$\begin{aligned} x_p &= x(t_p) \\ &= 1 - \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta \omega_n \frac{\pi}{\omega_d}} \sin(\omega_d(\pi/\omega_d) + \phi) \\ &= 1 + \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta \omega_n \frac{\pi}{\omega_d}} \sin(\phi) \\ &= 1 + \frac{1}{\sqrt{1-\zeta^2}} \left( e^{-\frac{\zeta}{\sqrt{1-\zeta^2}} \pi} \right) \sqrt{1-\zeta^2} \\ &= 1 + e^{-\frac{\zeta}{\sqrt{1-\zeta^2}} \pi} \end{aligned} \quad (422)$$

when considering that  $\sin(\phi + \pi) = -\sin(\phi)$  and  $\sin(\phi) = \sqrt{(1-\zeta^2)}$ .

The max overshoot ( $M_p$ ) can be found as

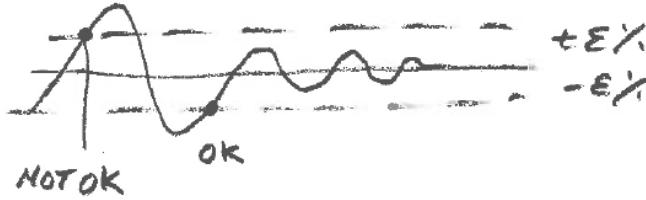
$$\begin{aligned} M_p &= \frac{x_p - x_{ss}}{x_{ss}} \\ &= \frac{1 + e^{-\frac{\zeta}{\sqrt{1-\zeta^2}} \pi} - 1}{1} \\ &= e^{-\frac{\zeta}{\sqrt{1-\zeta^2}} \pi} \end{aligned} \quad (423)$$

for a few select damping ratios, the overshoot percentages are shown in Table 3. However, the typical range of damping is  $0.4 < \zeta < 0.8$  and therefore the typical max overshoot is  $0.25\% < M_p < 1.5\%$ .

Table 3: Overshoot percentages for select damping ratios.

$\zeta$	0	0.2	0.4	0.6	0.8	1
$M_p$	100.00%	52.68%	25.40%	9.49%	1.52%	0.00%

The definition of settling time ( $t_s$ ) is to “get withing  $\pm\epsilon\%$  of  $x_{ss}$  and stay so”.



The settling time is defined as the time when both

$$|x_{ss} - x(t_s)| < \Delta \quad (424)$$

and

$$|x_{ss} - x(t > t_s)| < \Delta \quad (425)$$

are true where  $\Delta = \varepsilon \cdot x_{ss}$ . This is shown in figure 5.2.

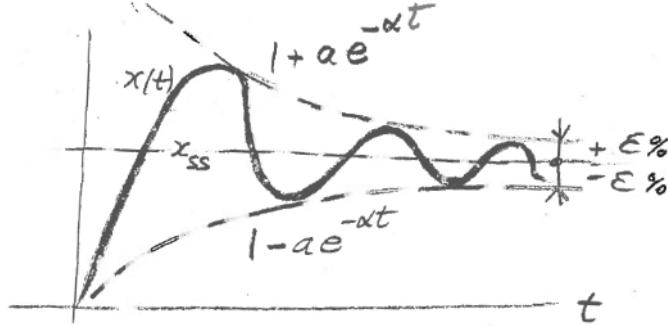


Figure 5.2: Settling time for a 2<sup>nd</sup>-order system.

We can find a generalized expression for  $t_s$  if we consider the system response for  $x_{ss} = 1$  as

$$x(t) = 1 - \left[ \frac{1}{\sqrt{1 - \zeta^2}} \right] e^{-[\zeta \omega_n]t} \sin(\omega_d t + \phi) \quad (426)$$

redefining the items in the brackets as  $a$  and  $\alpha$ , respectively, the system response can be simplified to

$$x(t) = 1 - ae^{-\alpha t} \sin(\omega_d t + \phi) \quad (427)$$

Therefore, the envelop of the system response is  $1 \pm ae^{-\alpha t}$  while the settling condition is  $ae^{-\alpha t} = \Delta$ . Considering that the final peak above the error range will happen when  $a \approx 1$ , we make the approximate calculation

$$a = \frac{1}{\sqrt{1 - \zeta^2}} \Big|_{\zeta \ll 1} \approx 1 \quad (428)$$

and considering that  $t_s$  is when the system decays under the value  $\varepsilon$ , we need to find the  $t$  value when

$$e^{-\alpha t} = \varepsilon \quad (429)$$

therefore, setting  $\varepsilon = 2\%$ , we can find

$$e^{-\alpha t} = 0.02 \quad (430)$$

$$-\alpha t = \log(0.02) \quad (431)$$

$$= -3.9$$

$$\approx -4$$

Therefore, knowing that  $\alpha = \zeta \omega_n$ , we can deduce

$$-\alpha t_s \approx -4 \quad (432)$$

$$-\zeta \omega_n t_s \approx -4$$

$$t_s \approx \frac{4}{\zeta \omega_n}$$

when  $\zeta \ll 1$  this simplified expression is within  $\pm 2\%$ .

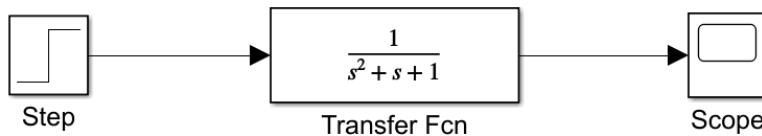
Importantly, one should always consider the effects of  $\zeta$  and  $\omega_n$  on performance.

- An increase in  $\omega_n$  shortens rise time ( $t_r$ ), peak time ( $t_p$ ), and settling time ( $t_s$ ).
- An increase in  $\zeta$  reduces max overshoot ( $M_p$ ) and shortens settling time ( $t_s$ ).

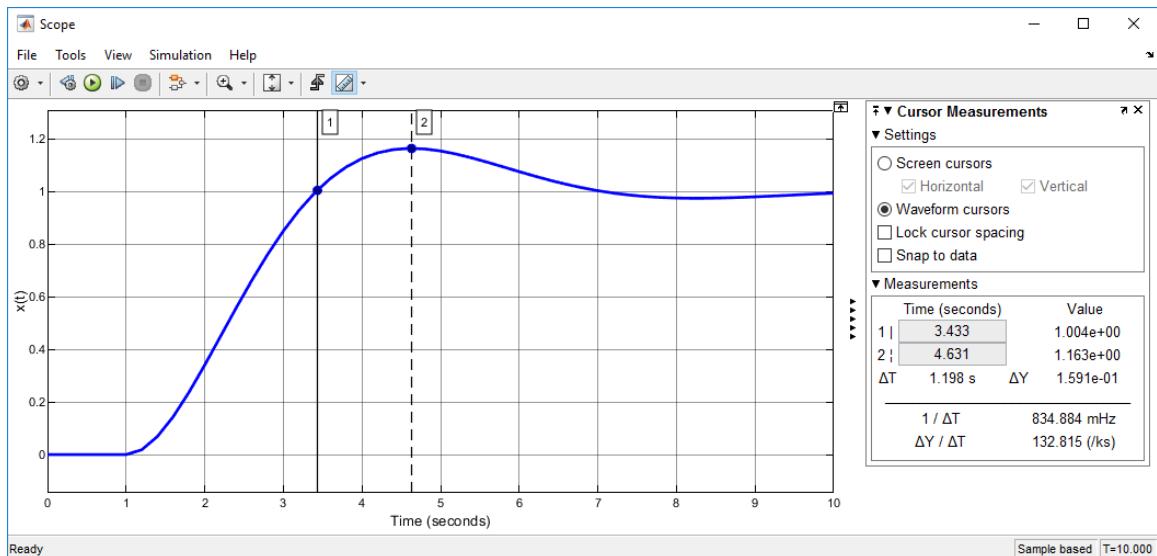
As before, the definition of time delay ( $t_d$ ) for a 2<sup>nd</sup>-order system is the time it takes to rise to 50% of  $x_{ss}$  the first time. Similarly, the max percentage overshoot ( $M_p$ ) for a 2<sup>nd</sup>-order system is calculated in the same way as a 1<sup>st</sup>-order system.

### Example 5.2 SIMULINK Tutorial on Measuring Performance Indicators

Build the simple 2<sup>nd</sup>-order system as shown below



Open the scope and press the ‘Cursor Measurements’ button to activate the cursors.



Use the first cursor to find the first crossing of  $x_{ss} = 1$ . Read the time as  $t_r = 2.433$  sec as the step function starts at 1 sec. Place the second cursor at peak value. Read the peak time  $t_p = 3.631$  sec and peak amplitude  $x_p = 1.163$ . Calculate  $M_p = 16.3\%$ .

## 6 System Identification

Given an experimental signal, the task is to find system parameters.

**Review 6.1** Harry Nyquist (February 7, 1889 ? April 4, 1976) was a Swedish physicist and electronic engineer. His parents emigrated to the U.S. in 1907. He attended the University of North Dakota starting in 1912 where he obtained a B.S. in 1914 and a M.S. in 1915, both in electrical engineering (entry to M.S. was 3 years!). Thereafter, he went to Yale University where he received a Ph.D. in physics in 1917.



Figure 6.1: Picture of Harry Nyquist from the American Institute of Physics. Fair use, via [Wikimedia Commons](#)

### 6.1 Experimental Signal Processing

to be updated based on Open Vibrations

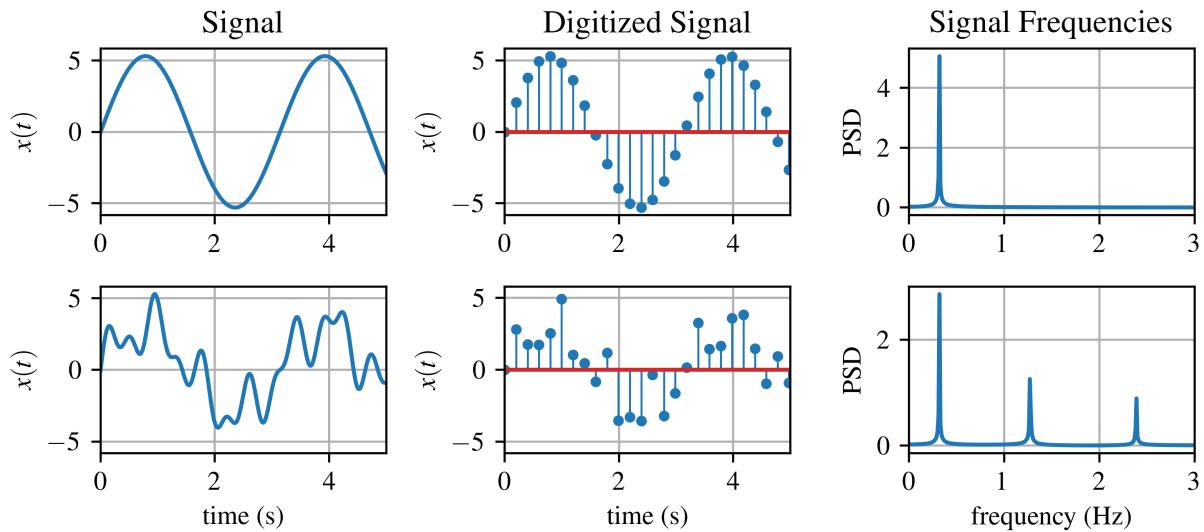


Figure 6.2: Digitization of two continuous time-series signals sampled at 5 S/s.

The Nyquist-Shannon sampling theorem is a theorem in the field of signal processing that defines the sample rate that permits a discrete sequence of samples (i.e. discrete-time) to sample a continuous-time signal of a finite bandwidth.

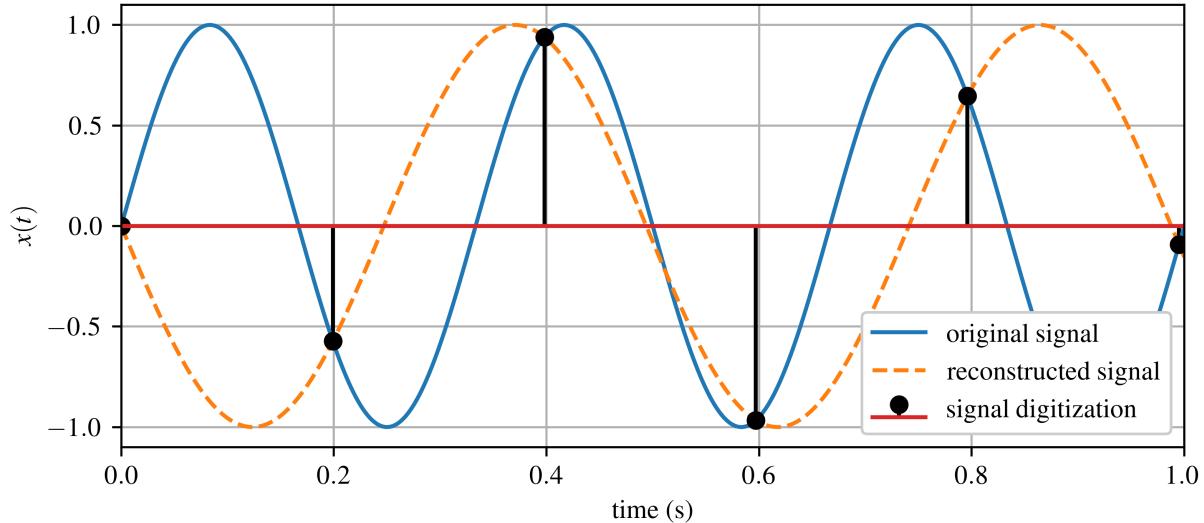
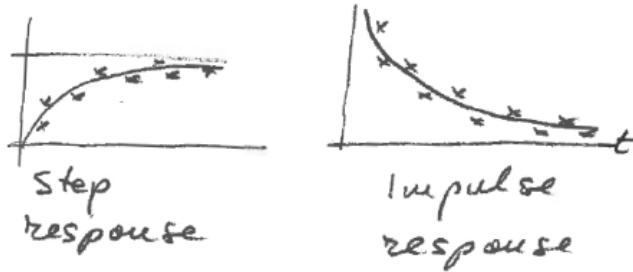


Figure 6.3: Aliasing of a 3 Hz signal that is sampled at 5 S/s.

In signal processing, aliasing is an effect that causes different signals to become indistinguishable from each other. In this way, the signals become aliases of one another when sampled. Aliasing also accounts for the development of distortion or artifact in a reconstructed signal when compared to the original continuous signal.

## 6.2 System Identification for 1<sup>st</sup>-order systems



Given a 1<sup>st</sup>-order system subjected to either a step or impulse input, there is a need to find the time constant ( $T$ ) for the first-order transfer function

$$G(s) = \frac{1}{Ts + 1} \quad (433)$$

### 6.2.1 Option 1 Optimization

Here, the task is to fit a curve to the experimental data. To do this, minimize the error in the function

$$x(t; T) = 1 - e^{-t/T} \quad (434)$$

for a series of experiments

$$\begin{aligned} x_{\text{exp}} &= x_1, x_2, x_3, \dots \\ t_{\text{exp}} &= t_1, t_2, t_3, \dots \end{aligned} \quad (435)$$

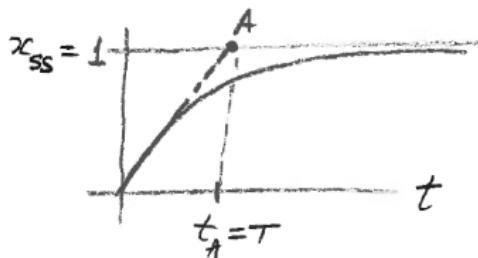
using any number of available curve fitting software to find  $T$ .

add example  
with curve  
fitting in  
Matlab

### 6.2.2 Option 2 Graphical Methods

In this method graphical methods can be used to provide quick estimates of the system parameter.

- **Tangent at Origin** This method shows us that  $T = t_A$ , as shown in the figure



$$x(t) = 1 - e^{-t/T} \quad (436)$$

and

$$\dot{x} = \frac{dx}{dt} = \left( -\frac{1}{T} \right) \left( -e^{-t/T} \right) = \frac{1}{T} e^{-t/T} \quad (437)$$

therefore

$$\dot{x}_0 = \frac{dx}{dt} \Big|_{t=0} = \frac{1}{T} \quad (438)$$

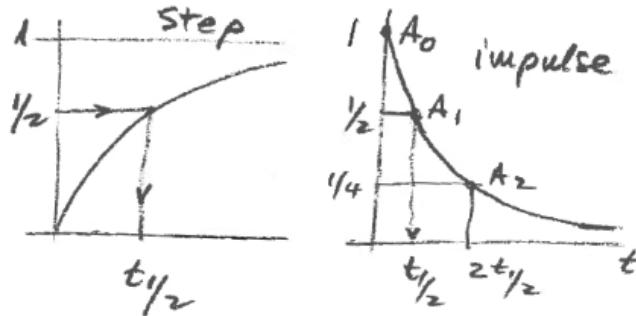
therefore, the tangent at the origin is

$$y(t) = \dot{x}_0 t = \frac{1}{T} t \quad (439)$$

where  $y(t)$  intersects  $x_{ss} = 1$  at  $\frac{1}{T}t_A = 1$ ; therefore

$$T = t_A \quad (440)$$

- **Half Time** Thought measuring the half time of a 1<sup>st</sup>-order systems subjected to either a step or impulse input, the time constant can be obtained.



- For a step function

$$x(t) = 1 - e^{-t/T} \quad (441)$$

solving for  $t = t_{1/2}$  leads to

$$\begin{aligned} x(t_{1/2}) &= 1/2 \\ 1/2 &= 1 - e^{-t_{1/2}/T} \\ &= e^{-t_{1/2}/T} \end{aligned} \quad (442)$$

solving for  $T$  results in

$$\begin{aligned} \frac{-t_{1/2}}{T} &= \ln(1/2) \\ T &= \frac{-t_{1/2}}{\ln(1/2)} \\ &= \frac{t_{1/2}}{0.693} \\ &\approx 1.4t_{1/2} \end{aligned} \quad (443)$$

- For a impulse function

$$x(t) = \frac{1}{T} e^{-t/T} \quad (444)$$

where

$$\begin{aligned} x(t_{1/2}) &= 1/2 \\ 1/2 &= \frac{1}{T} e^{-t_{1/2}/T} \end{aligned} \quad (445)$$

therefore

$$\begin{aligned} T &= \frac{-t_{1/2}}{\ln(1/2)} \\ T &\approx 1.4t_{1/2} \end{aligned}$$

Note that you may use the consecutive points  $A_1, A_2$  if  $A_0$  is not easy to determine, important is that signal halves between  $A_1$  and  $A_2$ .

This equation is correct, but I don't have a proof for it.

### Proof

For the expression

$$\begin{aligned} x(t_{1/2}) &= 1/2 \\ 1/2 &= 1 - e^{-t_{1/2}/T} \\ &= e^{-t_{1/2}/T} \end{aligned} \quad (446)$$

we can show the rest of this is only true for  $T=1$ , need a more robust proof.

$$T = -\frac{t_{1/2}}{W_0\left(-\frac{t_{1/2}}{2}\right)} \quad (447)$$

where  $W_0$  is the upper branch of the Lambert Function and  $W_0\left(-\frac{t_{1/2}}{2}\right) \neq 0, t_{1/2} \neq 0$  and  $n \in \mathbb{Z}$

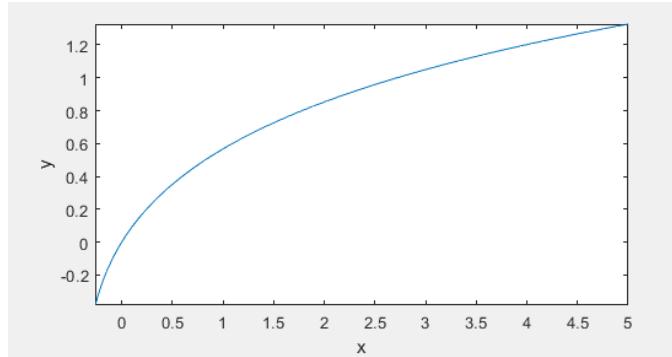


Figure 6.4: The upper branch of the Lambert Function graphed for  $y = W_0(x)$  for real components.

### 6.2.3 Option 3 Performance Indicators

Two methods can be used to estimate the time period:

- **Delay time** ( $t_d$ ) can be used to estimate the time period, as

$$t_d = -T \ln(0.5) \quad (448)$$

and therefore

$$T = -\frac{t_d}{\ln(0.5)} \quad (449)$$

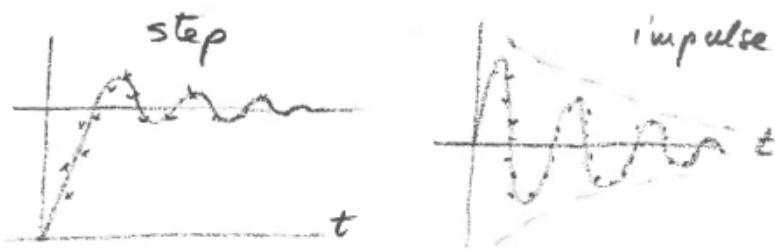
- **Settling time**  $t_s$  can be used to estimate the time period, here consider the 2% settling time value

$$t_s^{2\%} = -T \ln(0.02) \quad (450)$$

and therefore

$$T = -\frac{t_s^{2\%}}{\ln(0.02)} \quad (451)$$

## 6.3 System Identification for 2<sup>nd</sup>-order systems



Given a 2<sup>nd</sup>-order system subjected to either a step or impulse input, there is a need to find the natural frequency ( $\omega_n$ ) and damping ratio ( $\zeta$ ) for the 2<sup>nd</sup>-order transfer function

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (452)$$

### 6.3.1 Optimization

Here, the task is to fit a curve to the experimental data. To do this, minimize the error in the time series response of the system for a series of experiments

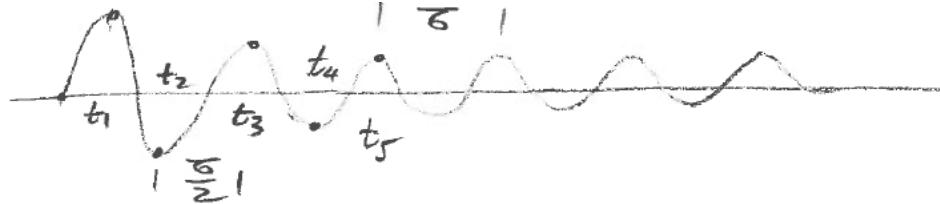
$$\begin{aligned} x_{\text{exp}} &= x_1, x_2, x_3, \dots \\ t_{\text{exp}} &= t_1, t_2, t_3, \dots \end{aligned} \quad (453)$$

using any number of available curve fitting software to find the natural frequency ( $\omega_n$ ) and damping ratio ( $\zeta$ ). add Matlab example, maybe with beam.

### 6.3.2 Frequency Estimation

Can be performed by finding the frequency at which the damped system crosses a given point, termed at  $f_d$ . This is either

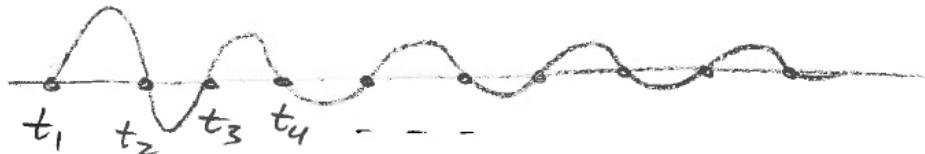
- Peak Detection



Is done by taking the average of the half period

$$f_d = \frac{\tau}{2} = \text{avg}[(t_2 - t_1), (t_2 - t_1), \dots] \quad (454)$$

- Zero Crossing



Is done by taking the average of the half period

$$f_d = \frac{\tau}{2} = \text{avg}[(t_2 - t_1), (t_2 - t_1), \dots] \quad (455)$$

In either case,

$$\omega_d = 2\pi f_d \quad (456)$$

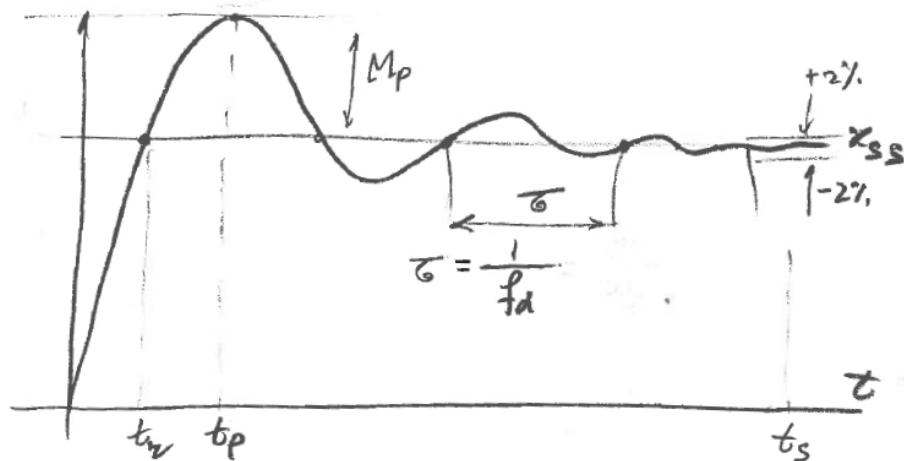
where

$$f_n = \frac{f_d}{\sqrt{1 - \zeta^2}} \quad (457)$$

and

$$\omega_n = 2\pi f_n \quad (458)$$

### 6.3.3 Step Response Analysis ( $\zeta \ll 1$ )



Recall that the rise time is

$$t_r = \frac{\pi - \phi}{\omega_n \sqrt{1 - \zeta^2}} = \frac{\pi - \phi}{\omega_d} \quad (459)$$

with

$$\omega_d = \omega_n \sqrt{1 - \zeta^2} \quad (460)$$

and

$$\phi = \tan^{-1} \frac{\sqrt{1 - \zeta^2}}{\zeta} = \sin^{-1} \sqrt{1 - \zeta^2} \quad (461)$$

while peak time is

$$t_p = \frac{\pi}{\omega_n \sqrt{1 - \zeta^2}} = \frac{\pi}{\omega_d} \quad (462)$$

The max percentage overshoot is

$$M_p = e^{-\frac{\zeta}{\sqrt{1-\zeta^2}}\pi} = \frac{x_p - x_{ss}}{x_{ss}} 100\% \quad (463)$$

and lastly, settling time is

$$t_s \approx \frac{4}{\zeta \omega_n} \quad (464)$$

From this, there are only two unknowns,  $\omega_n$  and  $\zeta$ . There is more information than minimally required to obtain these from  $t_r$ ,  $t_p$ ,  $M_p$ , and  $t_s$ .

- Obtain  $\omega_n$  and  $\zeta$  from  $t_r$  and  $t_p$ .

$$t_p = \frac{\pi}{\omega_d} \rightarrow \omega_d = \frac{\pi}{t_p} \quad (465)$$

and

$$\begin{aligned} t_r &= \frac{\pi - \phi}{\omega_d} \rightarrow \phi = \pi - t_r \omega_d \\ &= \pi - \frac{t_r}{t_p} \pi \\ &= \pi \left(1 - \frac{t_r}{t_p}\right) \end{aligned} \quad (466)$$

recall that

$$\phi = \sin^{-1} \sqrt{1 - \zeta^2} \quad (467)$$

$$\begin{aligned} 1 - \zeta^2 &= \sin^2 \phi \\ \zeta &= \sqrt{1 - [\sin(\phi)]^2} \end{aligned} \quad (468)$$

and

$$\omega_n = \frac{\omega_d}{\sqrt{1 - \zeta^2}} \quad (469)$$

- Obtain  $\zeta$  from  $M_p$ .

$$\begin{aligned} M_p &= e^{-\frac{\zeta}{\sqrt{1-\zeta^2}}\pi} \\ -\frac{\zeta}{\sqrt{1-\zeta^2}}\pi &= \ln(M_p) \\ \zeta^2 \pi^2 &= (1 - \zeta^2)(\ln(M_p))^2 \\ \zeta^2 \pi^2 &= (\ln(M_p))^2 - \zeta^2 (\ln(M_p))^2 \\ \zeta^2 \left[ \pi^2 + (\ln(M_p))^2 \right] &= (\ln(M_p))^2 \\ \zeta^2 &= \frac{(\ln(M_p))^2}{\pi^2 + (\ln(M_p))^2} \\ \zeta &= \frac{|\ln(M_p)|}{\sqrt{\pi^2 + (\ln(M_p))^2}} \end{aligned} \quad (470)$$

- Obtain  $\omega_n$  and  $\zeta$  from  $t_p$  and  $t_s$ .

$$t_p = \frac{\pi}{\omega_d} \rightarrow \omega_d = \frac{\pi}{t_p} \quad (471)$$

next

$$\begin{aligned}\omega_n &= \frac{\omega_d}{\sqrt{1 - \zeta^2}} \\ &= \frac{\frac{\pi}{t_p}}{\sqrt{1 - \zeta^2}} \\ &= \frac{\pi}{t_p \sqrt{1 - \zeta^2}}\end{aligned}\tag{472}$$

looking at the settling time

$$\begin{aligned}t_s &= \frac{4}{\zeta \omega_n} \\ &= \frac{4t_p \sqrt{1 - \zeta^2}}{\zeta \pi} \\ t_s \zeta \pi &= 4t_p \sqrt{1 - \zeta^2} \\ t_s^2 \zeta^2 \pi^2 &= 16t_p^2(1 - \zeta^2) \\ (16t_p^2 - \pi^2 t_s^2)\zeta^2 &= 16t_p \\ \zeta^2 &= \frac{16t_p}{16t_p^2 - \pi^2 t_s^2} \\ &= \frac{1}{1 - \left(\frac{\pi t_s}{4 t_p}\right)^2} \\ \zeta &= \frac{1}{\sqrt{1 - \left(\frac{\pi t_s}{4 t_p}\right)^2}}\end{aligned}\tag{473}$$

- Obtain  $\omega_n$  and  $\zeta$  from  $M_p$  and  $t_r$ .

$$\zeta = \frac{|\ln M_p|}{\sqrt{\pi^2 + (\ln M_p)^2}}\tag{474}$$

Recall that

$$t_r = \frac{\pi - \phi}{\omega_n \sqrt{1 - \zeta^2}} = \frac{\pi - \phi}{\omega_d}\tag{475}$$

Therefore

$$\omega_n = \frac{\pi - \phi}{t_r \sqrt{1 - \zeta^2}}\tag{476}$$

where

$$\phi = \sin^{-1} \sqrt{1 - \zeta^2}\tag{477}$$

System identification can be done through measuring performance indicators using Simulink.

### Example 6.1 SIMULINK Tutorial on Measuring Performance Indicators

System identification can be done through measuring performance indicators using Simulink. Consider a 2<sup>nd</sup>-order system with  $k = 700 \text{ N/m}$ ,  $c = 15 \text{ kg/s}$ , and  $m = 7 \text{ kg}$ . To implement this in Simulink, one could use MATLAB to generate the transfer function:

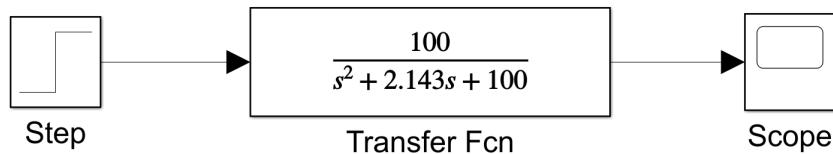
Listing 10: MATLAB code for Method II.

```

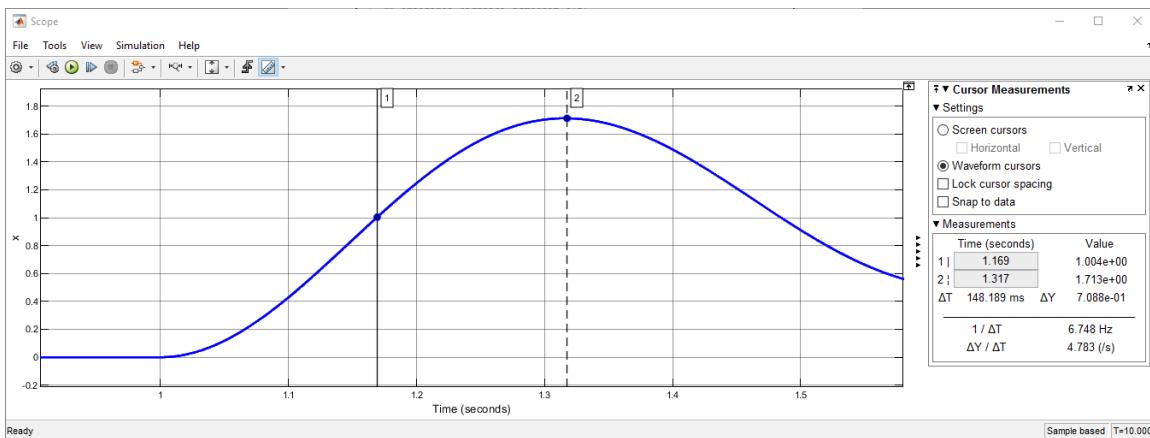
1 %% calculate system values
2 c = 15;          % N/(m/sec)
3 k = 700;         % N/m,
4 m = 7;           % kg
5 w_n = sqrt(k/m)    % natural frequency in rad/sec
6 z = c/(2*sqrt(k*m)) % damping ratio
7 w_d = w_n*sqrt(1-z^2);   % damped natural frequency in rad/sec
8
9 %% calculate transfer function G(s)
10 B = [w_n^2];
11 A = [1 2*z*w_n w_n^2];
12 G = tf(B,A) % build the transfer function

```

and then deploy the transfer function to SIMULINK as shown below



Using the scope and waveform cursors, several performance indicators can be read



these include

- $t_r = 0.169 \text{ sec}$
- $t_p = 0.317 \text{ sec}$

- $x_p = 1.173 \text{ m}$
- $M_p = 71.3\%$

From these, measurements the damping ratio of the system can be obtained using one of two methods,

- **Method 1** Estimate damping ratio using  $t_r, t_p$ . Knowing that

$$\begin{aligned}\phi &= \pi(1 - t_r/t_p) \\ &= 1.46674\end{aligned}\quad (478)$$

(479)

and

$$\begin{aligned}\zeta &= \sqrt{1 - [\sin(\phi)]^2} \\ &= 0.1039\end{aligned}\quad (480)$$

(481)

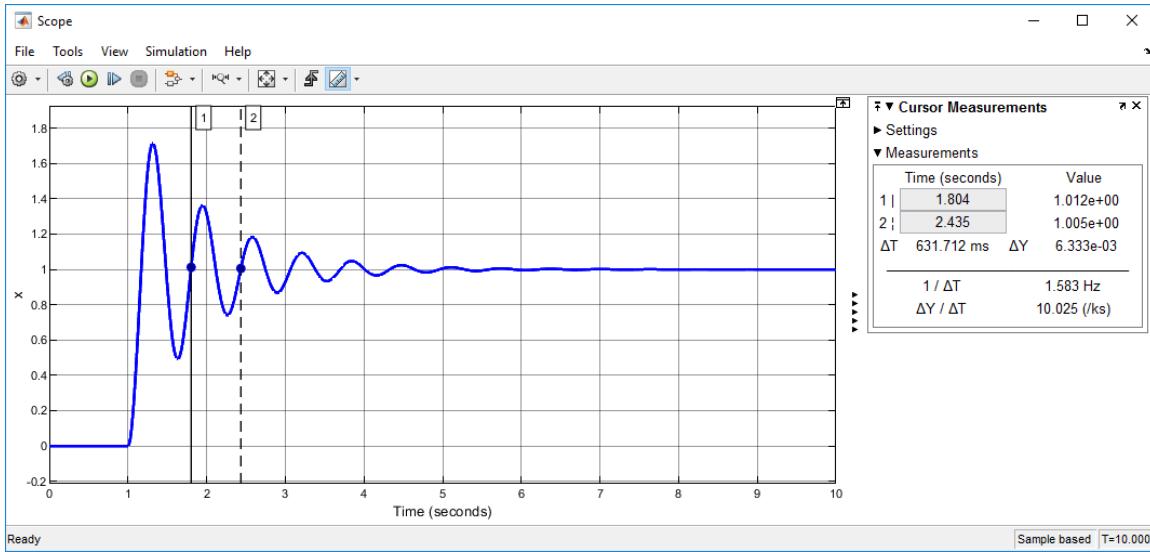
- **Method 2** Estimate damping ratio using  $M_p$ . Knowing that

$$\begin{aligned}\zeta &= \frac{|\ln M_p|}{\sqrt{\pi^2 + (\ln M_p)^2}} \\ &= \frac{|\ln(0.72)|}{\sqrt{\pi^2 + (\ln 0.72)^2}} \\ &= 0.1071\end{aligned}\quad (482)$$

(483)

These values give an error of 3.4% and 0% when compared to the correct value of  $\zeta = 0.1071$ , respectively. This is considered very good, but these are ideal systems with no noise and therefore excellent predictions are to be expected.

The SIMULINK model can also be used to estimate the natural frequency of the system using the zero crossing method by placing the cursors as close as possible to the “zero” value, 1 cycle apart. Here, “zero” is one due to the system being subjected to a step response.



The  $\Delta T$  of the cursors is 0.631712 sec, or 1.583 Hz. This is 0.53% off from the true value of 1.5915 Hz.

### 6.3.4 Logarithmic decrement

Logarithmic decrement ( $\zeta \ll 1$ ) For a vibrating system, the mass ( $m$ ) and stiffness ( $k$ ) can be measured using scales and static deflection tests. However, the damping coefficient ( $c$ ) is a more difficult quantity to determine. From  $k$  and  $m$  we can compute the natural frequency ( $\omega_n$ ) and the critical damping coefficient ( $c_{cr}$ ). Therefore, knowing that the critical damping ratio ( $\zeta$ ) is defined as:

$$\zeta = \frac{c}{c_{cr}} = \frac{c}{2\sqrt{km}} = \frac{c}{2m\omega_n} \quad (484)$$

if we calculate  $\zeta$ , we can obtain  $c$  for the system of interest. This is made possible because  $c_{cr}$  can be calculated from  $k$  and  $m$ . Observing the temporal response for the underdamped system,

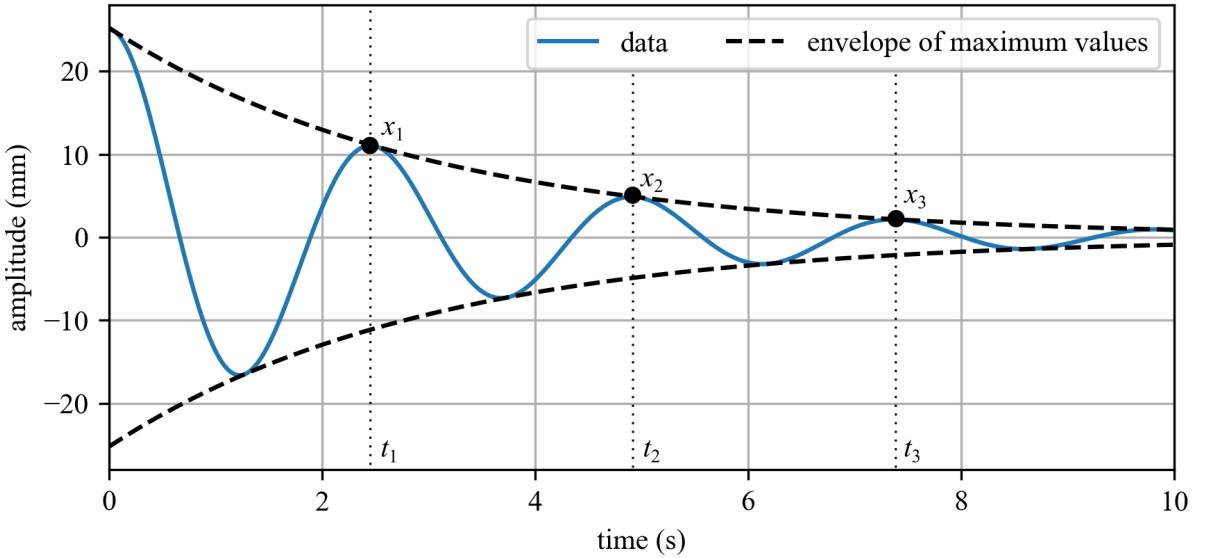


Figure 6.5: Measuring the peak displacements points in an experimental system with decay caused by damping.

we mark three points of maximum amplitude,  $x_1$ ,  $x_2$ , and  $x_3$  that happen at  $t_1$ ,  $t_2$ , and  $t_3$ , respectively. Considering displacement values for the first two points  $x_1$  and  $x_2$ , separated by a complete period ( $T$ ). Knowing that one cycle is  $2\pi$ , the time period for this complete cycle is given by:

$$t_2 - t_1 = \frac{2\pi}{\omega_d} = \frac{2\pi}{\omega_n \sqrt{1 - \zeta^2}} \quad (485)$$

where  $\omega_d$  is the damped natural frequency. This is the time period ( $T$ ) of damped oscillations. If we derive an equation for the values of the peaks, also called the envelope of maximum values, we get:

$$x_{\text{peaks}} = Ae^{-\zeta\omega_n t} \quad (486)$$

Knowing that the system is underdamped,  $A$  can be solved for using the initial conditions  $x_0$  and  $v_0$ , therefore:

$$A = \frac{\sqrt{(v_0 + \zeta\omega_n x_0)^2 + (x_0\omega_d)^2}}{\omega_d} \quad (487)$$

In terms of  $t_1$  and  $t_2$ , we can express the displacement at these times as:

$$x_1 = Ae^{-\zeta\omega_n t_1} \quad (488)$$

and

$$x_2 = Ae^{-\zeta\omega_n t_2} \quad (489)$$

therefore:

$$\frac{x_1}{x_2} = \frac{e^{-\zeta\omega_n t_1}}{e^{-\zeta\omega_n t_2}} = e^{\zeta\omega_n(t_2 - t_1)} \quad (490)$$

However, from before we know that  $t_2 - t_1 = \frac{2\pi}{\omega_d} = \frac{2\pi}{\omega_n \sqrt{1-\zeta^2}}$ . Therefore, we can express this last equation as:

$$\frac{x_1}{x_2} = e^{\left(\frac{2\pi\zeta}{\sqrt{1-\zeta^2}}\right)} \quad (491)$$

Next, we take the natural log of both sides to get the logarithmic decrement, denoted by  $\delta$ :

$$\delta = \ln\left(\frac{x_1}{x_2}\right) = \ln\left(\frac{x(t_1)}{x(t_1 + T)}\right) = \frac{2\pi\zeta}{\sqrt{1-\zeta^2}} \quad (492)$$

This shows us that the ratio of any two successive amplitudes for an underdamped system, vibrating freely, is constant and is a function of the damping only. Sometimes, in experiments, it is more convenient/accurate to measure the amplitudes after say “ $n$ ” peaks rather than two successive peaks (because if the damping is very small, the difference between the successive peaks may not be significant). The logarithmic decrement can then be given by the equation

$$\delta = \frac{1}{n} \ln\left(\frac{x_1}{x_{n+1}}\right) = \frac{1}{n} \ln\left(\frac{x(t_1)}{x(t_1 + nT)}\right) = \frac{2\pi\zeta}{\sqrt{1-\zeta^2}} \quad (493)$$

Once we use the experimental data to obtain  $\delta$ , and knowing that:

$$\delta = \frac{2\pi\zeta}{\sqrt{1-\zeta^2}} \quad (494)$$

we can calculate the value of  $\zeta$ :

$$\zeta = \frac{\delta}{\sqrt{4\pi^2 + \delta^2}} \quad (495)$$

Therefore, having  $\zeta$  we can solve for the coefficient of damping,  $c$ , as:

$$c = \zeta 2\sqrt{km} \quad (496)$$

**Example 6.2** Calculate the damping coefficient for the system with the measured amplitude as expressed below given that  $m = 3$  kg and  $k = 43$  N/m. Use  $t_1 = 1$  sec, and  $t_{n+1} = t_4 = 6$  sec. Use the peaks as marked in figure 6.6.

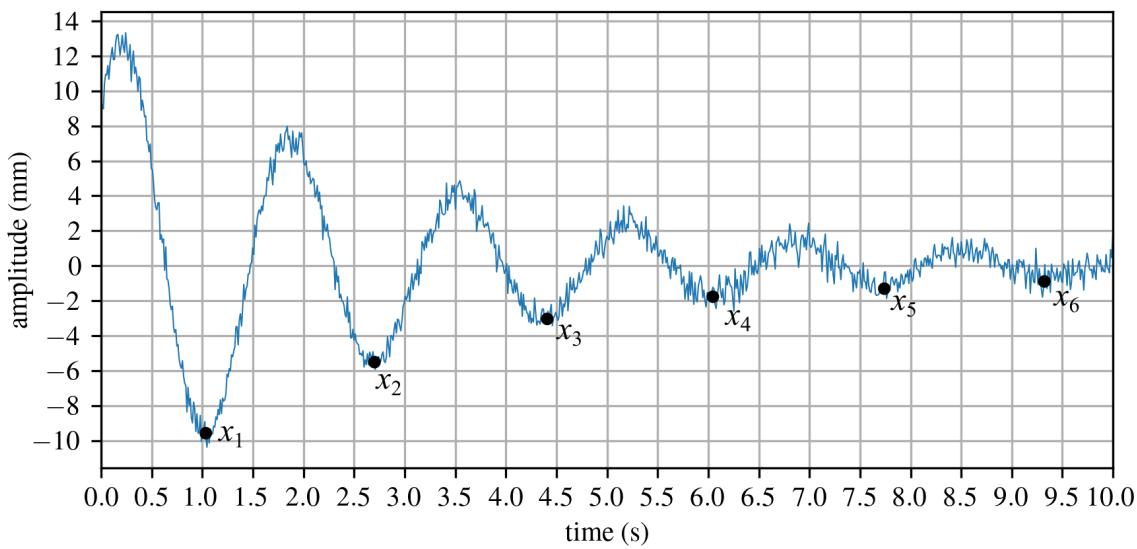


Figure 6.6: Response from an experimental system with noise.

**Solution:**

First, from the plot we can determine that  $x_1 = -9.5$  mm and  $x_4 = -1.8$  mm where  $n = 3$ . Thereafter, we can solve for  $\delta$ :

$$\delta = \frac{1}{3} \ln \left( \frac{x_1}{x_4} \right) = \frac{1}{3} \ln \left( \frac{-9.5}{-1.8} \right) = 0.554 \quad (497)$$

Next, we can calculate  $\zeta$ , as:

$$\zeta = \frac{\delta}{\sqrt{4\pi^2 + \delta^2}} = \frac{0.554}{\sqrt{4\pi^2 + 0.554^2}} = 0.0879 \quad (498)$$

And lastly:

$$c = \zeta 2\sqrt{km} = 0.0879 \cdot 2\sqrt{43 \cdot 3} = 2.0 \text{ kg/s} \quad (499)$$

**Example 6.3** A vehicle wheel, tire, and suspension can be modeled as a SDOF spring and mass as depicted below:

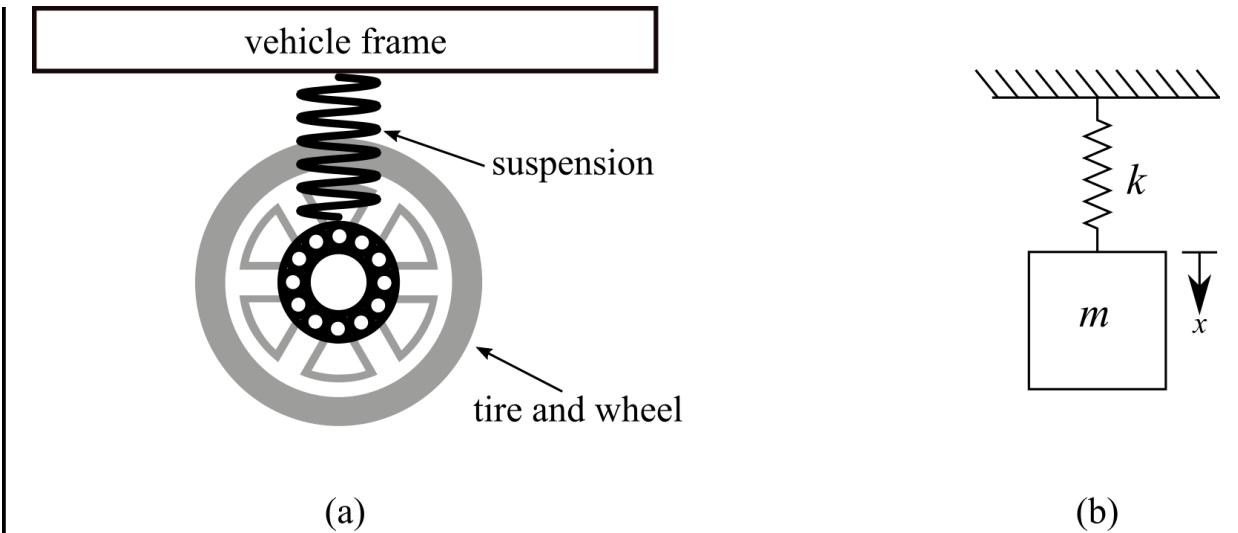


Figure 6.7: Modeling of a vehicle wheel, tire, and suspension showing: (a) Graphical representation; and (b) a spring-mass model.

The free response of a 1000-kg automobile with a stiffness of  $k = 400,000 \text{ N/m}$  is observed to be underdamped. Modeling the automobile as a single-degree-of-freedom oscillation in the vertical direction, as annotated in figure 6.7, determine the damping coefficient if the displacement at  $t_1$  is measured to be 2 cm and 0.22 cm at  $t_2$ .

**Solution:**

Knowing  $x_1 = 2 \text{ cm}$  and  $x_2 = 0.22 \text{ cm}$  and  $t_2 = T + t_1$ , therefore:

$$\delta = \ln \frac{x_1}{x_2} = \ln \frac{2}{0.22} = 2.207 \quad (500)$$

and:

$$\zeta = \left( \frac{\delta}{\sqrt{4\pi^2 + \delta^2}} \right) = \left( \frac{2.207}{\sqrt{4\pi^2 + 2.207^2}} \right) = 0.331 \quad (501)$$

therefore, we can obtain the damping coefficient as

$$c = 2\zeta\sqrt{km} = 2(0.331)\sqrt{400,000 \cdot 1,000} = 13,256 \text{ kg/s} \quad (502)$$

## 7 Control Systems

### 7.1 On-Off (Bang-Bang) Control

On-off (bang-bang) control is the most basic type of control system, operating by simply switching the output fully on or fully off in response to the measured variable. When the process value falls below a setpoint, the controller turns the system on; when it exceeds the setpoint, it turns it off. This approach is common in applications where precise control is not required, such as hot water heaters and HVAC systems, offering a simple, low-cost, and reliable method of maintaining conditions within an acceptable range.

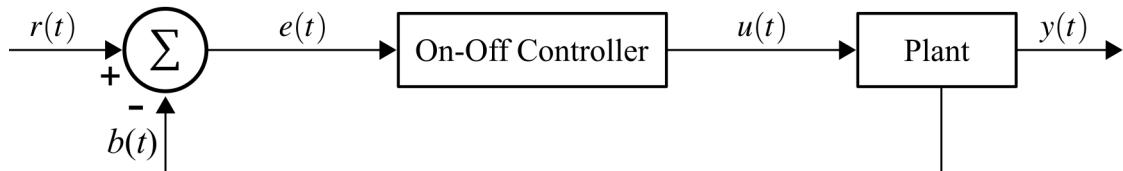


Figure 7.7: On-Off (Bang-Bang) controller for a system with feedback, where  $r(t)$  is the desired setpoint (SP) and  $y(t)$  is the measured process value (PV).

### 7.2 Proportional-Integral-Derivative (PID) Control

Proportional-Integral-Derivative (PID) Control is a three-term controller that employs feedback that is widely used in continuous control systems, including for the control of structural systems. A PID controller seeks to minimize the measured error value  $e(t)$  between a desired setpoint (SP) and a measured process variable (PV) by applying corrections based on the proportional ( $P$ ), integral ( $I$ ), and derivative ( $D$ ) terms (denoted P, I, and D respectively), from which it gets its name.

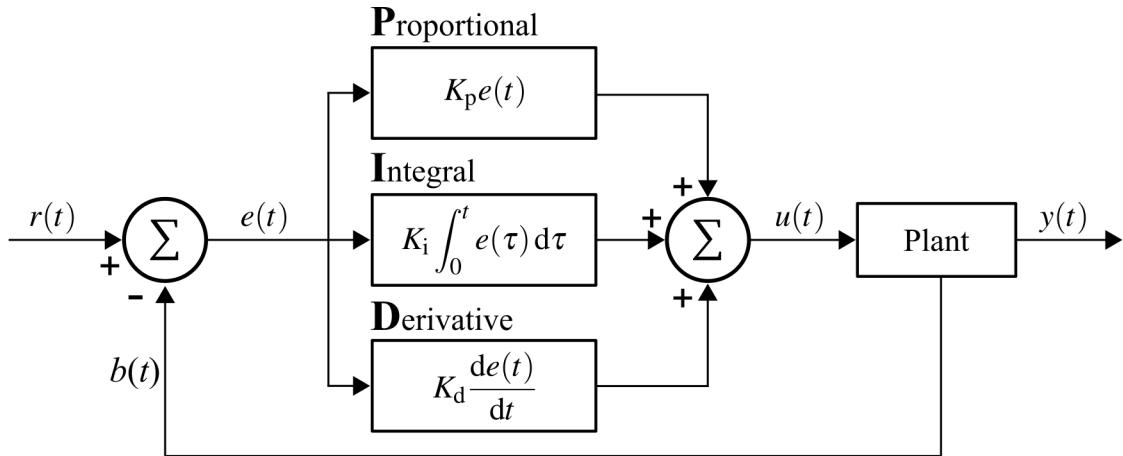


Figure 7.8: Generalized PID controller for a system with feedback, where  $r(t)$  is the desired setpoint (SP) and  $y(t)$  is the measured process value (PV).

The overall control equation is defined as

$$u(t) = K_p e(t) + K_i \int_0^t e(\tau) d\tau + K_d \frac{de(t)}{dt} \quad (7.7)$$

where  $K_p$ ,  $K_i$ , and  $K_d$  are non-negative coefficients for the proportional, integral, and derivative terms, respectively. The PID controller is diagrammed in figure 7.8 for a system with feedback control, such as that shown in figure ???. Moreover, in the Laplace-derived  $s$  domain, the transfer function of the PID controller is defined as

$$\mathcal{L}[s] = K_p + \frac{K_i}{s} + K_d s \quad (7.8)$$

where  $s$  is the complex frequency. A temporal response for the 1-DOF shown in figure ?? when controlled with a PID controller is reported in figure 7.9.

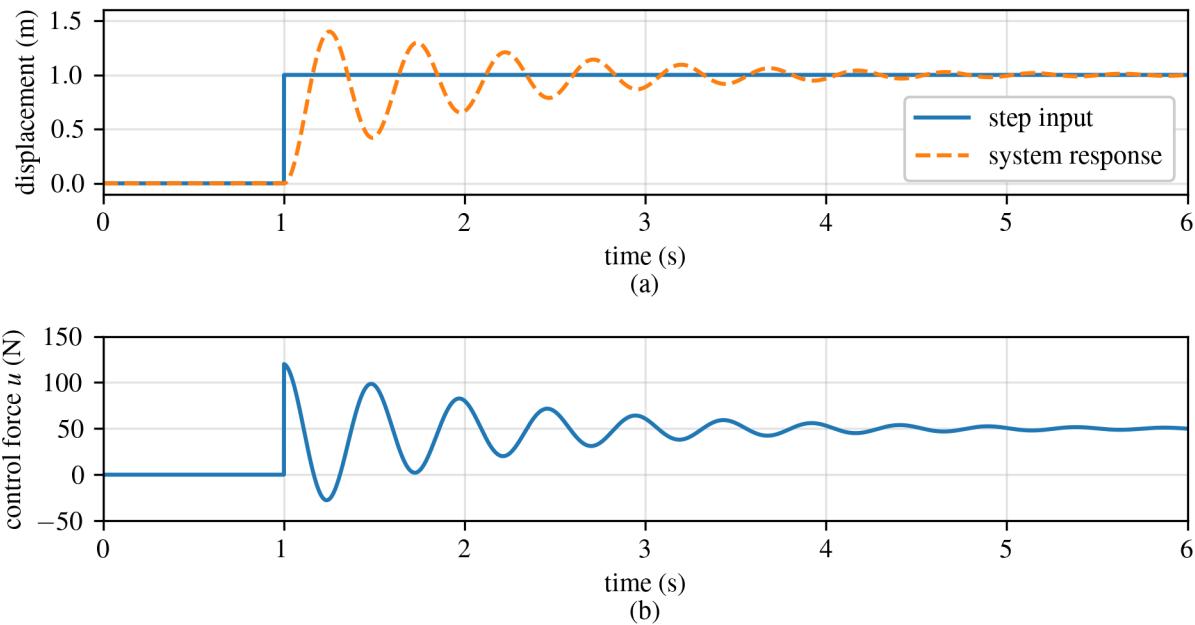


Figure 7.9: Response for a 1-DOF system controlled with a PID controller, showing: (a) system response and the step function input, and (b) the controller response imposed on the system to obtain such responses.

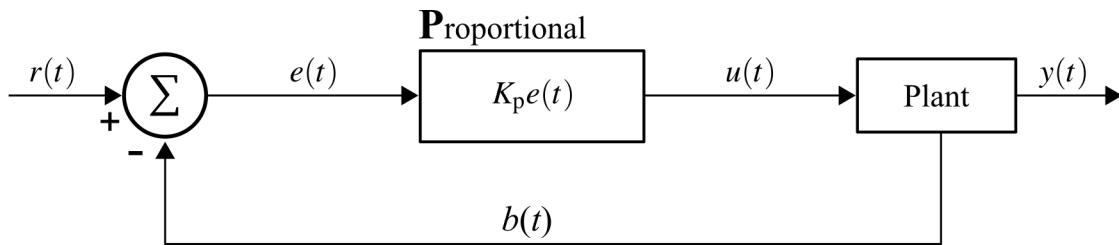


Figure 7.10: Generalized PID controller for a system with feedback, where  $r(t)$  is the desired setpoint (SP) and  $y(t)$  is the measured process value (PV).

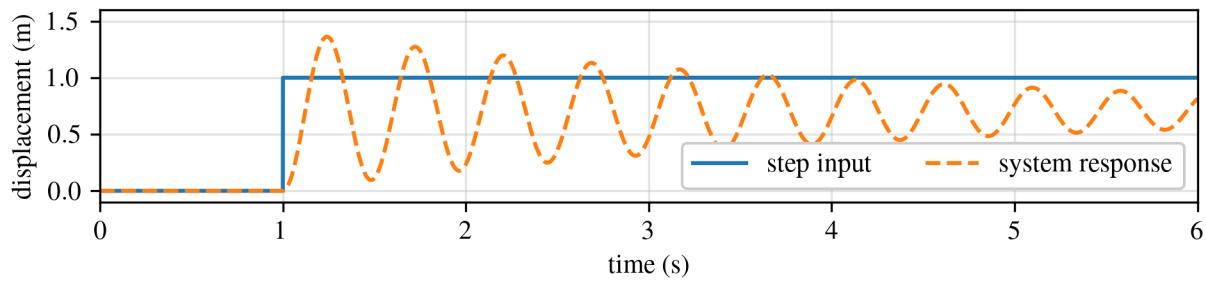


Figure 7.11: Response for a 1-DOF system controlled with a just a proportional (P) controller.

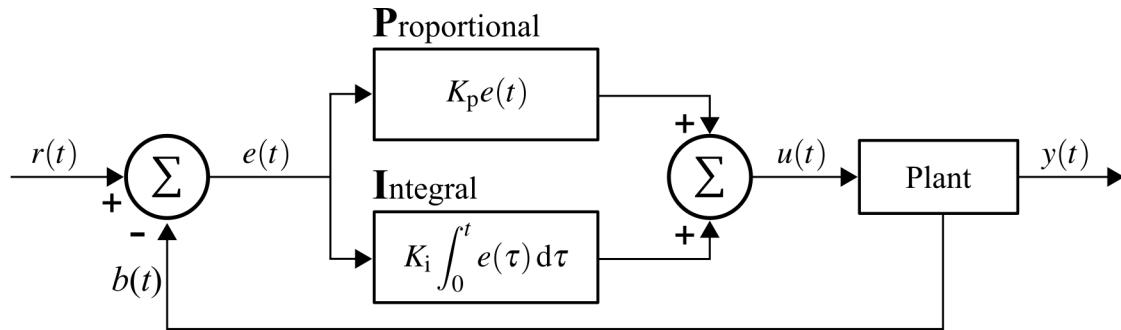


Figure 7.12: Generalized PID controller for a system with feedback, where  $r(t)$  is the desired setpoint (SP) and  $y(t)$  is the measured process value (PV).

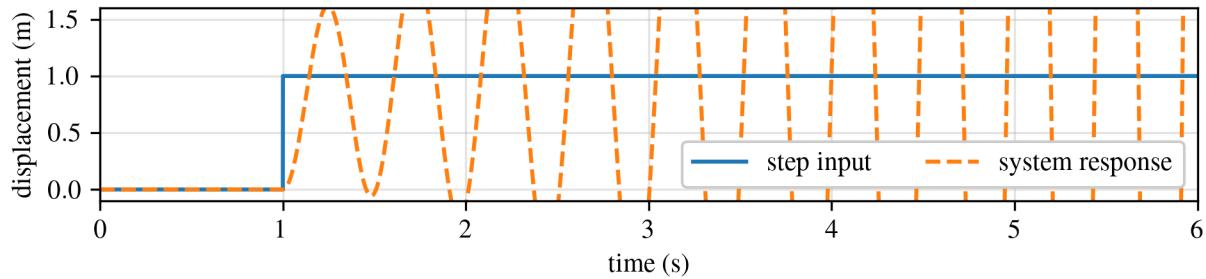


Figure 7.13: Response for a 1-DOF system controlled with a PI controller.

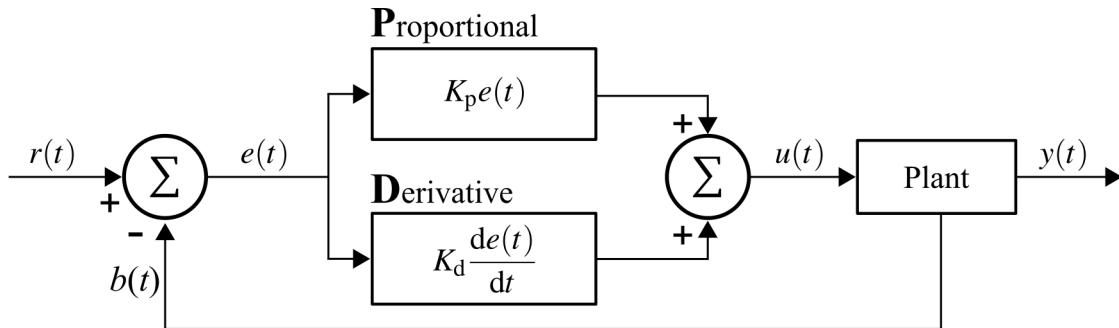


Figure 7.14: Generalized PID controller for a system with feedback, where  $r(t)$  is the desired setpoint (SP) and  $y(t)$  is the measured process value (PV).

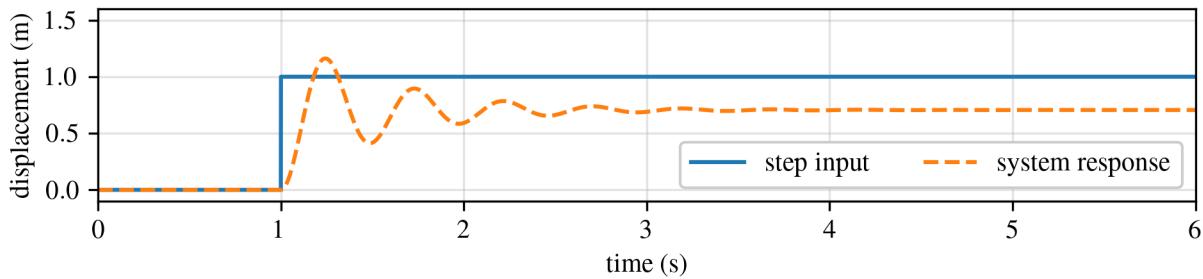
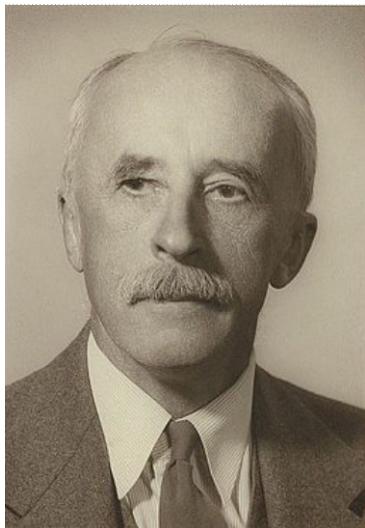


Figure 7.15: Response for a 1-DOF system controlled with a PD controller.

### Review 7.1 Nicolas Minorsky and the Need for Better Control

Continuous control systems have been widely used for centuries. For example, consider that the centrifugal governor which uses spinning weights was used by Christiaan Huygens in the 1600s in the Netherlands to regulate the gap between millstones in windmills or by James Watt who famously linked a stem regulator to a centrifugal governor to control steam turbines.

Arguably, the Russian American engineer Nicolas Minorsky was the first to develop the theoretical analysis for the three-term control we now call PID. This was done in 1922 while he was researching and designing automatic ship steering for the US Navy. He based his work on watching how a ship's helmsman responds to wave loading on a ship, with a delayed input to the helm that not only considered the current ship course but also past errors and the desired rate of change for the ship. For a helmsman, the goal is stability, not absolute control, which simplifies how one thinks about the challenge of control.



(a)



(b)

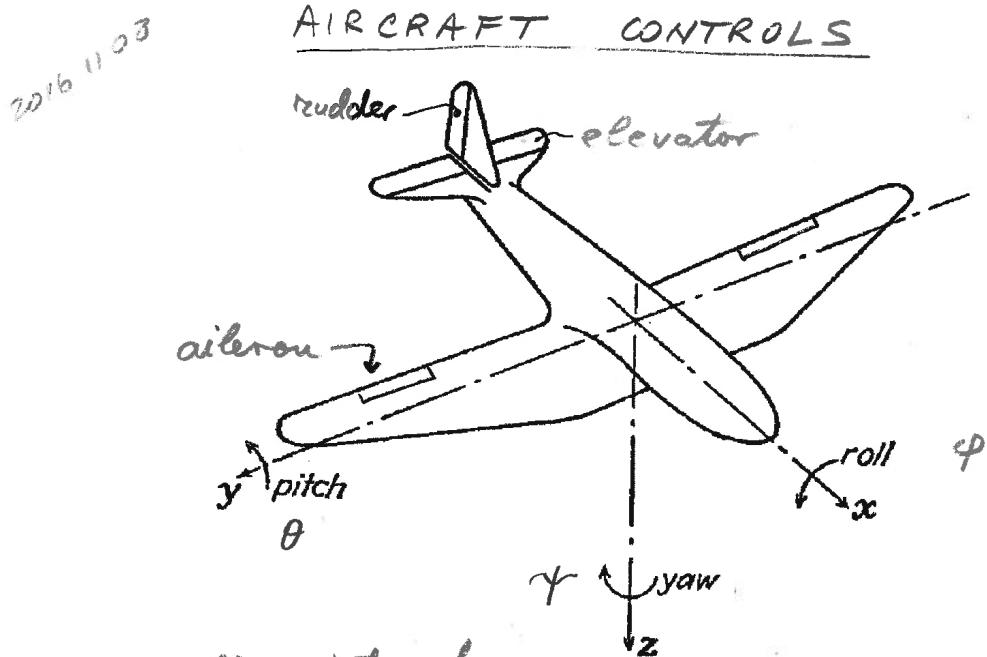
Figure 7.16: Historical perspective of PID control showing: (a) Portrait of Nicolas Minorsky<sup>1</sup> and (b) the battleship USS New Mexico (BB-40) of the United States Navy which was the first to implement PID control in its steering<sup>2</sup>.

<sup>1</sup>Peter Minorsky, grandson of Nicolas Minorsky,  
<<https://creativecommons.org/licenses/by-sa/1.0>>, via Wikimedia Commons

CC BY-SA 1.0

<sup>2</sup>U.S. Navy, Public domain, via Wikimedia Commons

## 7.3 Control Systems

Aircraft control surfaces

AILERONS control the rolling motion

$\varphi$  = roll angle (~bank angle)

pilot stick moves left-right to deflect ailerons

ELEVATOR controls the pitch motion  
nose up / nose down

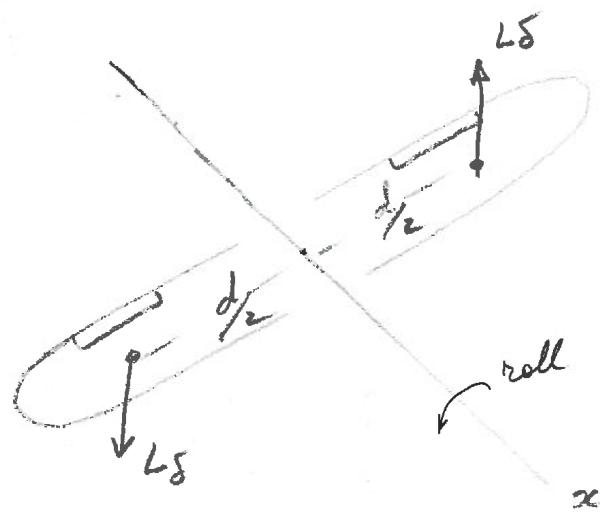
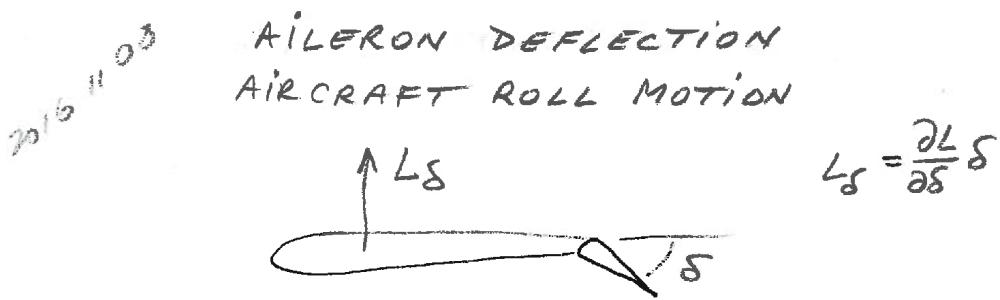
$\theta$  = pitch angle

pilot stick moves forward-backward to deflect elevator

RUDDER controls yaw motion

$\psi$  = yaw angle

pilot uses rudder pedals to deflect rudder



- Aileron deflection  $\delta$  produces additional lift  $L\delta$
- $L\delta$  is up/down because ailerons left/right move up/down
- Net effect is a rolling moment,

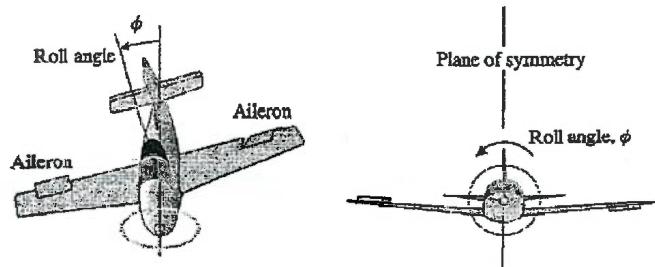
$$M = L\delta d = d \cdot \frac{\partial L}{\partial \delta} \cdot \delta \quad (1)$$

- Essentially, the rolling moment  $M$  is proportional to aileron deflection  $\delta$ , i.e.

$$M = K \delta \quad (K = \text{gain}) \quad (160/523)$$

RTF

## ROLL TRANSFER FUNCTION



$$EOM: J\ddot{\phi} + c\dot{\phi} = M \quad (1)$$

where

$J$  = inertia : mass moment of inertia about roll axis

$c$  = damping : air resistance to roll motion

$M$  = rolling moment produced  
by aileron deflection  $\delta$

Recall :

$$M = K\delta \quad (2)$$

$$(2) \rightarrow (1) : J\ddot{\phi} + c\dot{\phi} = K\delta \quad (3)$$

Take  $L T$  of Eq. (3) to get

$$L(3) : (Js^2 + cs)\tilde{\phi}(s) = K \tilde{\delta}(s) \quad (4)$$

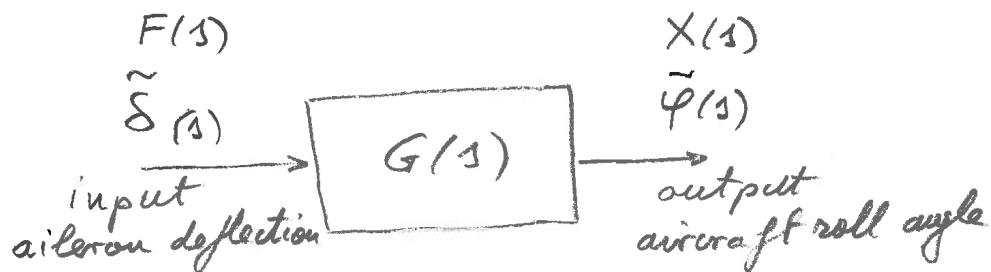
$$\text{where } \tilde{\phi}(s) = L\phi(t) \quad (5)$$

$$\tilde{\delta}(s) = L\delta(t)$$

Solution of Eq.(4) yields :

167/523

$$\tilde{\varphi}(s) = \frac{K}{Js^2 + Cs} \tilde{\delta}(s) \quad (6)$$



$$G(s) = \frac{K}{Js^2 + Cs} \quad (7)$$

$$X(s) = G(s) F(s). \quad (8)$$

The system described by Eq.(7) is an uncontrolled 2nd order dynamic system.

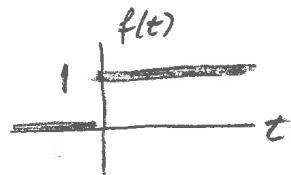
- This system is "uncontrolled" because a constant input will produce a continuously growing response (see roll response on next page).
- Another example of uncontrolled 2nd order dynamic system is the DC Motor where an applied constant voltage produces continuous rotation.

RE 09  
2016/11

### ROLL RESPONSE

#### TO CONSTANT AILERON DEFLECTION

Assume the pilot moves the stick laterally such as to create a constant aileron deflection  $\delta = \text{const}$ . This corresponds to a step input, i.e.,



$$\left\{ \begin{array}{l} f(t) = 1 \\ F(s) = \frac{1}{s} \end{array} \right. \quad (9)$$

$$(9) \rightarrow (8): \quad X(s) = \frac{K}{Js^2 + Cs} \cdot \frac{1}{s} = \frac{K}{s^2(Js + C)} \quad (10)$$

Table 2-1, # 19 has the pair

$$t - T(1 - e^{-t/T}) \quad \frac{LT}{s^2(Ts + 1)} \quad (11)$$

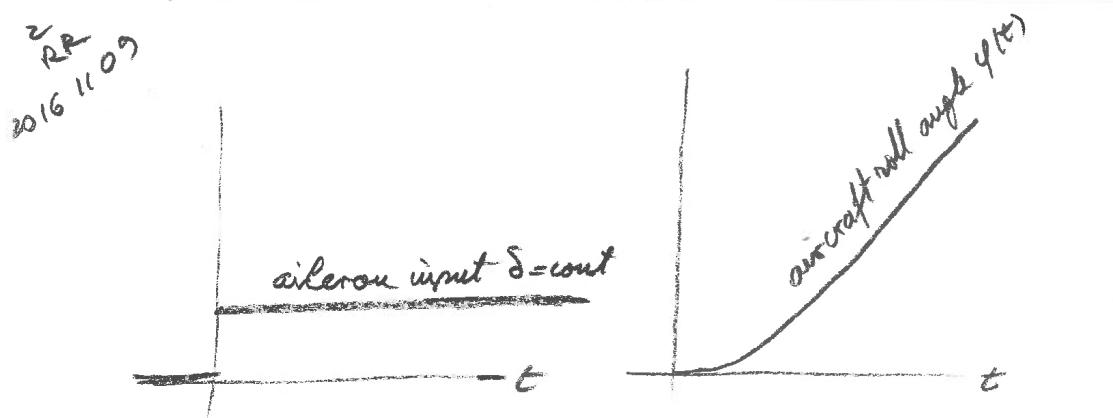
Write (10) such as to look like (11), i.e.,

$$X(s) = \frac{K}{C} \cdot \frac{1}{s^2(Ts + 1)}, \quad T = \frac{J}{C} \quad (12)$$

ILT of (12) gives:

$$x(t) = \frac{K}{C} \left[ t - T(1 - e^{-t/T}) \right] \quad (13)$$

$$x_{ss} = \lim_{t \rightarrow \infty} x(t) = \frac{K}{C} (t - T) \quad (14)$$



Plot of eq. (13) indicates that the aircraft will roll continuously with a constant roll rate.

Thus, a constant aileron input  $\delta = \text{const}$  produces continuously increasing roll angle of aircraft.

This is a general property of Type I systems: they cannot maintain position.

The step response of a Type I system is unconstrained

- DC motor spins continuously under constant voltage input
- Aircraft rolls continuously under constant aileron input

$\frac{3RR}{201b^{11}0^9}$

### Step response of Type 1 systems

$$G(s) = \frac{K}{s} \cdot \frac{(T_0 s + 1)(T_1 s + 1) \dots}{(T_1 s + 1)(T_2 s + 1) \dots} \quad \text{Type 1}$$

$$F(s) = \frac{1}{s} \quad \text{step excitation}$$

$$X(s) = G(s) F(s)$$

$$= \frac{K}{s} \cdot \frac{(T_0 s + 1) \cancel{(s)} \dots}{(T_1 s + 1) \cancel{(s)} \dots} \cdot \frac{1}{s}$$

$$X(s) = \frac{K}{s^2} \cdot \frac{(s) \dots}{(s) \dots} \quad (1)$$

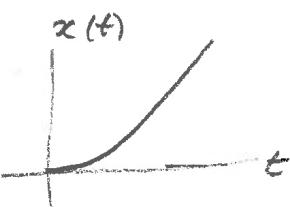
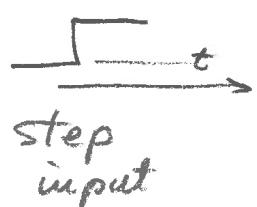
Steady state response is calculated with  
Final Value Theorem, i.e.,

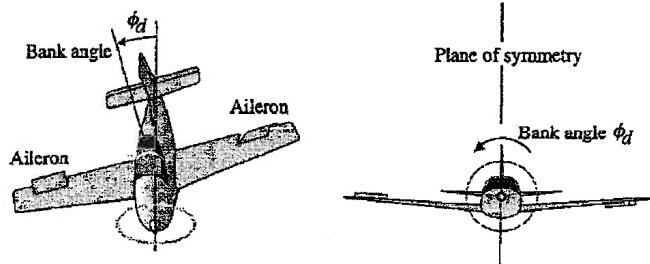
$$x_{ss} = \lim_{t \rightarrow \infty} x(t) = \lim_{s \rightarrow 0} s X(s) \quad (2)$$

(1)  $\rightarrow$  (2) :

$$x_{ss} = \lim_{s \rightarrow 0} s \cdot \frac{K}{s^2} \cdot \underbrace{\frac{(T_0 s + 1)(T_1 s + 1) \dots}{(T_1 s + 1)(T_2 s + 1) \dots}}_{\overset{s \rightarrow 0}{\longrightarrow} 1} \rightarrow 1$$

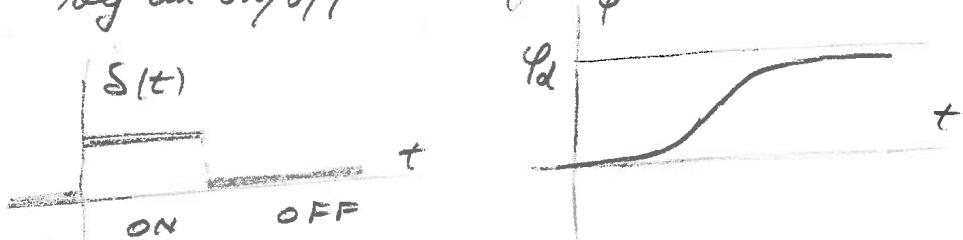
$$x_{ss} = \lim_{s \rightarrow 0} \frac{K}{s} = \infty$$



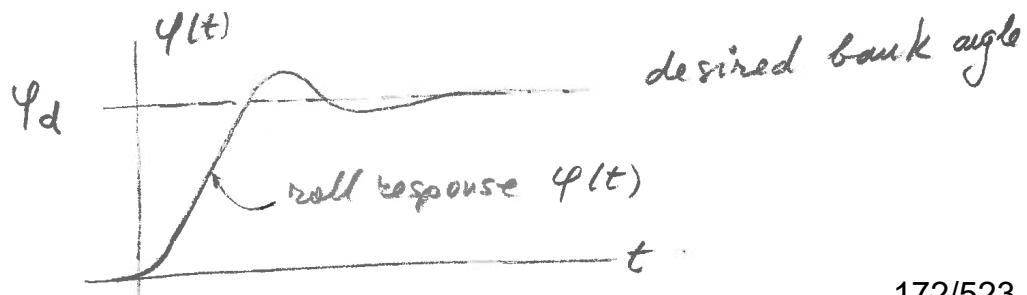
*BAC*BANK ANGLE CONTROL

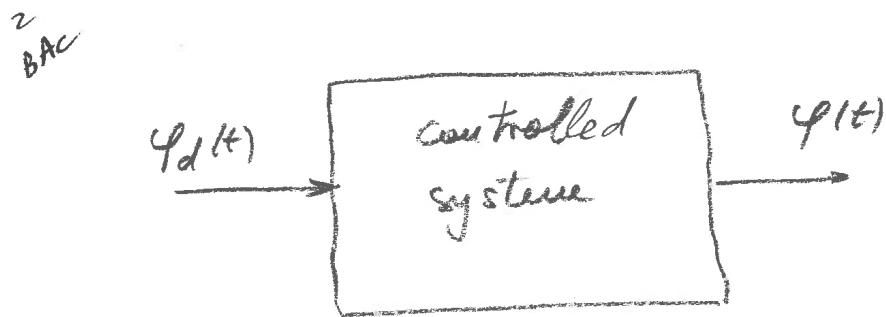
We want the aircraft to roll from flying straight & level to flying inclined with a bank angle  $\phi_d = \text{const.}$

- Manually, the pilot creates a bank angle by an on/off deflection of aileron



- We desire to build a FB control system to achieve this transition in a smooth way automatically.





The control system design process needs specifications. We choose two performance indicators,  $M_p$  and  $t_p$  and define their values as design specs.

### Control design specifications

DS1 : Fast response time as measured by rise time,  $t_r \leq 1.5$  sec.

DS2 : Maximum percentage overshoot for step input less than 20%  
 $M_p \leq 20\%$

BAC Details of:

### MANUAL BANK ANGLE CONTROL

- Aircraft flies straight & level



- Pilot wants to bank  $15^\circ$



- Pilot moves stick sideways.



- Aircraft starts to respond



- Pilot's eyes see the aircraft

rolling and estimates actual value  $\varphi$

- Pilot's mind processes the measured value  $\varphi$ , compares it with desired value  $\varphi_d = 15^\circ$ , and sends action order to hand muscles!

- if  $\varphi < \varphi_d$ , then continue to push stick

- if  $\varphi \approx \varphi_d$ , move stick back to neutral position to reduce rolling moment from ailerons

- if  $\varphi > \varphi_d$ , move stick the otherway because the aircraft overshot the target angle and needs to roll Backwards

FB  
2016/10/9

## FEEDBACK CONTROL

FB concept:

- adjust the system input to obtain a desired output

FB implementation in Laplace s-domain:

- measure output
- calculate "error", i.e. difference between "desired" and "measured"
- feed "error" to the system to adjust itself until the "error" is reduced to zero.

(i.e., "measured" = "desired")

In the time domain, the FB process is a transient process of repeated adjustments until output matches input ( $\text{error} \rightarrow 0$ ).

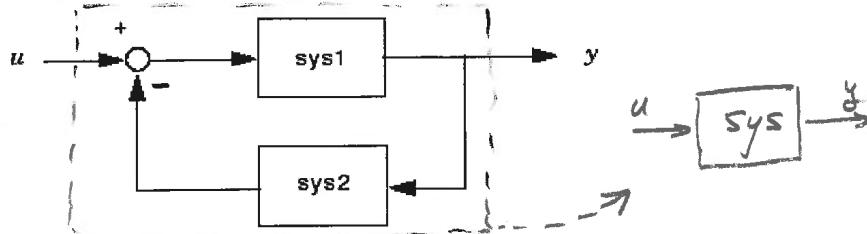
The process is based on Convolution Theorem of Laplace Transform

$$\mathcal{L}^{-1} G(s) F(s) = \int_0^t f(\tau) g(t-\tau) d\tau$$

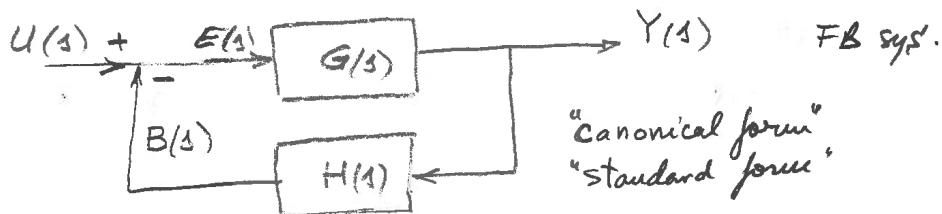
FB  
2016/11/10

## MATLAB:

`sys = feedback(sys1, sys2)` returns a model object `sys` for the negative feedback interconnection of model objects `sys1` and `sys2`.



The closed-loop model `sys` has `u` as input vector and `y` as output vector. The models `sys1` and `sys2` must be both continuous or both discrete with identical sample times. Precedence rules are used to determine the resulting model type



$$G_{CL}(s) = \frac{G(s)}{1 + G(s)H(s)}$$

closed loop  
transfer funct.

Proof

$$B(s) = H(s)Y(s)$$

$$E(s) = U(s) - B(s) = U(s) - H(s)Y(s)$$

$$Y(s) = G(s)E(s) = G(s)U(s) - G(s)H(s)Y(s)$$

$$Y(s) + G(s)H(s)Y(s) = G(s)U(s)$$

$$[1 + G(s)H(s)]Y(s) = G(s)U(s)$$

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

QED 176/523

3  
FB  
2016/11/10

### FB Nomenclature

$U(s)$ : input reference signal (desired result)

$Y(s)$ : output signal

$B(s)$ : feedback signal

$E(s)$ : error signal

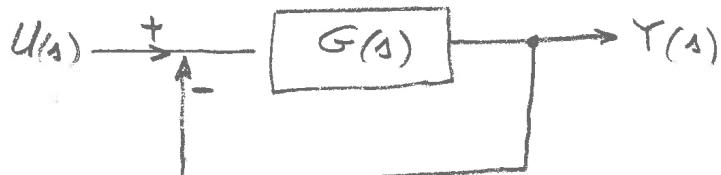
### Transfer functions

feed forward TF:  $G(s) = \frac{Y(s)}{E(s)}$

open loop TF:  $G(s)H(s) = \frac{B(s)}{E(s)}$

closed loop TF:  $G_{CL}(s) = \frac{Y(s)}{U(s)}$ .

### UNIT FEED BACK



"unit feedback" is obtained for  $H(s)=1$

$$G_{CL} = \frac{G}{1+G}$$

unit feedback  
closed loop TF.

## feedback

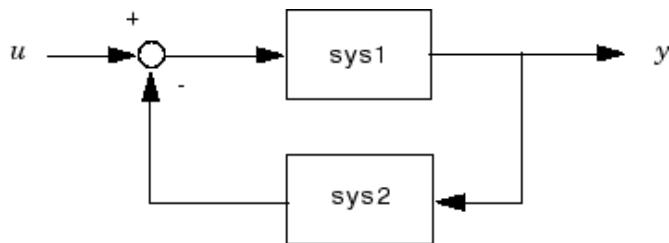
Feedback connection of two models

### Syntax

```
sys = feedback(sys1,sys2)
```

### Description

`sys = feedback(sys1,sys2)` returns a model object `sys` for the negative feedback interconnection of model objects `sys1` and `sys2`.



The closed-loop model `sys` has `u` as input vector and `y` as output vector. The models `sys1` and `sys2` must be both continuous or both discrete with identical sample times. Precedence rules are used to determine the resulting model type (see “Rules That Determine Model Type”).

To apply positive feedback, use the syntax

```
sys = feedback(sys1,sys2,+1)
```

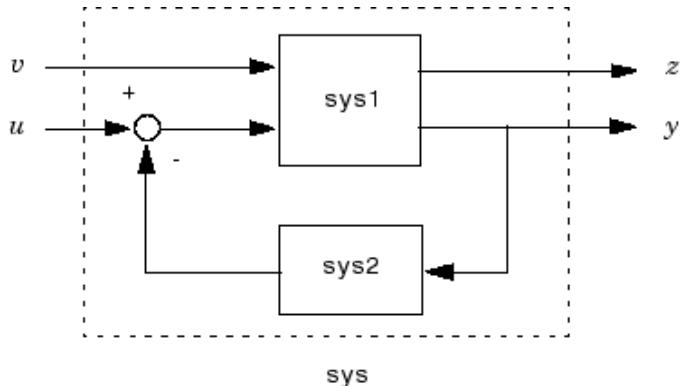
By default, `feedback(sys1,sys2)` assumes negative feedback and is equivalent to `feedback(sys1,sys2,-1)`.

Finally,

```
sys = feedback(sys1,sys2,feedin,feedout)
```

178/523

computes a closed-loop model **sys** for the more general feedback loop.



The vector **feedin** contains indices into the input vector of **sys1** and specifies which inputs **u** are involved in the feedback loop. Similarly, **feedout** specifies which outputs **y** of **sys1** are used for feedback. The resulting model **sys** has the same inputs and outputs as **sys1** (with their order preserved). As before, negative feedback is applied by default and you must use

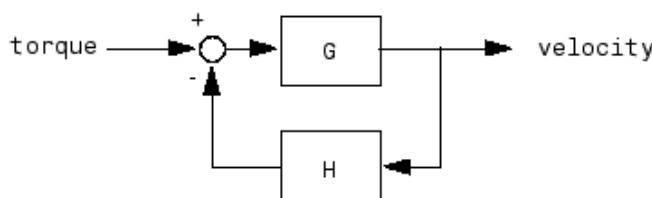
```
sys = feedback(sys1,sys2,feedin,feedout,+1)
```

to apply positive feedback.

For more complicated feedback structures, use **append** and **connect**.

## Examples

### Example 1



179/523

To connect the plant

$$G(s) = \frac{2s^2 + 5s + 1}{s^2 + 2s + 3}$$

with the controller

$$H(s) = \frac{5(s+2)}{s+10}$$

using negative feedback, type

```
G = tf([2 5 1],[1 2 3],'inputname','torque',...
       'outputname','velocity');
H = zpk(-2, -10, 5)
Cloop = feedback(G,H)
```

These commands produce the following result.

```
Zero/pole/gain from input "torque" to output "velocity":
0.18182 (s+10) (s+2.281) (s+0.2192)
-----
(s+3.419) (s^2 + 1.763s + 1.064)
```

The result is a zero-pole-gain model as expected from the precedence rules. Note that `Cloop` inherited the input and output names from `G`.

## Example 2

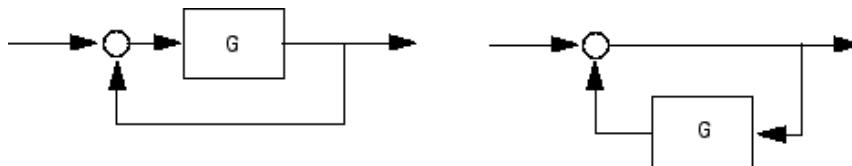
Consider a state-space plant `P` with five inputs and four outputs and a state-space feedback controller `K` with three inputs and two outputs. To connect outputs 1, 3, and 4 of the plant to the controller inputs, and the controller outputs to inputs 4 and 2 of the plant, use

```
feedin = [4 2];
feedout = [1 3 4];
Cloop = feedback(P,K,feedin,feedout)
```

## Example 3

You can form the following negative-feedback loops

180/523



by

```
Cloop = feedback(G,1)      % left diagram
Cloop = feedback(1,G)      % right diagram
```

## Limitations

The feedback connection should be free of algebraic loop. If  $D_1$  and  $D_2$  are the feedthrough matrices of `sys1` and `sys2`, this condition is equivalent to:

- $I + D_1D_2$  nonsingular when using negative feedback
- $I - D_1D_2$  nonsingular when using positive feedback.

### See Also

`series` | `parallel` | `connect`

Introduced before R2006a

$\hat{F}^B$ 

### Feedback control of Type 1 systems

- Type 1 system response is unconstrained
- Feedback can be used to control the response.

#### Examples

- DC motor cannot hold position; it rotates continuously under constant voltage  
With FB, a DC motor becomes a servomotor and holds position
- Aircraft rolls continuously; with FB, aircraft can maintain a constant bank angle.

#### Type 1 system transfer function

$$G(s) = \frac{K}{Js^2 + Cs} \quad \left\{ \begin{array}{l} K = \text{gain} \\ J = \text{inertia} \\ C = \text{damping} \end{array} \right.$$

2<sup>nd</sup> order system :  $s^2$  is highest power in denominator

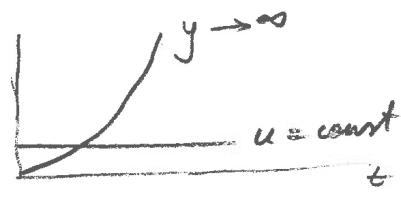
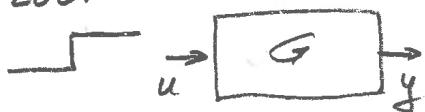
Type 1 system :  $G(s) = \frac{K}{s} \cdot \frac{1}{Js + c}$

$\overbrace{s}$   
"s to power 1"

<sup>8</sup>  
FB Step response of Type 1 system

$$G(s) = \frac{K}{Js^2 + Cs}$$

OPEN LOOP



CLOSED LOOP w FB



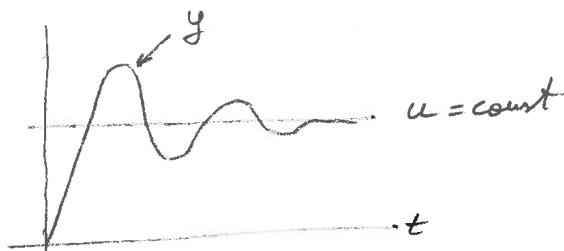
$$G_{CL} = \frac{G}{1+G}$$

$$= \frac{\frac{K}{Js^2 + Cs}}{1 + \frac{K}{Js^2 + Cs}} = \frac{K}{Js^2 + Cs + K} = \frac{\frac{K}{J}}{s^2 + \frac{C}{J}s + \frac{K}{J}}$$

$$G_{CL} = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad \left\{ \begin{array}{l} \text{2nd order system} \\ \text{Type 0 system} \end{array} \right.$$

$$\omega_n^2 = \frac{K}{J}$$

$$\zeta = \frac{c}{2\sqrt{JK}}$$



- FB has controlled the Type 1 system
- Type 1 system w FB can hold position

9  
fb

```
unit FB example
input data
  K | J | c =
  114   10   4
-----
calculated results

G =

  114
-----
  10 s^2 + 4 s

Continuous-time transfer function.

G_CL =

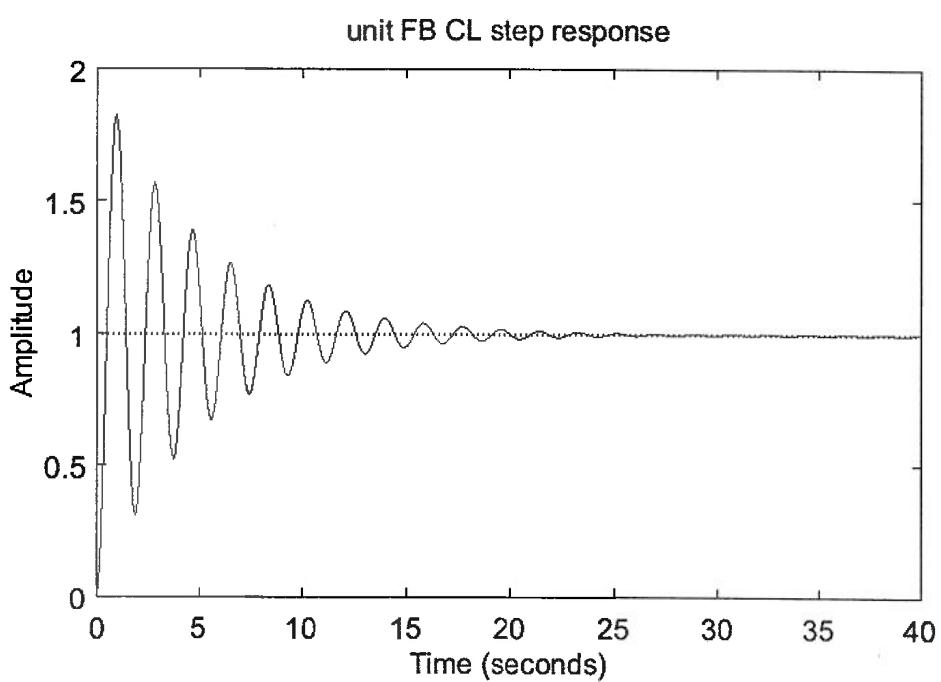
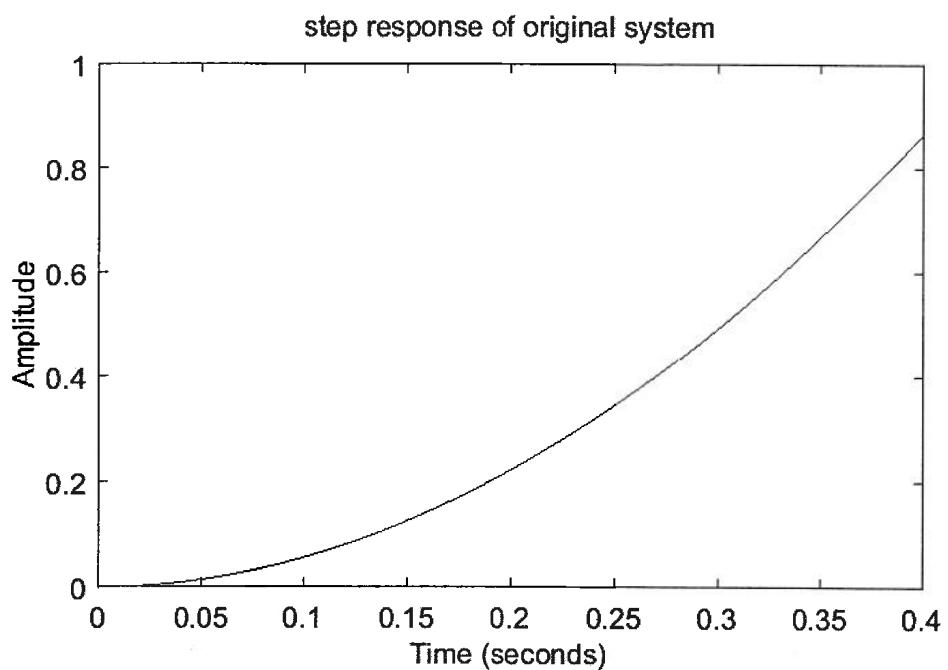
  114
-----
  10 s^2 + 4 s + 114

Continuous-time transfer function.

poles =
-0.2000 + 3.3705i
-0.2000 - 3.3705i

fn,Hz | f,Hz | zeta% =
0.5374  0.5364  5.9235
0.5374  0.5364  5.9235
```

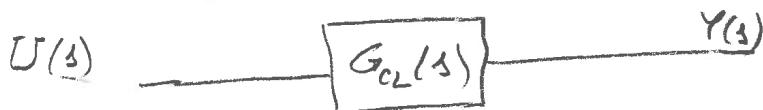
10  
FB



185/523

FB E  
2016/12/02

### Steady state error of feedback systems



Objective: calculate steady state error  $e_{ss}$  of CL system without calculating  $G_{CL}(s)$

Method: Use FVT or  $E(s)$

$$\text{Details: } \varepsilon(s) = U(s) - Y(s) \quad (1)$$

$$Y(s) = G(s)E(s) \quad (2)$$

$$(2) \rightarrow (1) \quad E = U - GE \rightarrow E(s) = \frac{1}{1+G(s)}U(s) \quad (3)$$

FVT:  $\lim_{t \rightarrow \infty} \varepsilon(t) = \lim_{s \rightarrow 0} E(s)$

$$e_{ss} = \lim_{s \rightarrow 0} \frac{s}{1+G(s)} U(s) \quad \text{steady state error (4)}$$

of FB system

SS error depends on:

- input function

$$\left. \begin{array}{l} \text{step } U(s) = \frac{1}{s} \\ \text{ramp } U(s) = \frac{1}{s^2} \end{array} \right\}$$

- system type

$$G(s) = \frac{K}{s^N} \frac{(T_0 s + 1)(T_1 s + 1) \dots}{(T_1 s + 1)(T_2 s + 1) \dots} \quad \left\{ \begin{array}{ll} \text{Type 0, } N=0 \\ \text{Type 1, } N=1 \\ \text{Type 2, } N=2 \end{array} \right.$$

$K = 186/523$

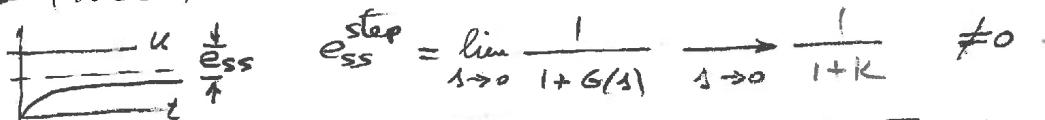
FB<sup>E</sup>Step error of FB systems

Figure of merit: "static position error constant" =  $\lim_{s \rightarrow 0} G(s)$  ←  
 misnomer: in fact, the larger, the better!

• Derivation:  $u_{\text{step}}(s) = \frac{1}{s}$  

$$e_{ss}^{\text{step}} = \lim_{s \rightarrow 0} \frac{s}{1+G(s)} \cdot \frac{1}{s} = \lim_{s \rightarrow 0} \frac{1}{1+G(s)} = \frac{1}{1+\lim_{s \rightarrow 0} G(s)}$$

Type 0 system :  $G(s) = K \frac{(T_0 s + 1)(\dots)}{(T_1 s + 1)(\dots)}$   $\xrightarrow[s \rightarrow 0]{} K = \text{count}$   
 $(N=0)$

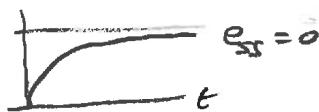


$$e_{ss}^{\text{step}} = \lim_{s \rightarrow 0} \frac{1}{1+G(s)} \xrightarrow[s \rightarrow 0]{} \frac{1}{1+K} \neq 0$$

Feedback creates non-zero ss error for Type 0 sys!

Type 1 system :  $G(s) = \frac{K}{s} \frac{(T_0 s + 1)(\dots)}{(T_1 s + 1)(\dots)} \xrightarrow[s \rightarrow 0]{} \frac{1}{0}$   
 $(N=1)$

$$e_{ss}^{\text{step}} = \lim_{s \rightarrow 0} \frac{1}{1+G(s)} = \frac{1}{1+\frac{1}{0}} = \frac{0}{0+1} = 0$$



Type 2 system :  $G(s) = \frac{K}{s^2} \frac{(T_0 s + 1)(\dots)}{(T_1 s + 1)(\dots)} \xrightarrow[s \rightarrow 0]{} \frac{1}{0}$   
 $(N=2)$

$$e_{ss}^{\text{step}} = \lim_{s \rightarrow \infty} \frac{1}{1+G(s)} = \dots = 0$$

$e_{ss}^{\text{step}} = 0$  for  $N \geq 1$

3  
FBE  
2016/12/02

### Ramp error of FB systems

Figure of merit: "static velocity error constant" =  $\lim_{s \rightarrow 0} s G(s)$  ←  
misnomer: in fact, the larger, the better!

Derivation:  $U_{(s)}^{\text{ramp}} = \frac{1}{s^2}$  

$$e_{ss}^{\text{ramp}} = \lim_{s \rightarrow 0} \frac{s}{1 + G(s)} \frac{1}{s^2} = \lim_{s \rightarrow 0} \frac{1}{s + sG(s)} = \frac{1}{\lim_{s \rightarrow 0} s G(s)}$$

Type 0 Sys

$$s G(s) = s K \frac{(T_a s + 1) \dots}{(T_1 s + 1) \dots} \xrightarrow[s \rightarrow 0]{} 0$$

$$e_{ss}^{\text{ramp}} = \frac{1}{\lim_{s \rightarrow 0} s G(s)} = \frac{1}{0} = \infty \quad \begin{matrix} \text{infinite} \\ \text{error!} \\ \text{cannot follow!} \end{matrix}$$

Type 1 sys

$$s G(s) = s \frac{K}{s} \frac{(T_a s + 1) \dots}{(T_1 s + 1) \dots} \xrightarrow[s \rightarrow 0]{} K = \text{const}$$



$$e_{ss}^{\text{ramp}} = \frac{1}{\lim_{s \rightarrow 0} s G(s)} = \frac{1}{K} \quad \begin{matrix} \text{ramp} \\ \text{offset} \end{matrix}$$

Type 2 sys

$$s G(s) = s \frac{K}{s^2} \frac{(T_a s + 1) \dots}{(T_1 s + 1) \dots} \xrightarrow[s \rightarrow 0]{} \frac{1}{0}$$



$$e_{ss}^{\text{ramp}} = \frac{1}{\lim_{s \rightarrow 0} s G(s)} = 0$$

$$e_{ss}^{\text{ramp}} = 0 \text{ for } N \geq 2$$

<sup>4</sup>  
~~EBG~~  
 2010, 2002

System		Steady State errors	
Type	Expression	Step error $e_{ss}^{step}$	Ramps error $e_{ss}^{ramp}$
0	$K \frac{(T_a s + 1)(\dots)}{(T_1 s + 1)(\dots)}$	$\frac{1}{1+K}$	$\infty$
1	$\frac{K}{s} \frac{(T_a s + 1)(\dots)}{(T_1 s + 1)(\dots)}$	0	$\frac{1}{K}$
2	$\frac{K}{s^2} \frac{(T_a s + 1)(\dots)}{(T_1 s + 1)(\dots)}$	0	0
$\vdots$			
$N > 2$	$\frac{K}{s^N} \frac{(T_a s + 1)(\dots)}{(T_1 s + 1)(\dots)}$	0	0

## 7.4 Stability of feedback control systems and root locus

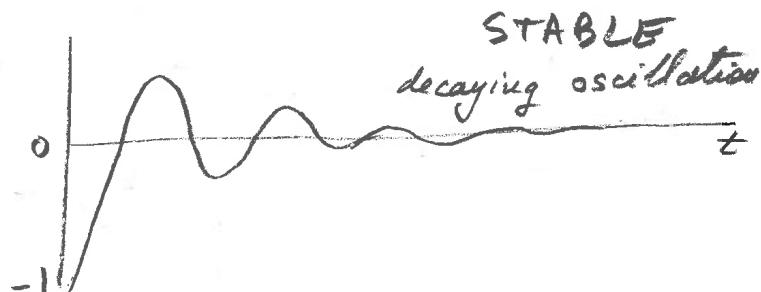
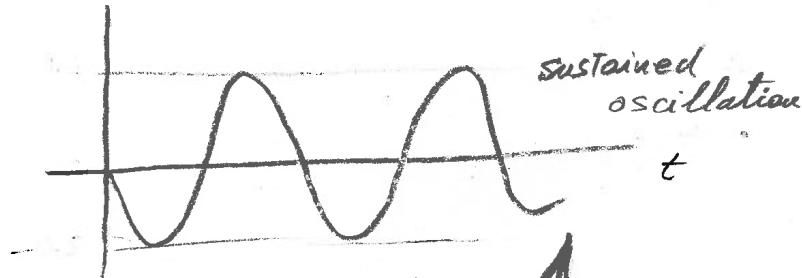
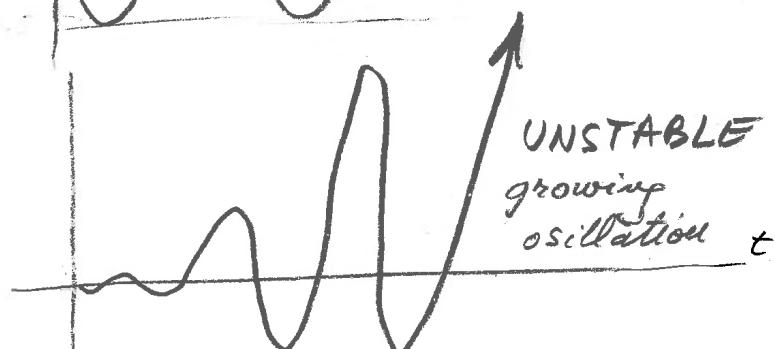
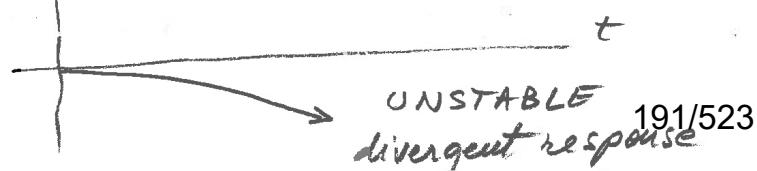
SFB

STABILITYOF FEEDBACK SYSTEMSEXAMPLE

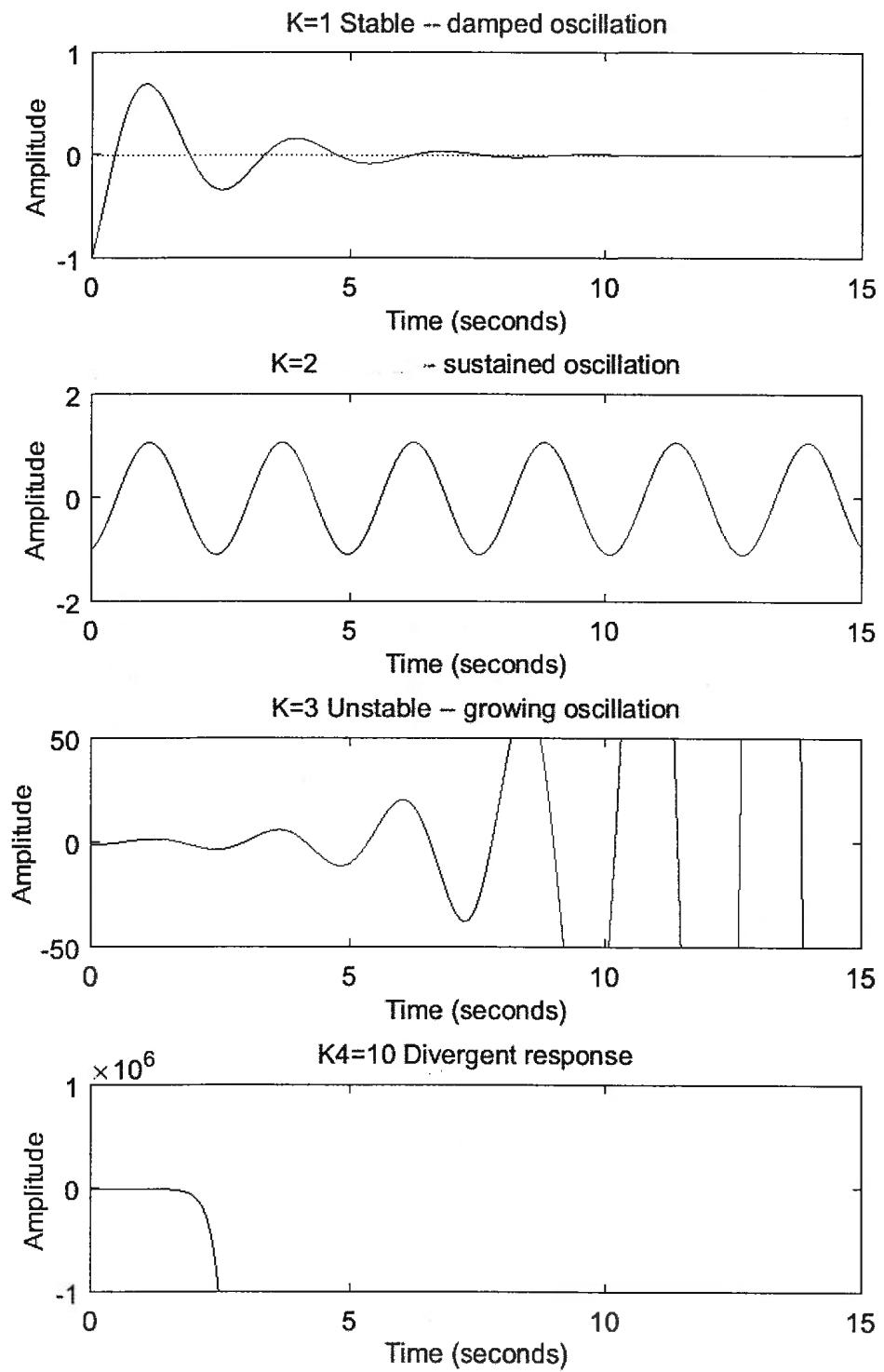
Run MATLAB code FB\_stability\_ex1



$$G(s) = \frac{1-s}{s^2 + 2s + 4}$$

 $K=1$  $K=2$  $K=3$  $K=10$ UNSTABLE  
divergent response 191/523

2  
SFB



3  
SFBExplanation

$$G_{CL} = \frac{G}{1+GK} = \frac{\frac{1-1}{s^2+2s+4}}{1+K \frac{\frac{1-1}{s^2+2s+4}}{s^2+2s+4}}$$

$$= \frac{\frac{1-1}{s^2+2s+4}}{s^2+2s+4 + K(1-1)} = \frac{1-1}{s^2 + (2-K)s + (K+4)}$$

Impulse response:  $X(s) = G_{CL}(s) = \frac{1-1}{s^2 + (2-K)s + (K+4)}$

Response type depends on characteristic eqn.

$$s^2 + (2-K)s + (K+4) = 0 \quad \text{characteristic eqn.}$$

$K=1 \quad s^2 + s + 5 = 0$  LHS  
 $p_{1,2} = \frac{-1 \pm \sqrt{1-4 \times 5}}{2} = -\frac{1}{2} \pm i \frac{\sqrt{19}}{2}$  decaying osc.  
STABLE

$K=2 \quad s^2 + 6 = 0$  imag. axis's  
 $p_{1,2} = \pm i\sqrt{6}$  sustained oscillation

$K=3 \quad s^2 - s + 7 = 0$  RHS  
 $p_{1,2} = \frac{1 \pm \sqrt{1-28}}{2} = \frac{1}{2} \pm i \frac{\sqrt{27}}{2}$  growing osc.  
UNSTABLE

$K=10 \quad s^2 - 8s + 14 = 0$   
 $p_{1,2} = 4 \pm \sqrt{4^2 - 14} = 4 \pm \sqrt{2} \quad \begin{cases} 5.41 & \text{RHS} \\ 2.59 & \text{UNSTABLE} \end{cases}$   
non-oscillatory

4

```
K1 =
    1

G1CL =

    -s + 1
    -----
    s^2 + s + 5
```

Continuous-time transfer function.

```
p1 =
    -0.5000 + 2.1794i
    -0.5000 - 2.1794i
wn1, rad/sec | zeta1 =
    2.2361    0.2236
    2.2361    0.2236
=====
```

```
K2 =
    2
```

```
G2CL =

    -s + 1
    -----
    s^2 + 6
```

Continuous-time transfer function.

```
p2 =
    0.0000 + 2.4495i
    0.0000 - 2.4495i
wn2, rad/sec | zeta2 =
    2.4495      0
    2.4495      0
```

&lt;

K3 =

3

G3CL =

$$\frac{-s + 1}{s^2 - s + 7}$$

Continuous-time transfer function.

p3 =

0.5000 + 2.5981i

0.5000 - 2.5981i

wn3, rad/sec | zeta3 =

2.6458 -0.1890

2.6458 -0.1890

K4 =

10

G4CL =

$$\frac{-s + 1}{s^2 - 8s + 14}$$

Continuous-time transfer function.

p4 =

5.4142

2.5858

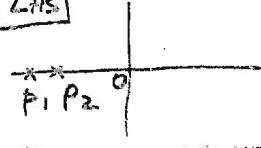
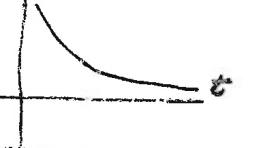
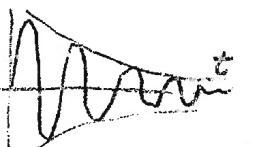
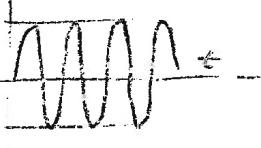
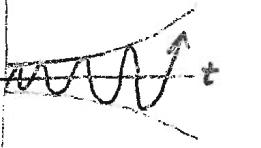
wn4, rad/sec | zeta4 =

5.4142 -1.0000

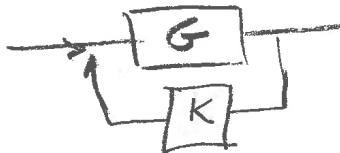
2.5858 -1.0000

6  
SFB

Recall stability analysis as function  
of pole location!

	pole location in complex plane	time response	
1	LHS $\rho_1, \rho_2 < 0$ negative real poles in LHS		monotonic
2	LHS $\rho_1 = \rho_2 < 0$ negative real double pole in LHS		stable
3	LHS $\rho_{1,2} = \sigma \pm i\omega$ $\sigma < 0$ complex poles in LHS		oscillatory
4	$\rho_{1,2} = \pm i\omega$ imaginary poles ( $\sigma = 0$ )		unstable
5	RHS $\rho_{1,2} = \sigma \pm i\omega$ $\sigma > 0$ complex poles in RHS		monotonic
6	RHS $\rho_1 = \rho_2 > 0$ positive real double pole in RHS		unstable
7	RHS $\rho_1 > \rho_2 > 0$ positive real poles in RHS		

20140305

Root LocusGiven  $G(s)$ 

Trace the poles of  $G_{CL}(s)$  as  
K increases

$$G(s) = \frac{B(s)}{A(s)} = \frac{\text{num}(s)}{\text{den}(s)}$$

$$G_{CL} = \frac{G}{1+KG} = \frac{B(s)}{A(s)+KB(s)}$$

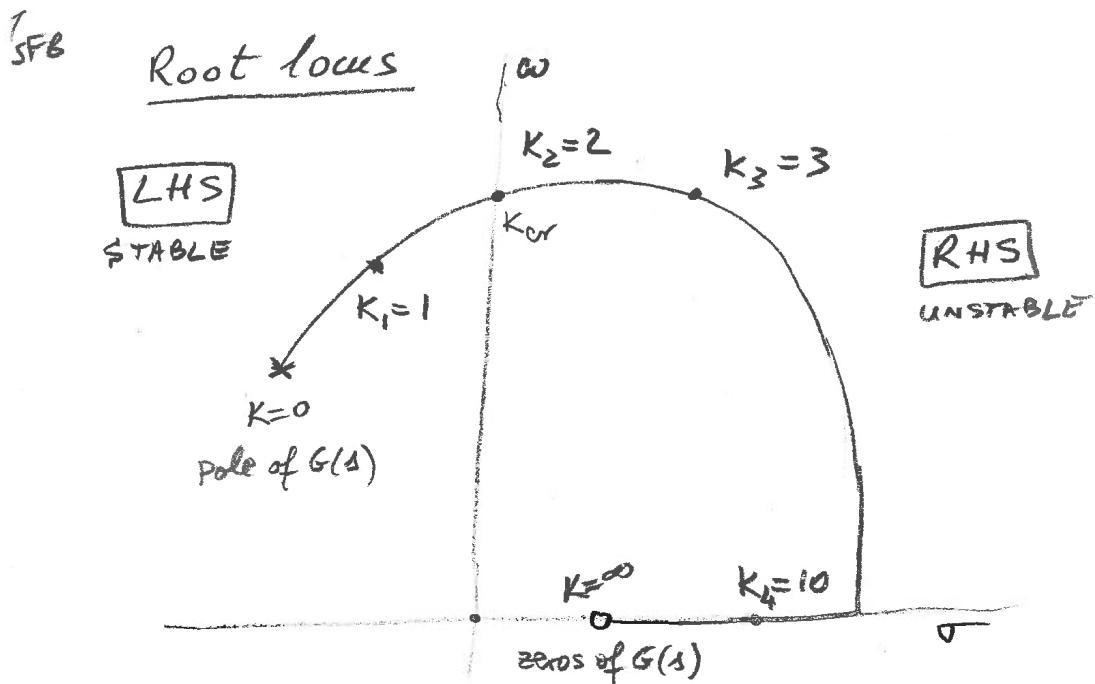
The root locus method looks at the root of the denominator of the CL system  $G_{CL}$

$$A(s)+KB(s) = 0$$

The roots are traced in the s-plane

for  $K=[0, \infty)$

For  $K=0$ , the poles of  $G_{CL}$  are the same as the poles of  $G$ .



$$\rho_{1,2} = \sigma \pm i\omega_d = -\zeta\omega_n \pm j\omega_d$$

$$x(t) = C e^{-\zeta\omega_n t} \sin(\omega_d t + \varphi)$$

$$K_1 : \sigma < 0 \rightarrow \zeta > 0$$

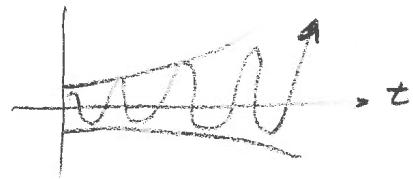


$$K_2 : \sigma = 0 \rightarrow \zeta = 0$$

$K_{cr}$



$$K_3 : \sigma > 0 \rightarrow \zeta < 0$$



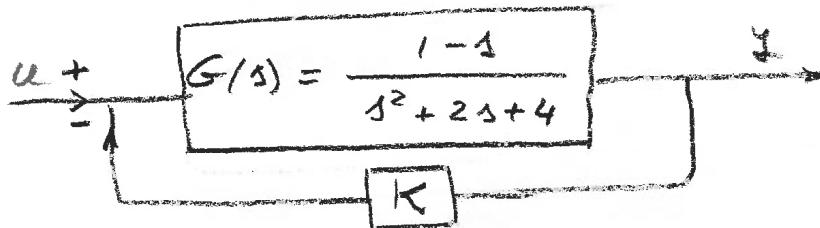
$$K_4 : \omega = 0$$

$$\rho_1 = \sigma_1, \rho_2 = \sigma_2 > 0$$

$$x(t) = C_1 e^{\rho_1 t} + C_2 e^{\rho_2 t}$$



SFB

MATLAB root locus function

System: G

Gain: 2  $K_1 = 2$ Pole:  $-0.00116 + 2.45i$ 

Damping: 0.000474

Overshoot (%): 99.9

Frequency (rad/s): 2.45

~~unstable~~  $\rightarrow$ 

K vs Root Locus

System: G

Gain: 1.01  $K_2 = 1$ Pole:  $-0.494 + 2.18i$ 

Damping: 0.221

Overshoot (%): 49.1

Frequency (rad/s): 2.24

System: G

Gain: 3  $K_3 = 3$ Pole:  $0.5 + 2.6i$ 

Damping: -0.189

Overshoot (%): 183

Frequency (rad/s): 2.64

$$\phi = \sigma + i\omega$$

$$\sigma > 0$$

*unstable oscillation*

~~unstable~~  $\rightarrow$

 $\phi = \sigma$ 

Imaginary Axis (seconds)

0

-1

-2

-3

*stable oscillation*

$\phi = \sigma + i\omega$

$\sigma < 0$

*divergence*

System: G

Gain: 10

 $K_4 = 10$ 

Pole: 2.57

Damping: -1

Overshoot (%): 0

Frequency (rad/s): 2.1

Real Axis (seconds<sup>-1</sup>)

7  
SF&

## rlocus

Evans root locus

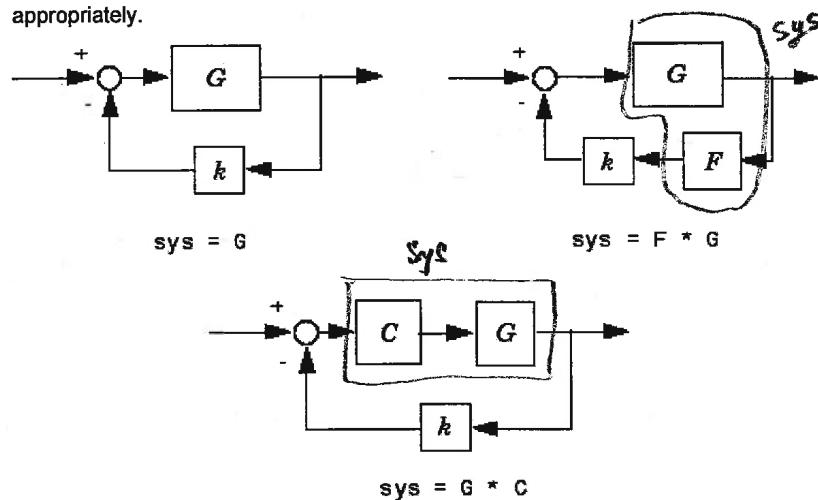
### Syntax

- rlocus
- rlocus(sys)
- rlocus(sys1,sys2,...)

### Description

rlocus computes the Evans root locus of a SISO open-loop model. The root locus gives the closed-loop pole trajectories as a function of the feedback gain  $k$  (assuming negative feedback). Root loci are used to study the effects of varying feedback gains on closed-loop pole locations. In turn, these locations provide indirect information on the time and frequency responses.

rlocus(sys) calculates and plots the root locus of the open-loop SISO model sys. This function can be applied to any of the following *negative* feedback loops by setting sys appropriately.



If sys has transfer function

$$h(s) = \frac{n(s)}{d(s)}$$

the closed-loop poles are the roots of

$$d(s) + k n(s) = 0$$

rlocus adaptively selects a set of positive gains  $k$  to produce a smooth plot. Alternatively,

10  
FB

```
rlocus(sys, k)
```

uses the user-specified vector  $k$  of gains to plot the root locus.

`rlocus(sys1, sys2, ...)` draws the root loci of multiple LTI models  $sys1$ ,  $sys2$ , ... on a single plot. You can specify a color, line style, and marker for each model, as in

```
rlocus(sys1, 'r', sys2, 'y:', sys3, 'gx').
```

When invoked with output arguments,

```
[r, k] = rlocus(sys)
r = rlocus(sys, k)
```

return the vector  $k$  of selected gains and the complex root locations  $r$  for these gains. The matrix  $r$  has  $\text{length}(k)$  columns and its  $j$ th column lists the closed-loop roots for the gain  $k(j)$ .

### Remarks

You can change the properties of your plot, for example the units. For information on the ways to change properties of your plots, see [Ways to Customize Plots](#).

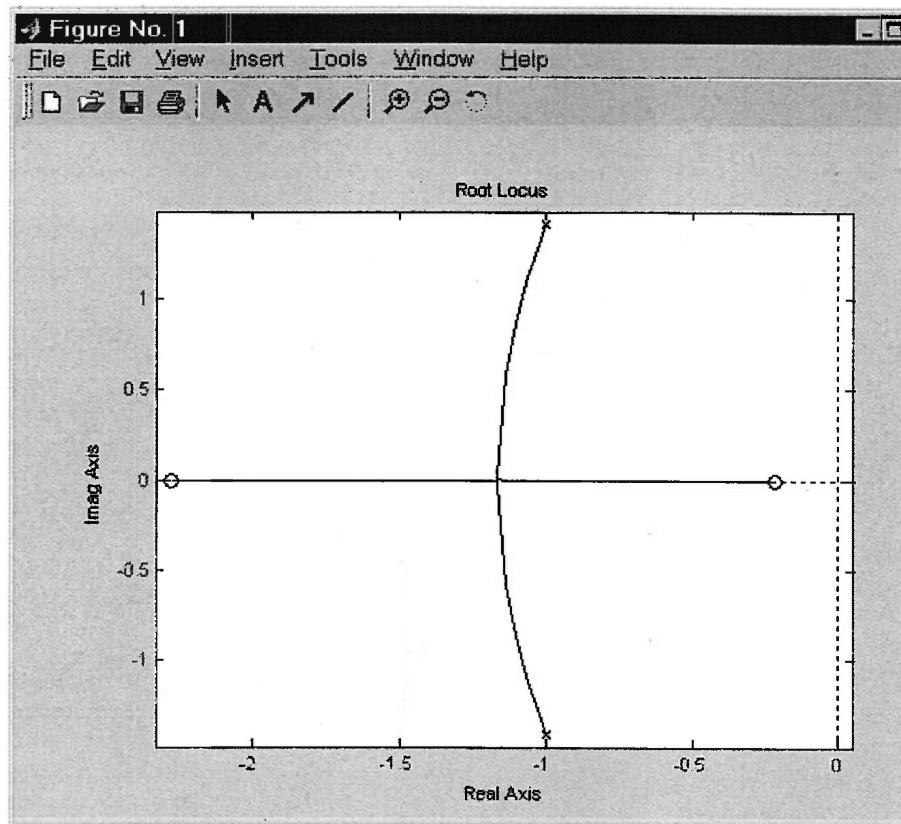
### Example

Find and plot the root-locus of the following system.

$$h(s) = \frac{2s^2 + 5s + 1}{s^2 + 2s + 3}$$

```
h = tf([2 5 1], [1 2 3]);
rlocus(h)
```

11  
SFB



You can use the right-click menu for rlocus to add grid lines, zoom in or out, and invoke the Property Editor to customize the plot. Also, click anywhere on the curve to activate a data marker that displays the gain value, pole, damping, overshoot, and frequency at the selected point.

### See Also

[pole](#), [pzmap](#)

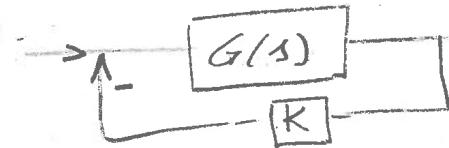
[Provide feedback about this page](#)

[reshape](#)

[rlocusplot](#)

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[Acknowledgments](#)

<sup>IV</sup>  
SFB ROOT LOCUS METHOD  $G(s) = \frac{B(s)}{A(s)} = \frac{\text{num}}{\text{den}}$



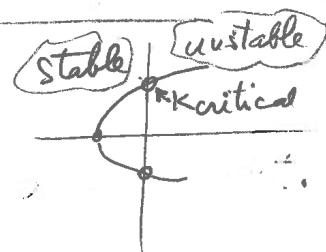
$$G_{CL} = \frac{KG}{1+KG}$$

$$G(s) = \frac{(s+z_1)(s+z_2) \dots (s+z_m)}{(s+p_1)(s+p_2) \dots (s+p_n)} = \frac{\text{num}}{\text{den}}$$

$-z_1, \dots, -z_m$  zeros of open loop transfer function  
 $-p_1, \dots, -p_n$  poles

charact. eq<sup>n</sup>:  $1 + KG = 0$ .

$$A(s) + KB(s) = 0$$



$$(s+p_1)(s+p_2) \dots (s+p_n) + K(s+z_1) \dots (s+z_m) = 0.$$

selection of this equation gives the  $\text{CL}$  poles  
 $s = \sigma + i\omega$

Root loci: trajectory of these poles as

$$K = 0 \rightarrow \infty$$

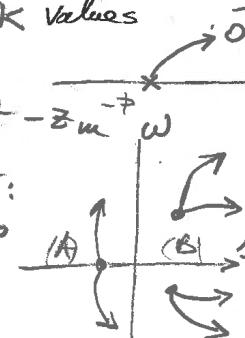
rlocus (num, den)  $\rightarrow$  automatic K generation  
 $K \in [0, \infty)$

rlocus (num, den, K)  $\rightarrow$  must give K values

$K=0$ : roots are the OL poles:  $-p_1, -p_2, \dots, -p_n$   
 $K=\infty$ : roots are the OL zeros:  $-z_1, -z_2, \dots, -z_m$

Breakaway: multiple roots branch out:

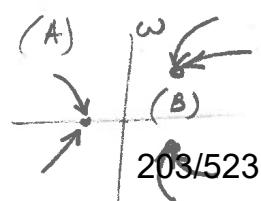
- (A) from real poles on real axis
- (B) from conjugate poles



Break in: branches coalesce into multiple roots

(A) into real poles

(B) into conjugate poles



<sup>13</sup>  
SFBAngle condition

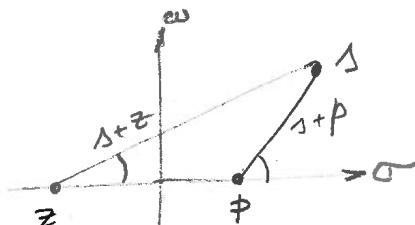
$$1 + KG = 0$$

$$KG = -1$$

$$\angle KG = \angle -1 = \pm 180^\circ (2k+1), k=0, 1, \dots$$

$$\angle KG = \angle K \frac{(s+z_1)(s+z_2) \dots}{(s+p_1)(s+p_2) \dots}$$

$$= \cancel{\angle s+z_1} + \cancel{\angle s+z_2} + \dots - \cancel{\angle s+p_1} - \cancel{\angle s+p_2} - \cancel{\angle s+p_3} \dots = \pm 180^\circ (2k+1).$$

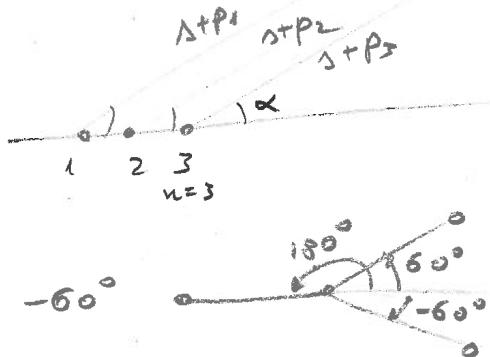
Asymptotes

$$s \rightarrow \infty$$

• angles of asymptotes

$$\text{ang asympt} = \frac{\pm 180^\circ (2k+1)}{3}$$

$$= 60^\circ ; 180^\circ, -60^\circ$$



• axis crossing of asymptotes

$$\text{SFB } (s+p_1)(s+p_2)\dots + K(s+z_1)(s+z_2)\dots = 0$$

$$K=0 \rightarrow (s+p_1)(s+p_2)\dots(s+p_n) = 0$$

roots are the OL poles



$$K \rightarrow \infty : K(s+z_1)(s+z_2)\dots(s+z_m) = 0$$

roots are the OL zeros

## 7.5 Stability Criteria

RH  
20140306

## ROUTH CRITERION

### FOR HIGHER ORDER POLYNOMIALS

Given:  $A(s) = a_n s^n + a_{n-1} s^{n-1} + \dots + a_0$

Find if any roots of  $A(s)$  are in the RHS

Solution by Routh criterion

(1) If any of the coefficients  $a_0, a_1, \dots, a_n$  is negative, then at least one root is in RHS and the system is UNSTABLE. **STOP**

(2) If all coefficients are positive, do the Routh table:

2  
RHR Table

$$A(s) = a_n s^n + a_{n-1} s^{n-1} + a_{n-2} s^{n-2} + \dots + a_1 s + a_0 = 0$$

$s^n$	$a_n$	$a_{n-2}$	$a_{n-4}$	$\dots$
$s^{n-1}$	$a_{n-1}$	$a_{n-3}$	$a_{n-5}$	$\dots$
$s^{n-2}$	$b_1$	$b_2$	$b_1 = \frac{a_{n-1} a_{n-2} - a_n a_{n-3}}{a_{n-1}}$	
$s^{n-3}$	$c_1$	$c_2$	$b_2 = \frac{a_{n-1} a_{n-4} - a_n a_{n-5}}{a_{n-1}}$	
$\vdots$			$c_1 = \frac{b_1 a_{n-3} - a_{n-1} b_2}{b_1}$	$\vdots$
$\vdots$				

Count the number of sign changes and zeros in the first column to get the number of roots in RHS

use MATLAB file:

Routh\_Hurwitz\_Stability\_criterion\_modified\_vg1.m

Routh criterion Example VGI

$$G(s) = \frac{4s+2}{s^3 + 3s^2 + 4s + 2} = \frac{B(s)}{A(s)}$$

Examine  $A(s) = s^3 + 3s^2 + 4s + 2$

$$\begin{array}{ll} a_3 = 1 & a_1 = 4 \\ a_2 = 3 & a_0 = 2 \end{array}$$

$s^3$	1	4	$b_1 = \frac{3 \times 4 - 1 \times 2}{3} = 10/3$
$s^2$	3	2	$b_2 = \frac{0 - 0}{3} = 0$
$s^1$	$10/3$		$c_1 = \frac{\frac{10}{3} \times 2 - 3 \times 0}{10/3} = 2$
$s^0$	2		

Routh criterion: number of sign changes = zero (0)  
 i.e. No root in RHS  $\rightarrow$  STABLE!

Verification by MATLAB

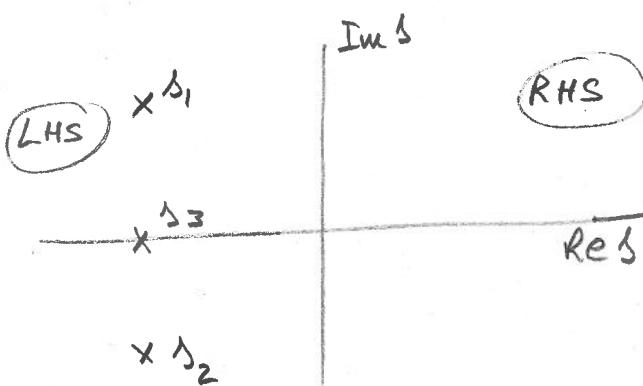
$$A = [1 \ 3 \ 4 \ 2]$$

$\text{roots}(A)$ :

$$\lambda_1 = -1 + i$$

$$\lambda_2 = -1 - i$$

$$\lambda_3 = -1$$



All roots are in LHS  $\rightarrow$  System is STABLE!

4 RT  
2014 03 03

## R Criterion

## Example VG 2

$$G(s) = \frac{31}{s^3 + 5s^2 + 6s + 31}$$

$$a_3 = 1 \quad a_2 = 5 \quad a_1 = 6 \quad a_0 = 31$$

$$\begin{array}{c|cc} s^3 & 1 & 6 \\ s^2 & 5 & 31 \\ \hline s^1 & -0.2 \\ s^0 & 31 \end{array} \quad L_1 = \frac{5 \times 6 - 1 \times 31}{5} = -0.2$$

$$L_2 = \frac{-0.2 \times 31}{-0.2} = 31$$

→ 2 sign changes  
2 roots in RHS → unstable !

See Matlab print out on next page .

5  
RH  
20140303

## MATLAB Command Window

Input coefficients of characteristic equation  
 $[a_n a_{n-1} a_{n-2} \dots a_0] = [1 5 6 31]$

-----  
 Roots of characteristic equation is:

ans =

$p_1$  -5.0319  
 $p_2$  0.0160 + 2.4820i  
 $p_3$  0.0160 - 2.4820i

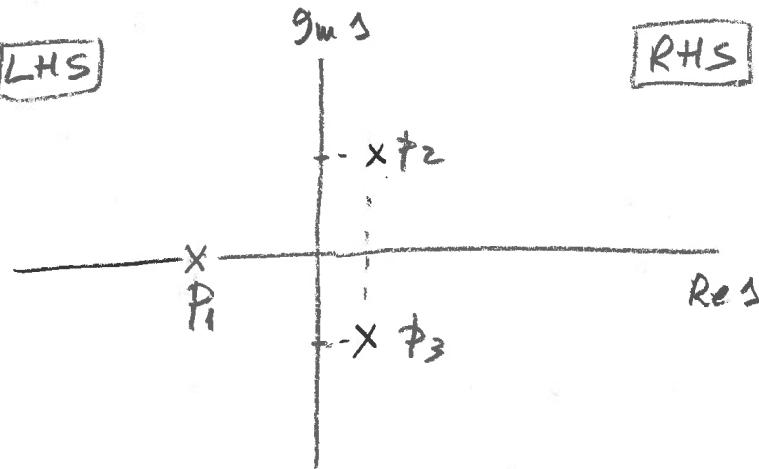
-----The R array is:-----

m =

1.0000	6.0000
5.0000	31.0000
-0.2000	0
31.0000	0

----> System is Unstable <----

>>



5a  
R criterion Example VG3

$$G(s) = \frac{5s^2 + 8s + 3}{s^6 + 3s^5 - s^4 - 7s^3 + 10s^2 + 14s - 20}$$

$A(s)$  has some negative coefficients  
 At least one root of  $A(s)$  is in RHS  
 system is unstable

<sup>6</sup>  
R.H.S.R criterion Example VG4

$$G(s) = \frac{5s^2 + 8s + 3}{s^6 + 3s^5 + s^4 + 7s^3 + 10s^2 + 14s + 20}$$

$$\begin{array}{llll} a_6 = 1 & a_4 = 1 & a_2 = 10 & a_0 = 20 \\ a_5 = 3 & a_3 = 7 & a_1 = 14 & \end{array}$$

$s^6$	1	1	10	20	
$s^5$	3	7	14		Two (2) sign changes
$\textcircled{1} \rightarrow s^4$	-1.33	5.33	-20		2 roots in RHS
$\textcircled{2} \rightarrow s^3$	19	59			UNSTABLE!
$s^2$	9.4737	20			
$s^1$	18.8889				
$s^0$	20				

MATLAB

$$A = [1 \ 3 \ 1 \ 7 \ 10 \ 14 \ 20]$$

roots(A):

$$-3.14 + 0i$$

$$-1.30 + 0i$$

LHS

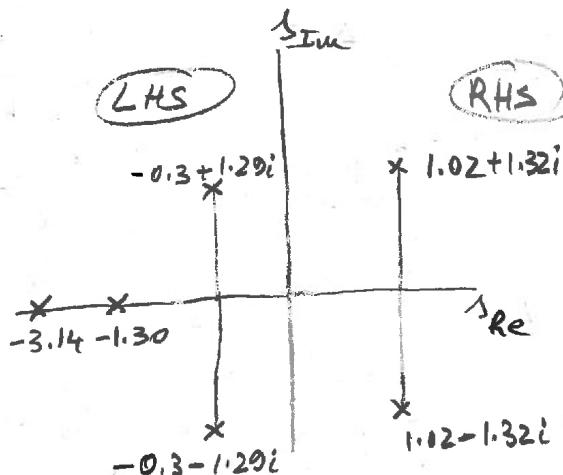
$$-0.30 + 1.29i$$

$$-0.30 - 1.29i$$

RHS

$$1.02 + 1.32i$$

$$1.02 - 1.32i$$



See book

for more about Routh criterion

$\gamma$   
RH

Input coefficients of characteristic equation  
 $[a_n \ a_{n-1} \ a_{n-2} \dots \ a_0] = [1 \ 3 \ 1 \ 7 \ 10 \ 14 \ 20]$

-----  
Roots of characteristic equation are:

ans =

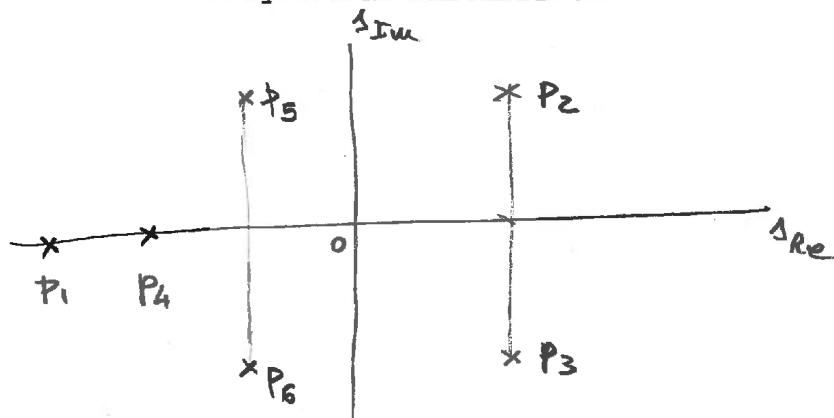
$p_1$	-3.1462 + 0.0000i	LHS
$p_2$	1.0202 + 1.3244i	RHS
$p_3$	1.0202 - 1.3244i	RHS
$p_4$	-1.2962 + 0.0000i	LHS
$p_5$	-0.2990 + 1.2905i	LHS
$p_6$	-0.2990 - 1.2905i	LHS

-----The Routh array is:-----

m =

1.0000	1.0000	10.0000	20.0000
3.0000	7.0000	14.0000	0
-1.3333	5.3333	20.0000	0
19.0000	59.0000	0	0
9.4737	20.0000	0	0
16.5889	0	0	0
20.0000	0	0	0

----> System is Unstable <----



8  
RH

(page 116)

#### 5.4 HURWITZ STABILITY CRITERION

The Hurwitz criterion is another method for determining whether all the roots of the characteristic equation of a continuous system have negative real parts. This criterion is applied using determinants formed from the coefficients of the characteristic equation. It is assumed that the first coefficient,  $a_n$ , is positive. The determinants  $\Delta_i$ ,  $i=1, 2, \dots, n-1$ , are formed as the principal minor determinants of the determinant

$$\Delta_n = \begin{vmatrix} a_{n-1} & a_{n-3} & \cdots & \begin{cases} a_0 & \text{if } n \text{ odd} \\ a_1 & \text{if } n \text{ even} \end{cases} & 0 & \cdots & 0 \\ a_n & a_{n-2} & \cdots & \begin{cases} a_1 & \text{if } n \text{ odd} \\ a_0 & \text{if } n \text{ even} \end{cases} & 0 & \cdots & 0 \\ 0 & a_{n-1} & a_{n-3} & \cdots & \cdots & \cdots & 0 \\ 0 & a_n & a_{n-2} & \cdots & \cdots & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ 0 & \cdots & \cdots & \cdots & \cdots & \cdots & a_0 \end{vmatrix}$$

The determinants are thus formed as follows:

$$\Delta_1 = a_{n-1}$$

$$\Delta_2 = \begin{vmatrix} a_{n-1} & a_{n-3} \\ a_n & a_{n-2} \end{vmatrix} = a_{n-1}a_{n-2} - a_n a_{n-3}$$

$$\Delta_3 = \begin{vmatrix} a_{n-1} & a_{n-3} & a_{n-5} \\ a_n & a_{n-2} & a_{n-4} \\ 0 & a_{n-1} & a_{n-3} \end{vmatrix} = a_{n-1}a_{n-2}a_{n-3} + a_n a_{n-1}a_{n-5} - a_n a_{n-3}^2 - a_{n-4}a_{n-1}^2$$

and so on up to  $\Delta_{n-1}$ .

**Hurwitz Criterion:** All the roots of the characteristic equation have negative real parts if and only if  $\Delta_i > 0$ ,  $i = 1, 2, \dots, n$ .

**EXAMPLE 5.5.** For  $n = 3$ ,

$$\Delta_3 = \begin{vmatrix} a_2 & a_0 & 0 \\ a_3 & a_1 & 0 \\ 0 & a_2 & a_0 \end{vmatrix} = a_2 a_1 a_0 - a_0^2 a_3, \quad \Delta_2 = \begin{vmatrix} a_2 & a_0 \\ a_3 & a_1 \end{vmatrix} = a_2 a_1 - a_0 a_3, \quad \Delta_1 = a_2$$

Thus all the roots of the characteristic equation have negative real parts if

$$a_2 > 0 \quad a_2 a_1 - a_0 a_3 > 0 \quad a_2 a_1 a_0 - a_0^2 a_3 > 0$$

9  
 R<sup>2</sup>  
 2016/12/01

Q: What is R criterion?

A: R criterion is a tabular method to determine if a polynomial

$$A(s) = a_n s^n + a_{n-1} s^{n-1} + \dots + a_1 s + a_0$$

has roots in the RHS

Q: What is R criterion good for?

A: R criterion is used to evaluate the stability of a system  $G(s) = \frac{B(s)}{A(s)}$

If  $A(s)$  has roots in RHS, then the system is UNSTABLE

Q: How do I use R criterion?

A: - If  $A(s)$  has at least one -ve coefficient, then the system is UNSTABLE. **STOP**

- If all coefficients of  $A(s)$  are +ve, then do R Table to find if any roots are in RHS

10  
R<sup>st</sup>

Q : What is H criterion?

A : H criterion is another method of using the coefficients of a polynomial equation to determine whether all the roots have negative real parts

Q : How is H criterion different from R criterion?

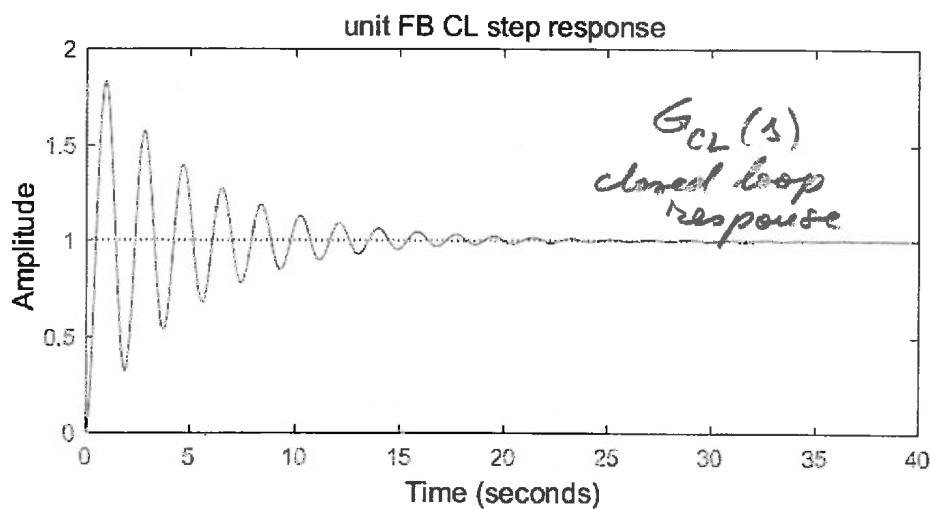
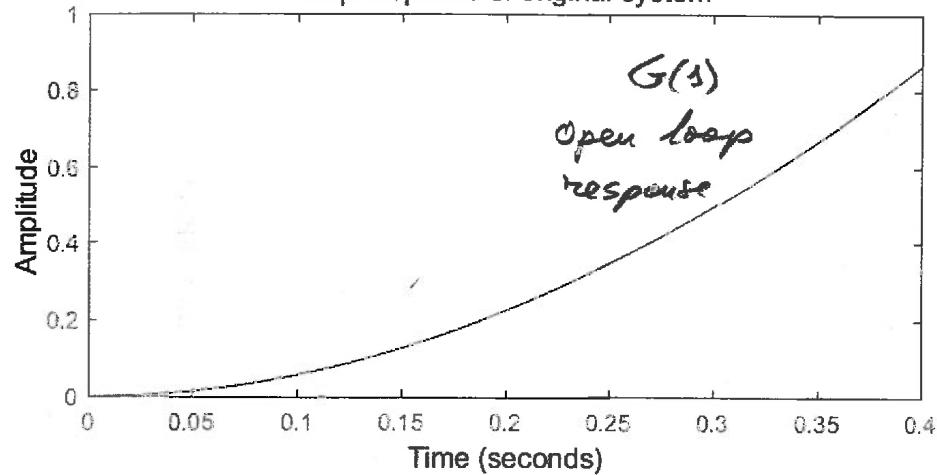
A : H criterion uses determinants, whereas R criterion uses a table

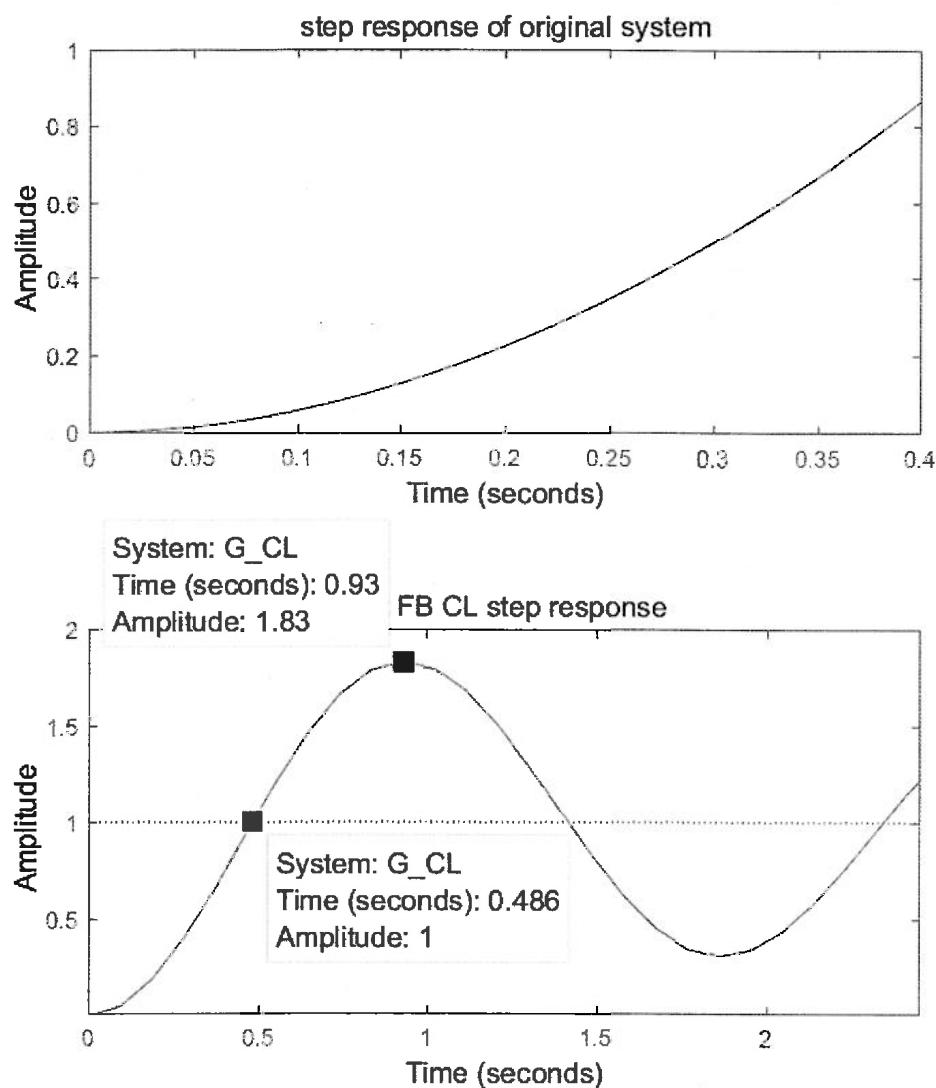
## 7.6 Feedback Controllers

Aircraft roll  
step response

$G(s) = \frac{114}{10s^2 + 4s}$

step response of original system



D6  
C

Measured:  $t_{\frac{1}{2}} = 0.486 \text{ sec}$

$M_p = 83\%$

Design Specs  
DS1:  $t_{\frac{1}{2}} \leq 1.5 \text{ sec}$

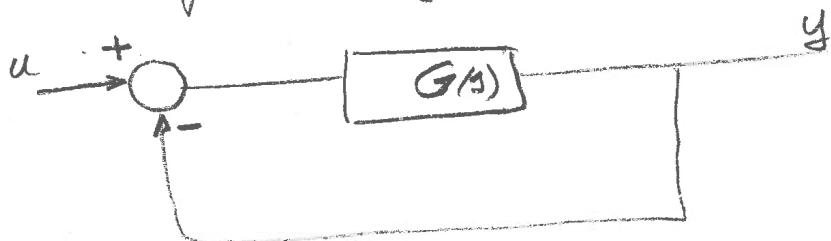
DS2:  $M_p \leq 20\%$

DS1: satisfied

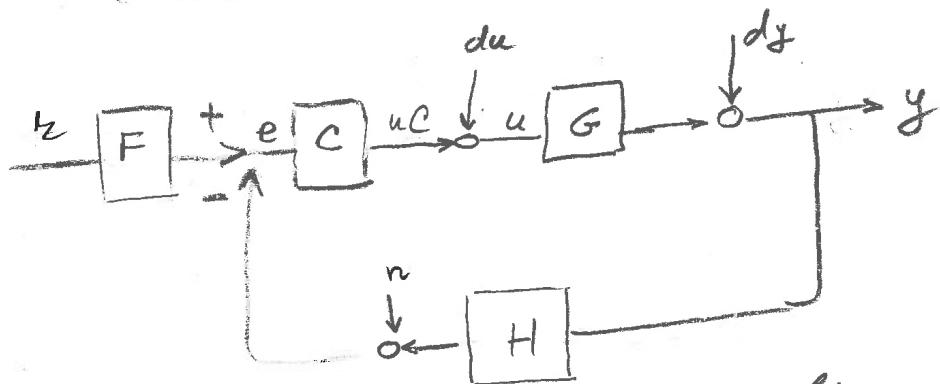
DS2: NOT satisfied

*C  
2014 03 17* CONTROLLERS

Basic feedback system



Enhanced FB control systems



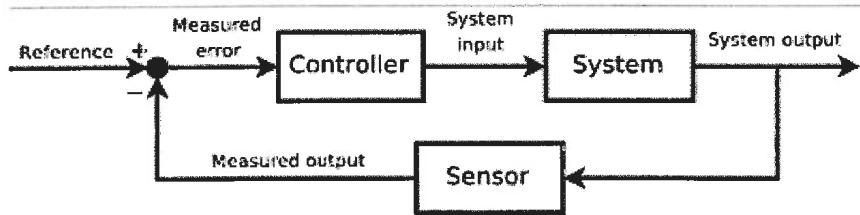
Addition "boxes" with specific transfer functions may be added to improve performance

Examples:

- PID controllers
- Filters
- Compensators
- Pre-filters
- Post-filters

2  
C

## CONTROLLERS



Controllers modulate the feedback error to improve the performance of the feedback control system.

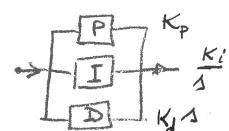
$a_c$ Filters

$$G_c(s) = \frac{1}{T_1 s + 1} \quad (\text{low pass filter})$$

$$G_c(s) = \frac{T_1}{T_1 s + 1} \quad (\text{high pass filter})$$

Compensators

$$G_c(s) = \frac{T_a s + 1}{T_1 s + 1}$$

PID controllers

$$G_c = K_p + \frac{K_i}{s} + K_d s$$

$$G_c = K \left( 1 + \frac{1}{T_i s} + T_d s \right)$$

t-domain    s-domain

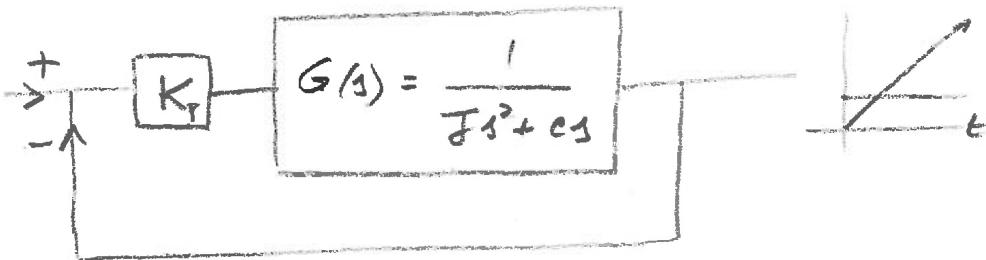
Comments:

	t-domain	s-domain		
P (proportional)	K	K	<ul style="list-style-type: none"> <li>simplest</li> <li>adds stiffness, increases frequency</li> </ul>	
I (integral)	$\int dt$	$\frac{K_i}{s}$	<ul style="list-style-type: none"> <li>eliminates offsets</li> <li>increases system order</li> <li>may be unstable</li> </ul>	"under compensator"
D (derivative)	$d/dt$	$K_d s$	<ul style="list-style-type: none"> <li>adds damping</li> <li>decreases system order</li> <li>increases sensitivity</li> <li>used as PD or PID</li> </ul>	"anticipates & corrects the error"

Note : D control is not physically realizable. It is done as  $\frac{Ns}{s+N} \xrightarrow{N \gg 1} s$  224/523

<sup>8</sup> P control of Type I system.

Recall:



$$G_{CL}(s) = \frac{K_p}{Js^2 + cs + K_p}$$

$$G_{CL}(s) = \frac{K_p}{Js^2 + cs + K_p} = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

$$\omega_n^2 = \frac{K_p}{J}, \quad \omega_n = \sqrt{\frac{K_p}{J}}$$

$$\zeta = \frac{c}{2\sqrt{J}K_p}$$

$K$  = factor of proportionality; "P-control"  
Modification of  $K_p$  can modify  
frequency  $\omega_n$  and damping  $\zeta$

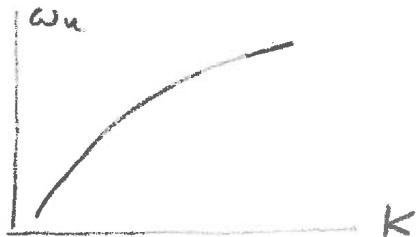
<sup>2</sup>

P-control of frequency and damping

$K_p$  = factor of proportionality  $\rightarrow$  P-control

Modification of  $K_p$  can modify  
the frequency  $\omega_n$  and the damping  $\zeta$

$$\omega_n^2 = \frac{K_p}{J}$$



Frequency increases with  $K$

$$\zeta = \frac{c}{2\sqrt{J}K_p}$$



Damping decreases with  $K_p$

$$\text{Critical damping: } \zeta = 1 \rightarrow \frac{c}{2\sqrt{J}K_{cr}} = 1 : K_{cr} = \frac{c^2}{4J}$$

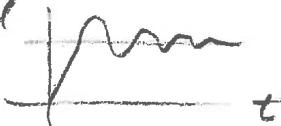
$0 < K_p \leq K_{cr}$ , overdamped

converging exponential response.



$K_p < K_{cr}$

underdamped  
oscillatory response

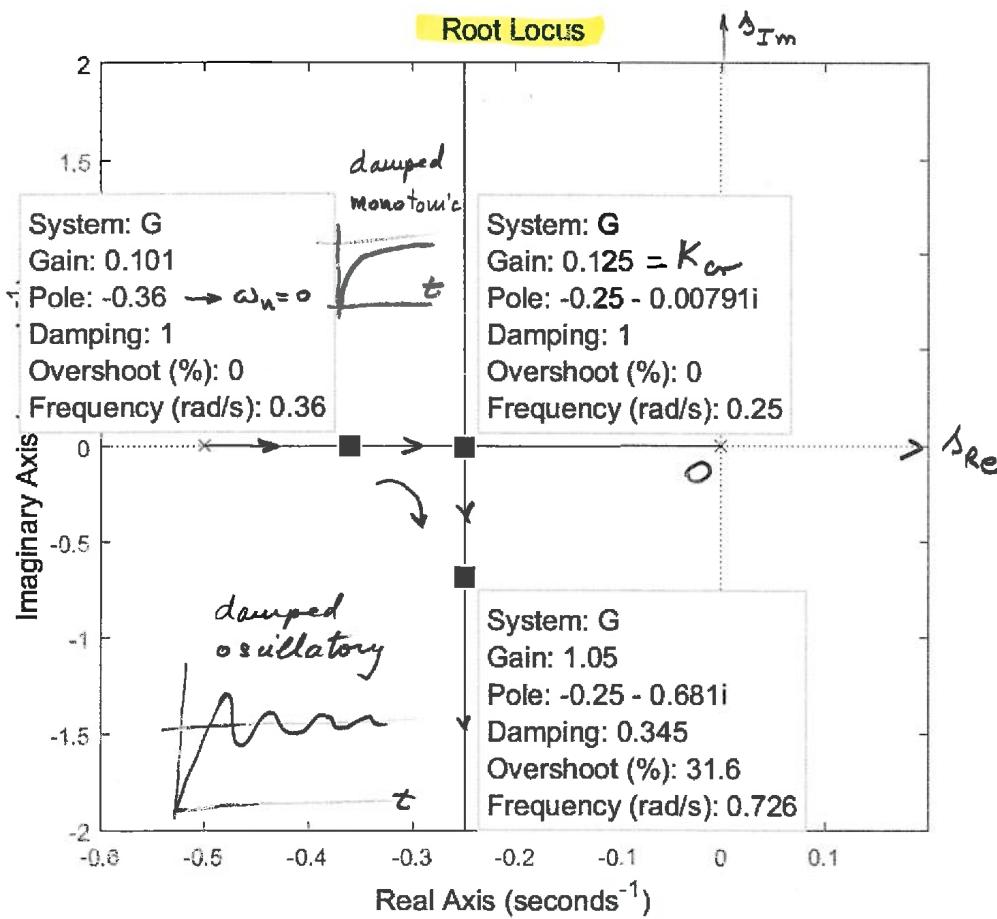


3<sup>rd</sup>

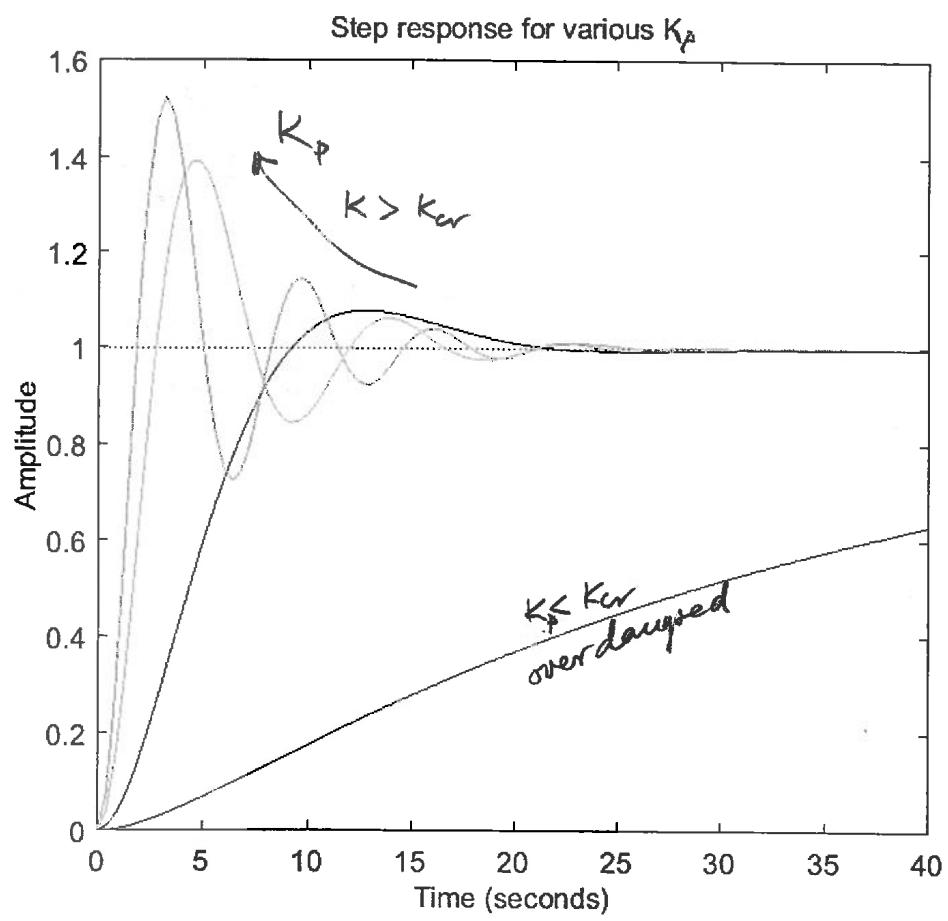
$$G(s) = \frac{1}{2s^2 + s}$$

$$K_{cr} = \frac{c^2}{4J} = \frac{1}{8}$$

$$= 0.125$$



$K_p$   
P-CONTROL  
Type 1 sys



$s_p$ 

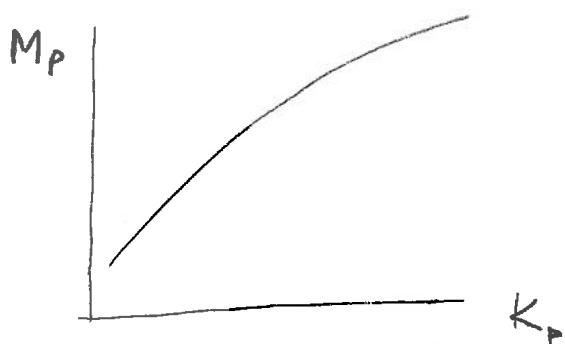
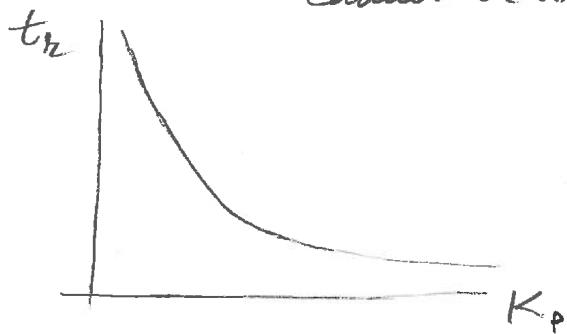
P-control of  $t_r$  and  $M_p$

Recall:

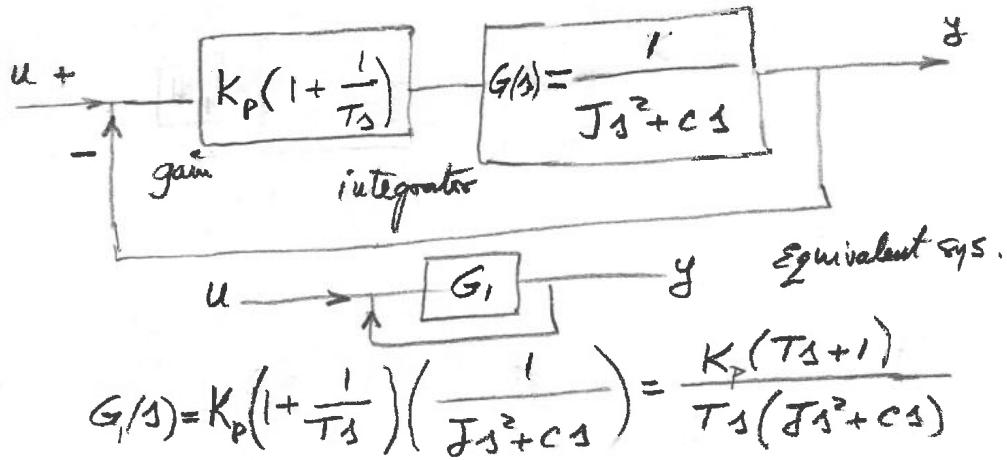
$$\text{rise time } t_r = \frac{\pi - \varphi}{\omega_d} = \frac{\pi - \varphi}{\omega_n \sqrt{1-\xi^2}}, \quad \varphi = \sin^{-1} \sqrt{1-\xi^2}$$

$$\text{overshoot } M_p = e^{-\frac{\pi}{\sqrt{1-\xi^2}}}.$$

Small overshoot  $M_p$  and small rise time  $t_r$   
cannot be simultaneously met!



Run MATLAB program

$\frac{1}{s}$ PI Control Principle

$$G_{CL} = \frac{G_1}{1 + G_1} = \frac{K_p(T_s + 1)}{(T_J)s^3 + (T_C)s^2 + (K_p T)s + K_p}$$

, 3<sup>rd</sup> order system

- May become unstable
- One could use R-H criterion to predict critical T for instability, i.e.,

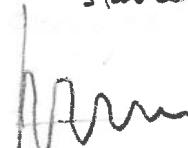
$$T > T_{cr} = J/C \quad (\text{see next page})$$

Example :  $J=2$ ,  $C=1$ ,  $T_{cr} = \frac{2}{1}=2$

(See results next page)

$$T=4$$

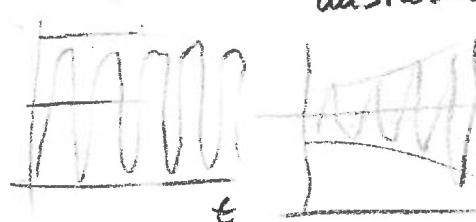
stable



$$T=T_{cr}=2$$

$$T=1$$

unstable



230/523

<sup>10</sup>  
<sup>PS</sup> R.H (Routh-Hurwitz) Criterion Table.

$$(T\mathcal{J})s^3 + (Tc)s^2 + (K_p T)s + K = 0$$

$s^3$	$T\mathcal{J}$	$K_p T$	
$s^2$	$Tc$	$K_p$	
$s^1$	$b_1$		$b_1 = \frac{(Tc)(K_p T) - (T\mathcal{J})K_p}{Tc} = \frac{(CT - \mathcal{J})K_p}{C}$
$s^0$	$K_p$		$c_1 = \frac{b_1 K_p}{b_1} = K_p$

Discussions

$s^3$	$T\mathcal{J}$	+ve
$s^2$	$Tc$	+ve
$s^1$	$b_1$	may be +ve or -ve
$s^0$	$K_p$	+ve

If  $b_1$  is -ve, then sign change; i.e., INSTABILITY

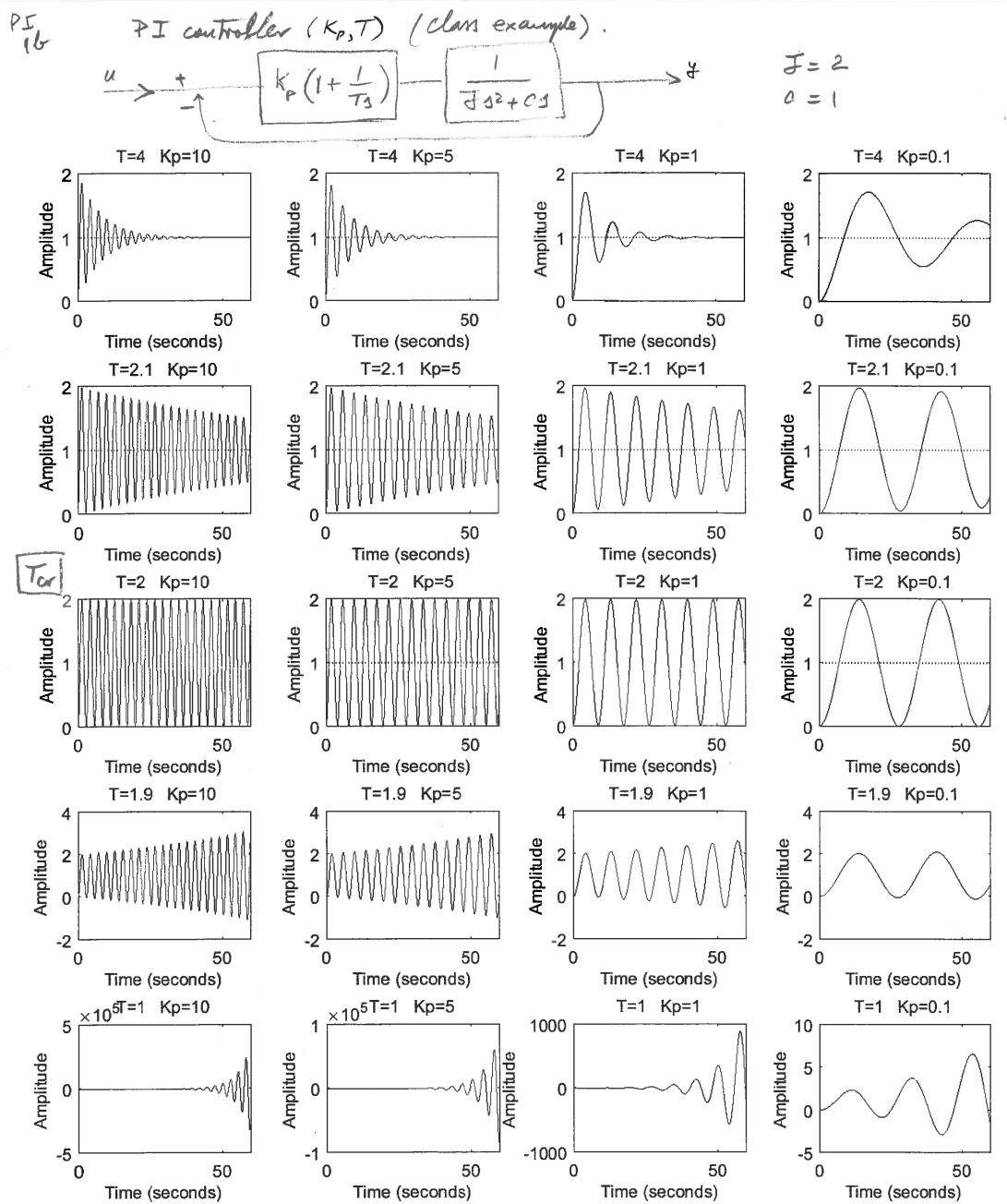
For stability,  $b_1 > 0$ , i.e.  $CT - \mathcal{J} > 0$ .

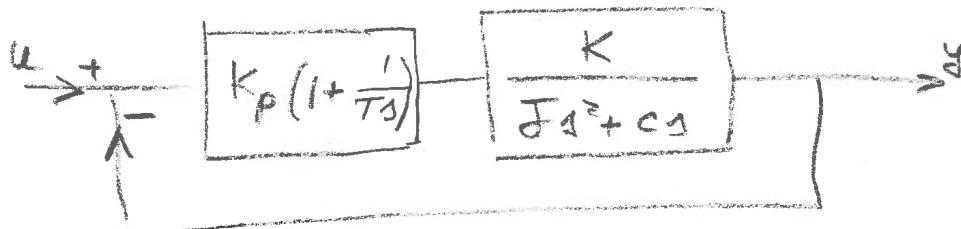
Need  $CT > \mathcal{J}$

$$T > \frac{\mathcal{J}}{C}$$

$$\text{denote } T_{cr} = \frac{\mathcal{J}}{C}$$

Need  $T > T_{cr}$  for stability.



<sup>2</sup>  
PSPI controller  $(K_p, T)$ 

$$G(s) = K_p \left(1 + \frac{1}{T_1}\right) \left(\frac{K}{J s^2 + C_1}\right) = \frac{K_p K (T_1 + 1)}{T_1 (J s^2 + C_1)}$$

$$G_{CL} = \frac{G}{1+G} = \frac{K_p K (T_1 + 1)}{T J s^3 + T C s^2 + K_p K T_1 + K_p K}$$

- 3<sup>rd</sup> order system

- May become unstable

- R-H stability criterion requires  $T > T_{cr}$ ,

$$T_{cr} = \frac{J}{C} \rightarrow \text{for stability (next page)}$$

Example :  $K=114$ ,  $J=10$ ,  $C=4$ ,  $T_{cr} = \frac{10}{4} = 2.5$   
(aircraft)

$\frac{Z_0}{P_1}$ R-H Table

$$(T_J s^3 + T_C s^2 + K_p K T) s + (K_p K) = 0$$

$s^3$	$T_J$	$K_p K T$
$s^2$	$T_C$	$K_p K$
$s^1$	$b_1$	$b_1 = \frac{T_C K_p K T - T_J K_p K}{T_C}$
$s^0$	$K_p K$	$= \frac{T_C - J}{C} K_p K$

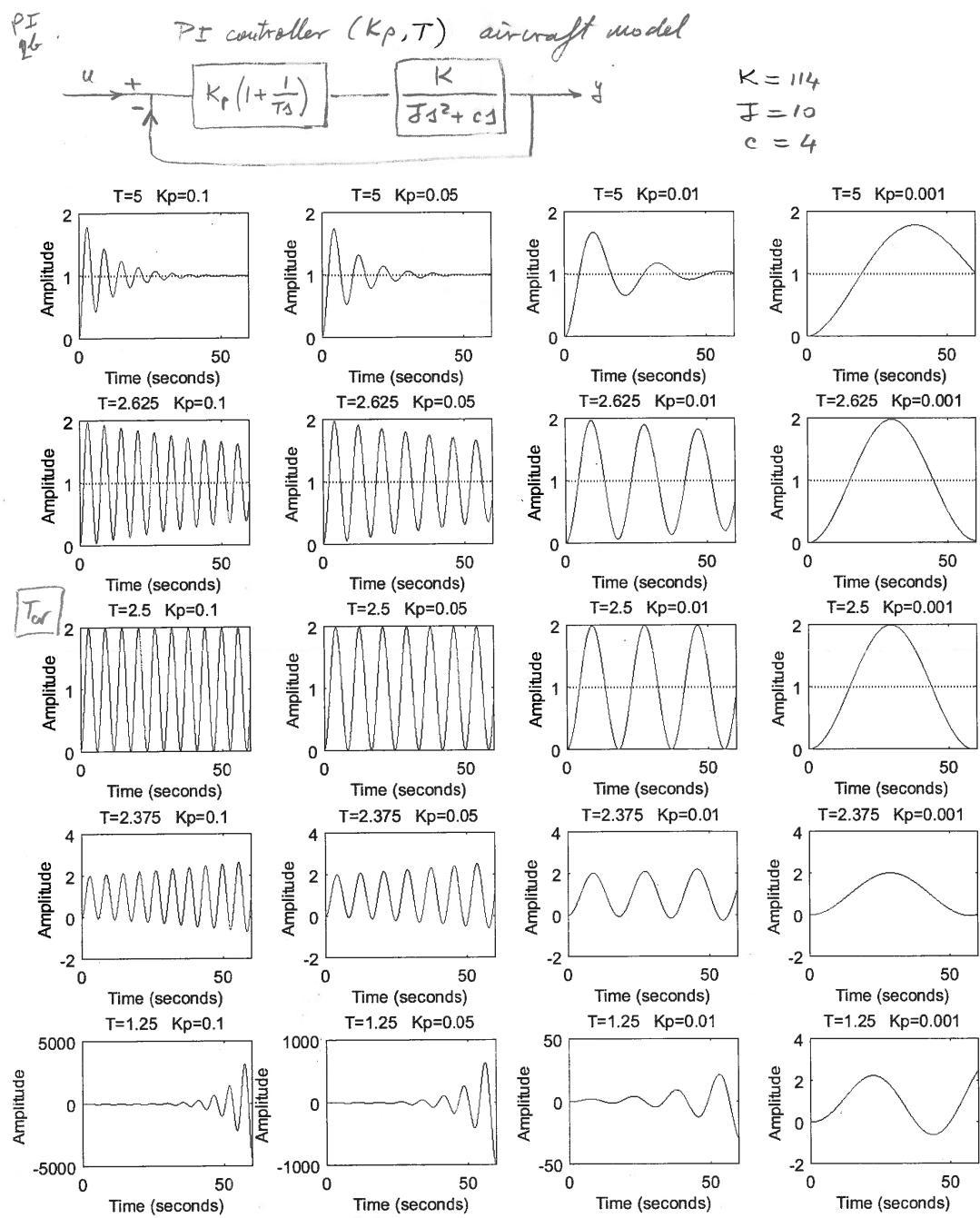
## Discussion

$s^3$	$T_J$	+ve
$s^2$	$T_C$	+ve
$s^1$	$b_1$	may be +ve or -ve
$s^0$	$K_p K$	+ve

INSTABILITY happens if  $b_1 < 0$ , i.e.,  $T_C < J$ .

For stability, need  $T_C > J$  or

$$T > T_{cr}, \quad T_{cr} = \frac{J}{C}$$



$\eta^C$   
P>

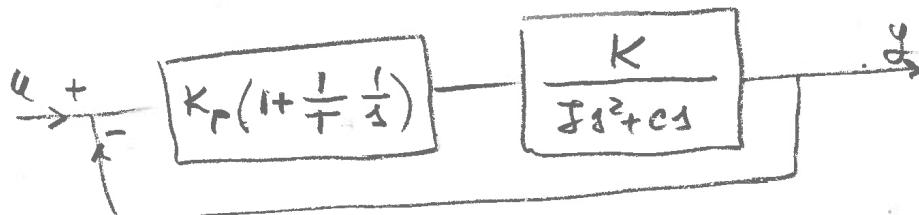
Aircraft roll response  
with  $(K_p, T)$  PI controller

$$K_p = 1/K = 1/114$$

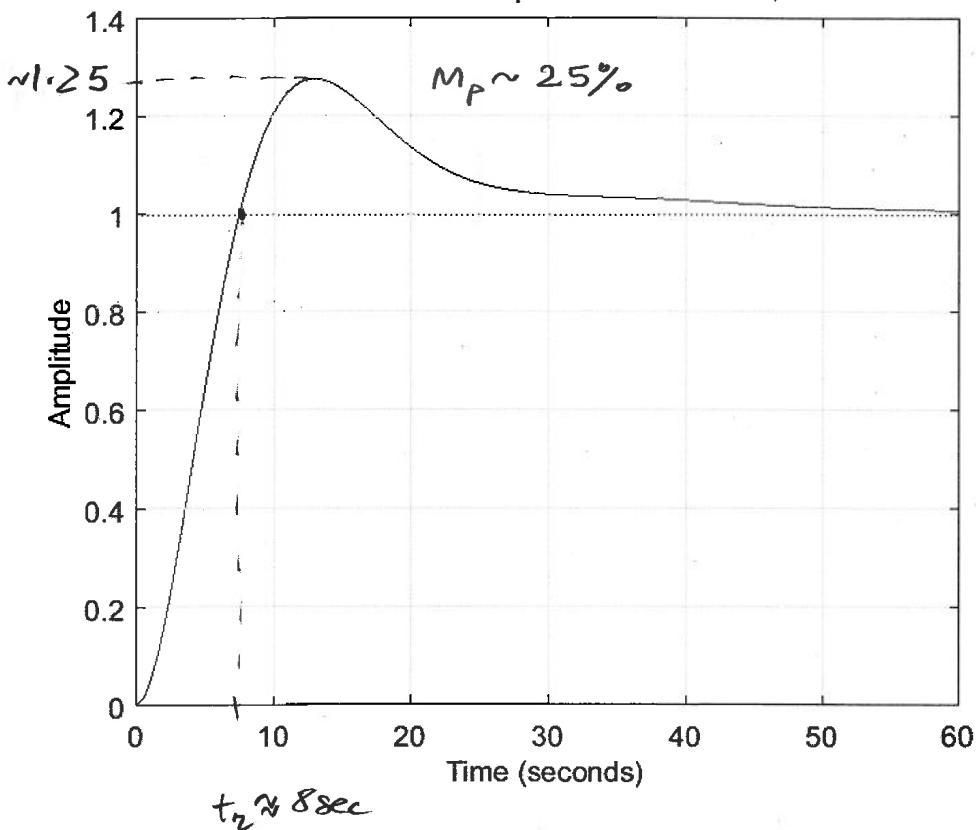
$$T = 20 > T_{cr}$$

$$\begin{aligned} K &= 114 \\ J &= 10 \\ C &= 4 \end{aligned}$$

$$T_{cr} = \frac{T}{C} = 2.5$$

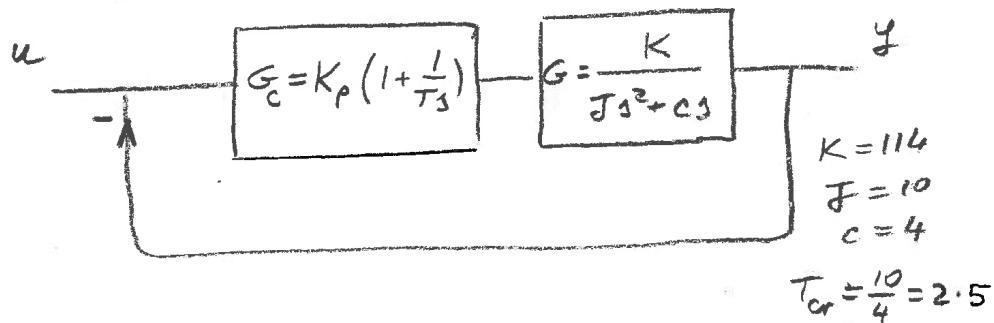


$T=20 \quad K_p=0.0087719$

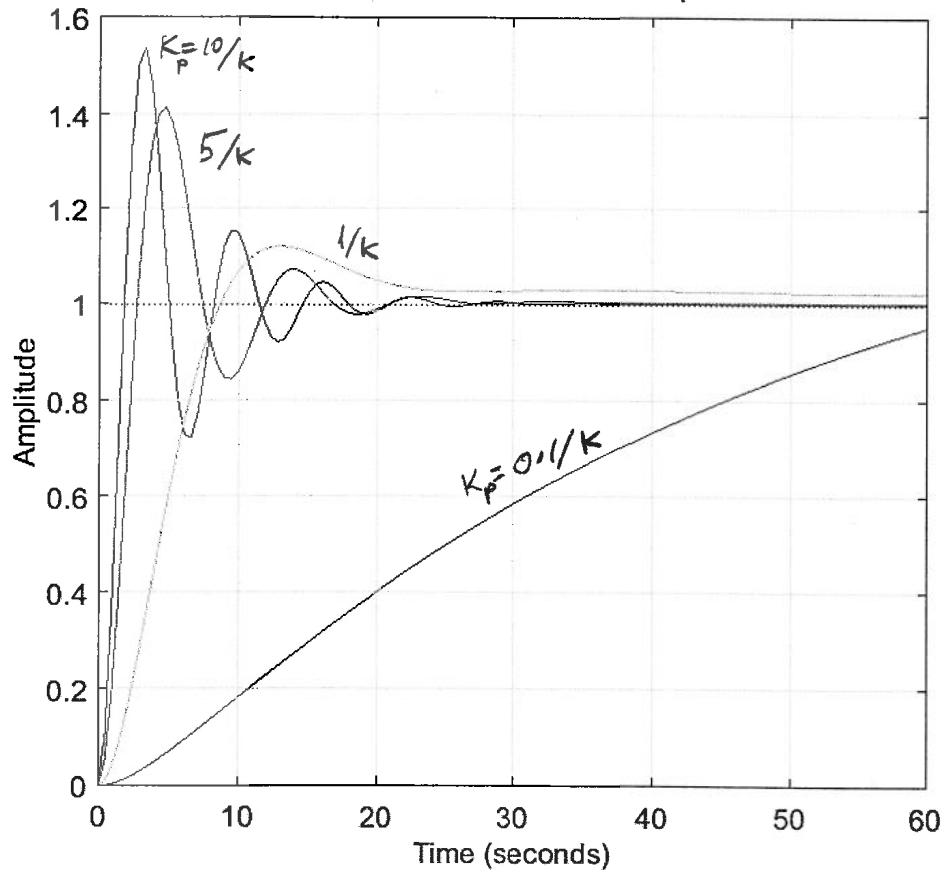


3

PI-control



$$T = 100 > T_{cr}$$

step response for various  $K_p$ 

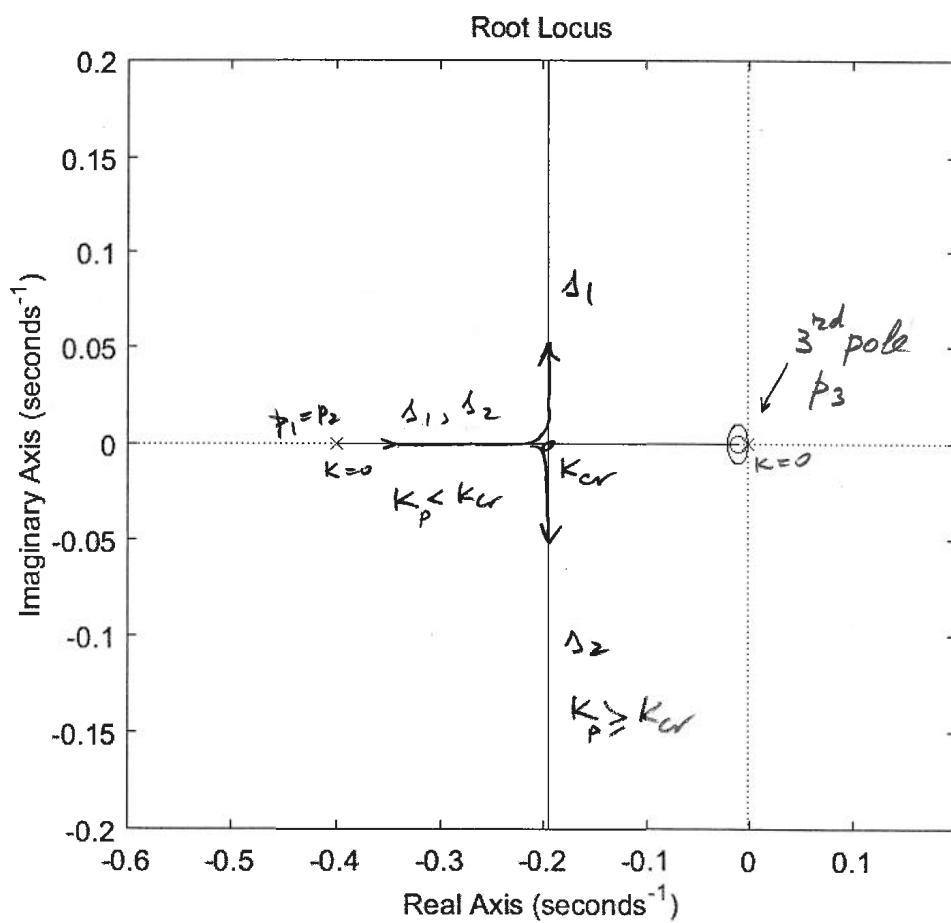
4

*P1 control*  
*Root locus*

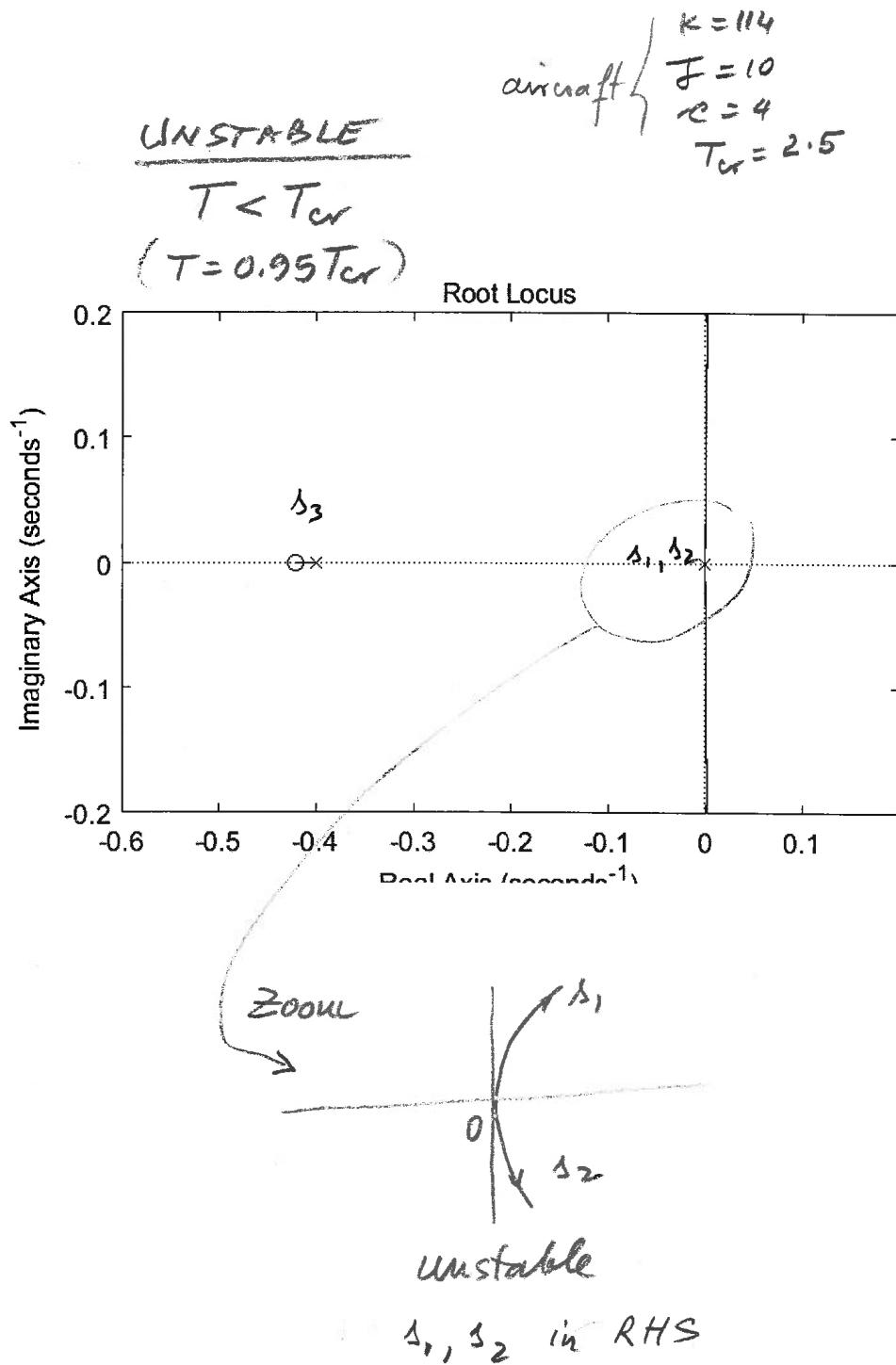
aircraft

$$\left. \begin{array}{l} K=114 \\ J=10 \\ c=4 \\ T_{cr}=2.5 \end{array} \right\}$$

$$T=100 > T_{cr}$$



5  
UNSTABLE PI Control  
Root locus



6

UNSTABLE PI control  
Step response

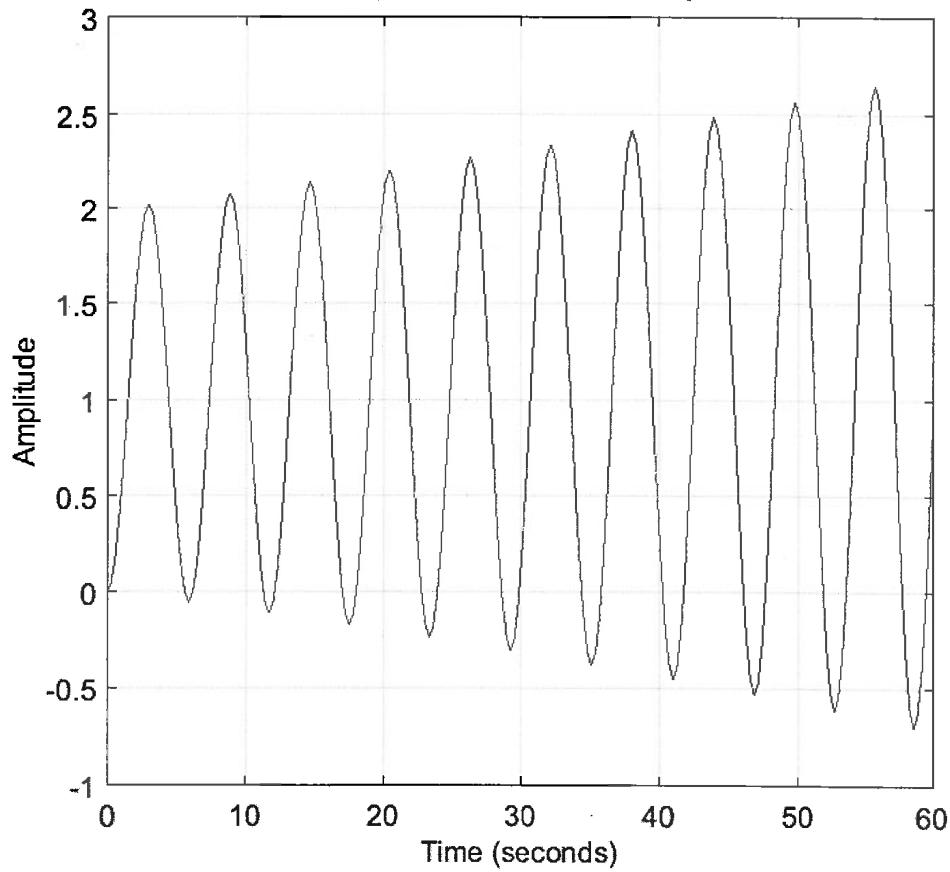
aircraft  
 $K = 114$   
 $J = 10$   
 $c = 4$   
 $T_{cr} = 2.5$

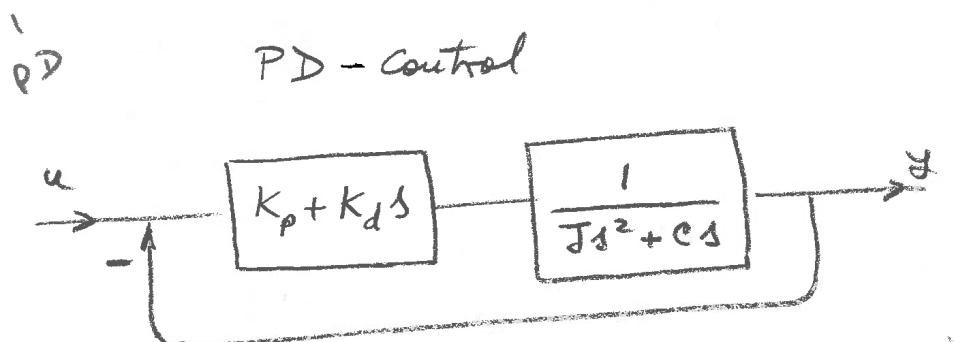
UNSTABLE

$T < T_{cr}$

( $T = 0.95T_{cr}$ )

step response for various  $K_p$





$$G(s) = \frac{1}{J s^2 + c s} \quad (\text{assume } K=1 \text{ for ease})$$

$$G_c(s) = K_p + K_d s$$

$$G_1(s) = \frac{K_p + K_d s}{J s^2 + c s}$$

$$G_{CL}(s) = \frac{K_p + K_d s}{J s^2 + c s + K_p + K_d s}$$

$$= \frac{K_p + K_d s}{J s^2 + (c + K_d) s + K_p}$$

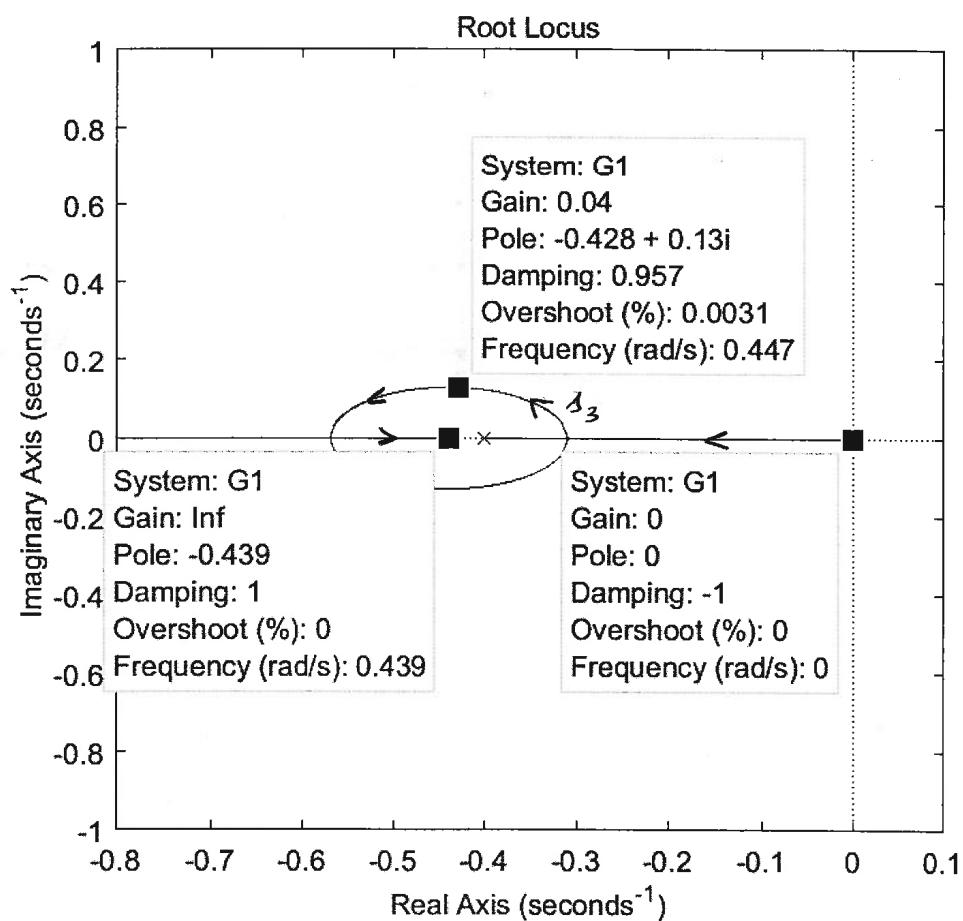
$\curvearrowright$  damping is augmented by  $K_d$

### strategy

- Adjust frequency with  $K_p$  to get small  $t_h$
- Reduce overshoot by increasing damping  
 $\curvearrowright$  with  $K_d$

<sup>2</sup>  
PDPD controller

Interesting behavior of the poles  $s_1, s_2$   
in the root locus



P D

PD control

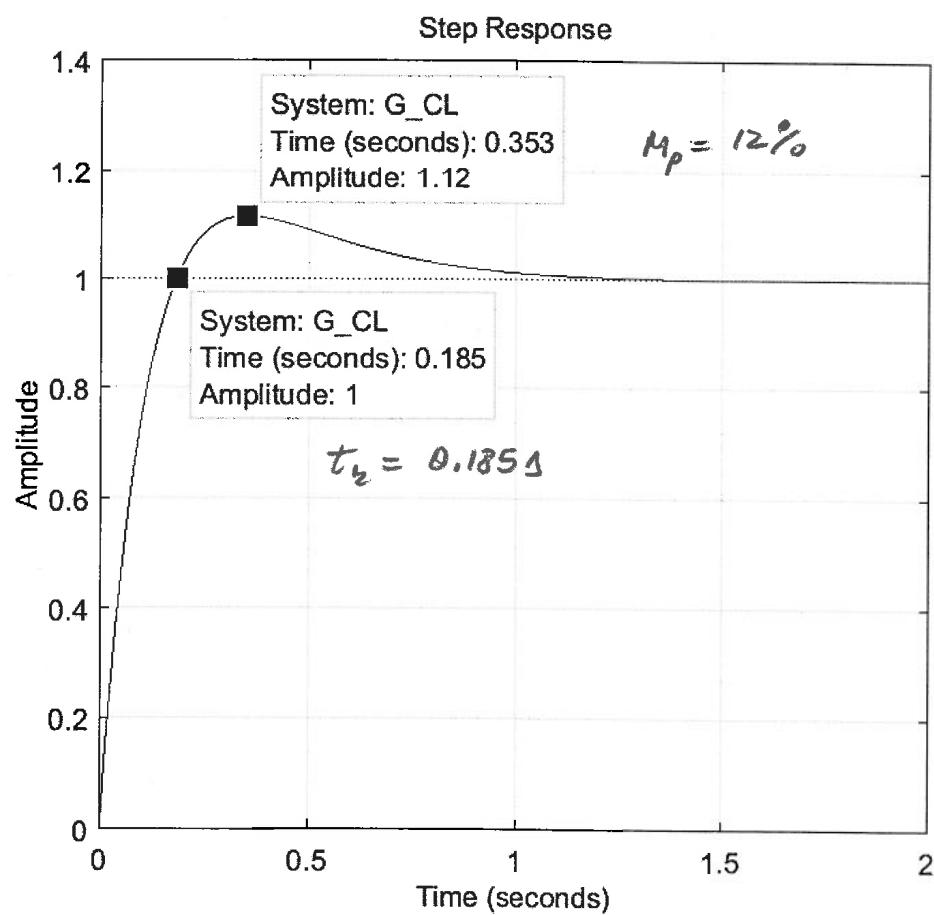
$$K = 114$$

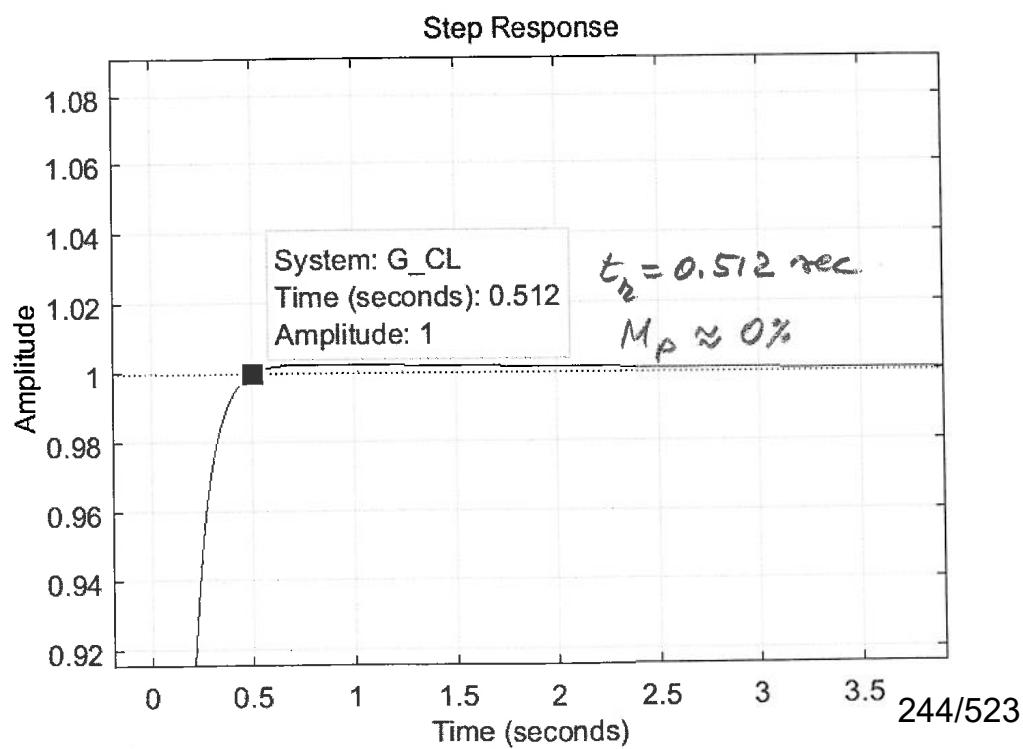
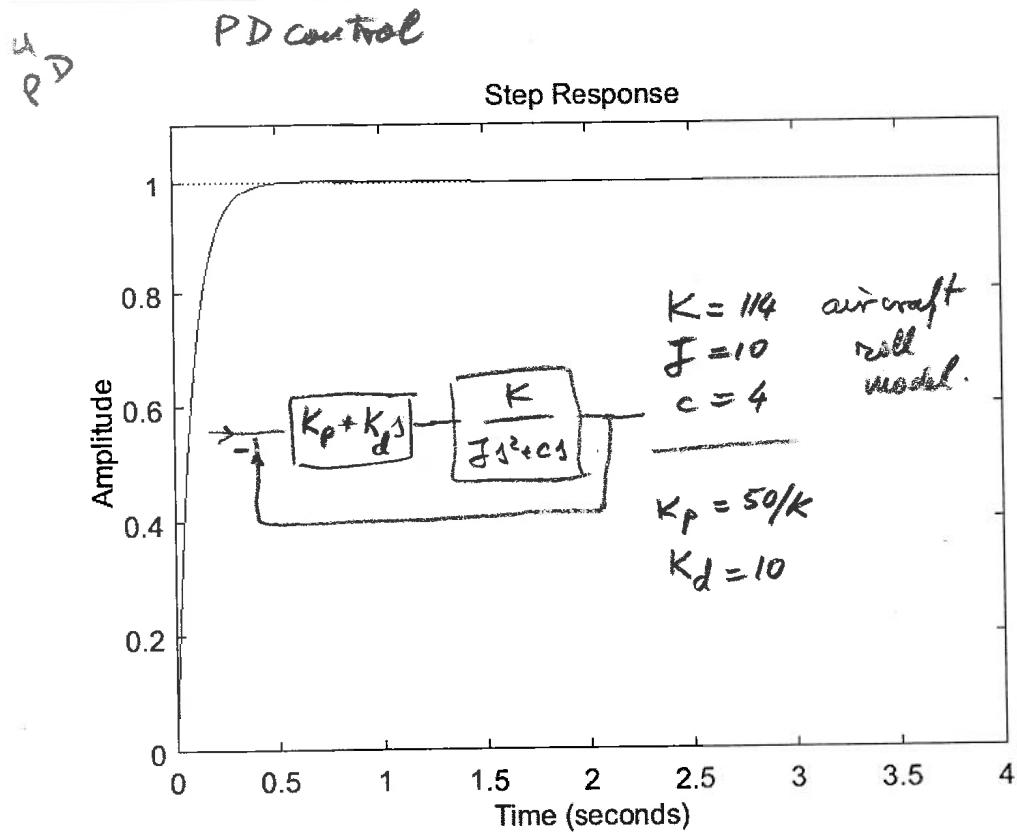
$$J = 10$$

$$C = 4$$

$$K_p = 3$$

$$K_d = 1$$





## 7.7 SIMULINK Aircraft Roll Motion

## SIMULINK

### Aircraft Roll Motion Autopilot Development

Table of Contents:

<b>1</b>	<b>CONTROL SYSTEM DESIGN OBJECTIVES.....</b>	<b>2</b>
<b>2</b>	<b>UNCONSTRAINED AIRCRAFT ROLL MOTION .....</b>	<b>3</b>
2.1	MODEL.....	3
2.2	SIMULATION PARAMETERS .....	4
2.3	DISPLAY SETUP.....	5
<b>3</b>	<b>UNIT FEEDBACK CONTROL OF AIRCRAFT ROLL MOTION.....</b>	<b>7</b>
3.1	SIMULATION PARAMETERS .....	7
3.2	DISPLAY SETUP.....	8
3.3	AIRCRAFT ROLL RESPONSE WITH UNIT FB .....	9
<b>4</b>	<b>VARIABLE GAIN FB CONTROL OF AIRCRAFT ROLL MOTION (P CONTROL) .....</b>	<b>10</b>
4.1	P-CONTROLLED AIRCRAFT ROLL MODEL.....	10
4.2	DISPLAY SETUP.....	10
4.3	P-CONTROL GAIN CHANGES .....	11
4.3.1	$K_p=1$ Response.....	12
4.3.2	$K_p=10$ Response.....	13
4.3.3	$K_p=0.1$ Response.....	14
4.3.4	$K_p=0.01$ Response.....	15
<b>5</b>	<b>PI CONTROL .....</b>	<b>17</b>
5.1	PI CONTROL SETUP.....	17
5.2	AIRCRAFT ROLL RESPONSE WITH PI CONTROL .....	18
<b>6</b>	<b>PD CONTROL .....</b>	<b>19</b>
6.1	PD CONTROL SETUP .....	19
6.2	AIRCRAFT ROLL RESPONSE WITH PD CONTROL .....	19
<b>7</b>	<b>PID CONTROL.....</b>	<b>22</b>
7.1	PID CONTROL SETUP .....	22
7.2	AUTOMATIC TUNING OF THE PID CONTROLLER .....	23
7.3	MANUAL ADJUSTMENT OF THE PID TUNER .....	25
7.4	VERIFICATION OF THE PID TUNING PROCESS .....	26
<b>8</b>	<b>VELOCITY FEEDBACK CONTROL .....</b>	<b>30</b>

## 1 CONTROL SYSTEM DESIGN OBJECTIVES

Consider the aircraft roll transfer function  $G(s) = \frac{K}{Js^2 + cs} = \frac{114}{10s^2 + 4s}$

Design a feedback control system to control the aircraft roll motion with the following control design objectives:

Design Objective 1: Control the unconstraint aircraft motion resulting from an aileron input.  
Have a autopilot system that can maintain the aircraft at a constant bank angle

We wish to achieve this objective through feedback (FB)

Design Objective 2: Achieve a reasonable aircraft roll response.

We define ‘reasonable response’ using two control design specifications:

- DS1: Fast response time as measured by rise time  
 $t_r \leq 1.5$  sec
- DS2: maximum percentage overshoot for step input less than 20%  
 $M_p \leq 20\%$

## 2 UNCONSTRAINED AIRCRAFT ROLL MOTION

### 2.1 MODEL

Open a new model canvas and save it as 'SIMULINK\_airplane\_roll\_unitFB'

From Simulink Library Browser drag onto the new canvas the following blocks:

- select 'Step' from 'Sources'
- select 'Transfer Fcn' from 'Continuous'
- select 'Scope' from 'Sinks', enter 'Number of inputs' = 2

Create annotation text boxes above each box as follows:

- 'reference signal' above the 'Step' box
- 'aircraft roll dynamics' above 'Transfer Fcn' box
- 'display' above 'Scope'

Change properties of 'Transfer Fcn' block to represent the aircraft roll transfer function

$$G(s) = \frac{K}{Js^2 + cs} = \frac{114}{10s^2 + 4s}$$

Connect the blocks:

- 'reference signal' → 'aircraft dynamics' → 'display' (1<sup>st</sup> port)
- 'reference signal' → 'display' (2<sup>nd</sup> port)

Use the pull down menu: 'Edit → Copy Current View to Clipboard → Metafile' to capture only the model as shown in Figure 1.

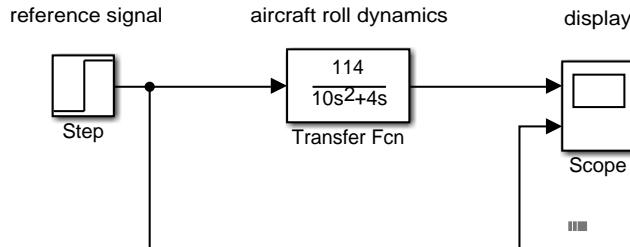


Figure 1

## 2.2 SIMULATION PARAMETERS

Use the pull down menu to open Configuration Parameters window, i.e.,  
Simulation → Model Configuration Parameters

Make:

- Max step size: 1e-4
- Stop time: 2

The rest should remain unchanged (verify that they are the same as inFigure 2)

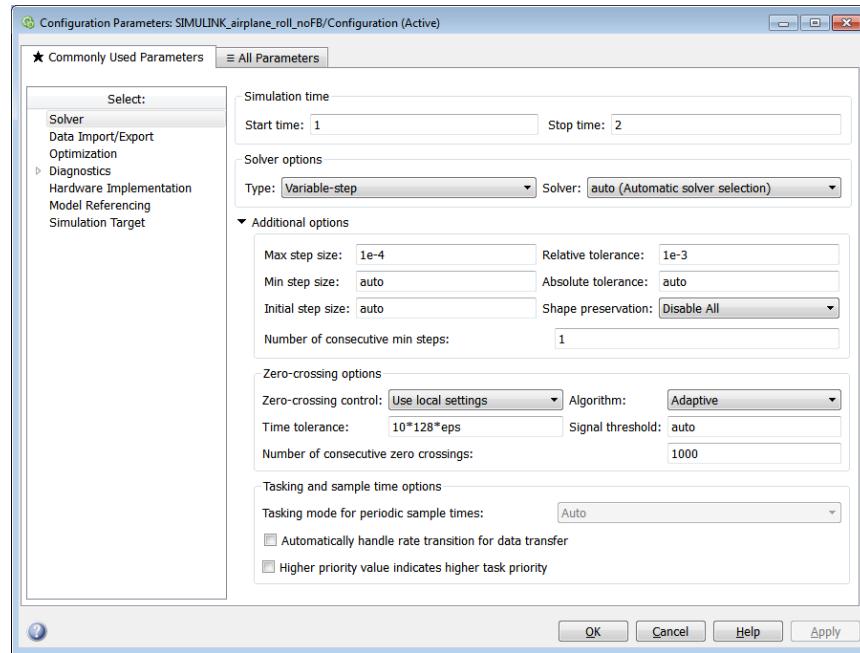


Figure 2

### 2.3 DISPLAY SETUP

Double click on the ‘Scope’ block in the SIMULINK model. The ‘Scope’ display should open in a new window.

Select ‘View →Style’ to open ‘Style Scope’ dialog box (Figure 3a).

Choose color white for:

- Figure color: ‘bucket’
- Axes colors: ‘bucket’

Choose color black for Axes colors: ‘brush’

Choose color blue for ‘Properties for line: Input step’ (Figure 3a)

Choose color red for ‘Properties for line: Transfer Fcn’ (Figure 3b)

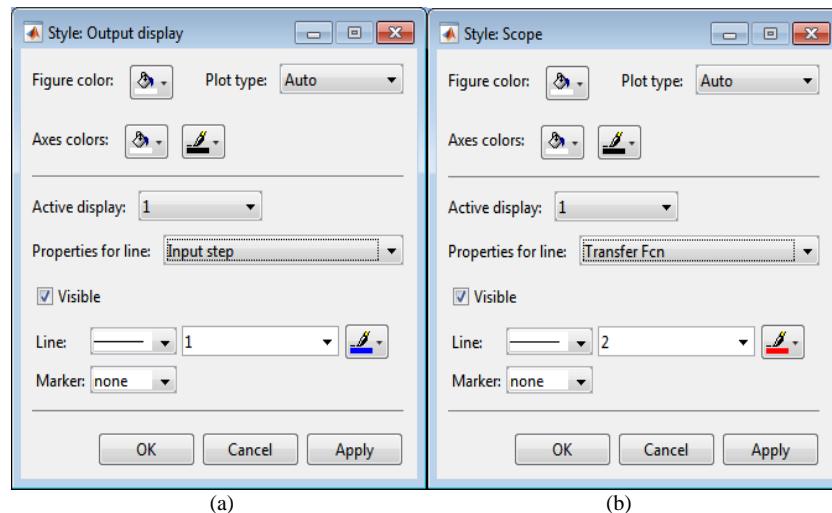


Figure 3

The ‘Scope’ display should look as shown in Figure 4:

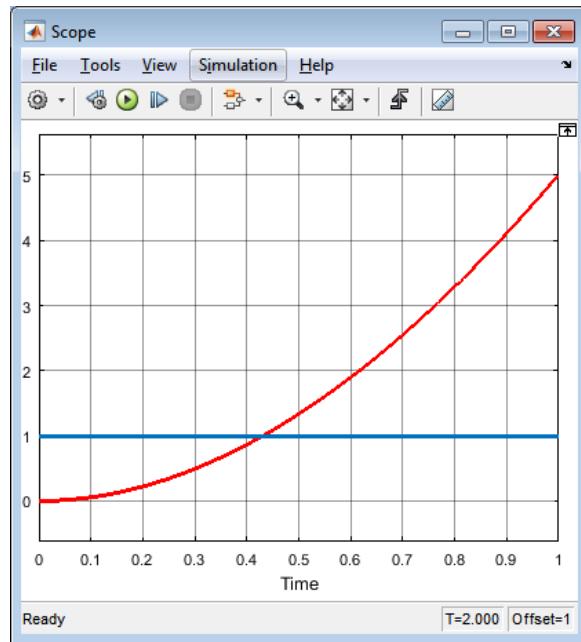


Figure 4

Note that the blue line represents the step input whereas the red line represent the aircraft roll response, which growth continuously. This situation is unacceptable. To counteract it, use feedback control as shown in next section.

### 3 UNIT FEEDBACK CONTROL OF AIRCRAFT ROLL MOTION

Open a new model canvas and save it as ‘SIMULINK\_airplane\_roll\_unitFB’  
From Simulink Library Browser drag onto the new canvas the following blocks:

- ‘Step’ from ‘Sources’
- ‘Sum’ from ‘Commonly Used Blocks’
- ‘Scope’ from ‘Sinks’
- ‘Transfer Fcn’ from ‘Continuous’

Create annotation text boxes above each box as follows:

- ‘reference signal’ above the ‘Step’ box
- ‘aircraft roll dynamics’ above ‘Transfer Fcn’ box
- ‘output signal’ above ‘Scope’

Change properties of ‘Transfer Fcn’ block to represent the aircraft roll transfer function

$$G(s) = \frac{114}{10s^2 + 4s}$$

Change the second port of the ‘Sum’ block to negative (-). Do this by using ‘right-click’ → ‘Block Parameters: Sum’ and then putting ‘+–‘ in ‘List of signs’

Connect the blocks:

- ‘reference signal’ → ‘Sum’ → ‘aircraft dynamics’ → ‘display’
- output from ‘aircraft dynamics’ to negative (-) port of the ‘Sum block’ (this is the unit feedback closed loop)

Use the pull down menu: ‘Edit → Copy Current View to Clipboard → Metafile’ to capture only the model (Figure 5).

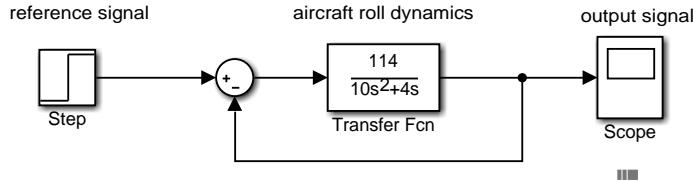


Figure 5

#### 3.1 SIMULATION PARAMETERS

Use the pull down menu to open Configuration Parameters window, i.e.,  
Simulation → Model Configuration Parameters

Make:

- Max step size: 1e-4
- Stop time: 40

The rest should remain unchanged (verify that they are the same as in the figure)

### 3.2 DISPLAY SETUP

Double click on the 'Scope' block in the SIMULINK model. The 'Scope' display should open in a new window.

Select 'View →Style' to open 'Style Scope' dialog box.

Choose color white for:

- Figure color
- Axes colors: bucket

Choose color black for Axes colors: brush

'Properties for line: Transfer Fcn': 1.5 weight, red (Figure 6).

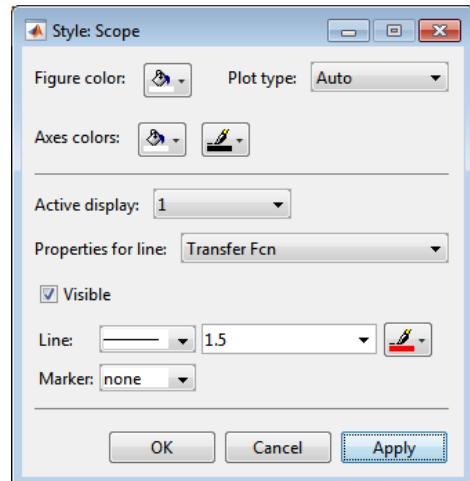


Figure 6

### 3.3 AIRCRAFT ROLL RESPONSE WITH UNIT FB

Press the 'Run' button. The aircraft roll response with unit FB is displayed (Figure 7).

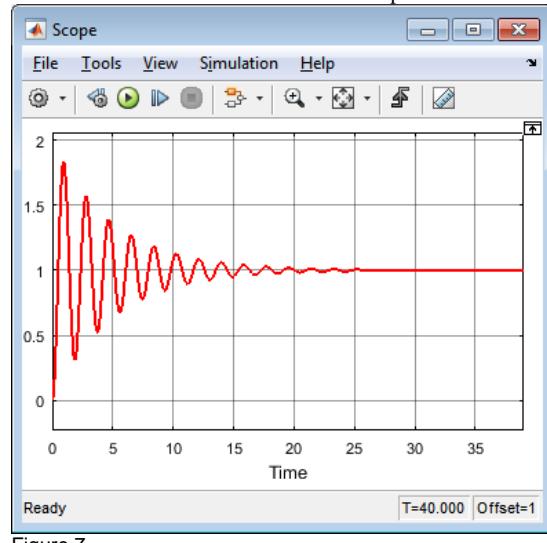


Figure 7

It is apparent that the response is unsatisfactory because:

- many oscillations until it settles down to  $x_{ss} = 1$
- large overshoot ( $x_p \approx 1.8$ ,  $M_p \approx 80\%$ )

#### 4 VARIABLE GAIN FB CONTROL OF AIRCRAFT ROLL MOTION (P CONTROL)

To improve the aircraft FB response, we can try to use variable gain control. This is known as proportional control or ‘P control’.

##### 4.1 P-CONTROLLED AIRCRAFT ROLL MODEL

Add a gain box to the model to obtain the P-controller (Figure 8).

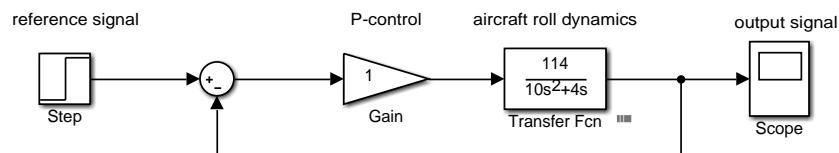


Figure 8

Save this model as ‘SIMULINK\_aircraft\_roll\_P\_control’

##### 4.2 DISPLAY SETUP

Have the display setup as before.

In addition, stop automatic axes scaling by doing the following:

- Tools → Axes Scaling → Axis Scaling Properties

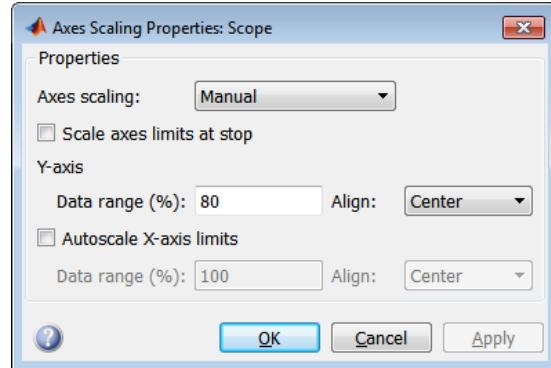


Figure 9

#### 4.3 P-CONTROL GAIN CHANGES

The P-control gain  $K_p$  can be modified as follows:

Right click 'Gain' box, select 'Block Parameters'. The 'Block Parameters: Gain' dialog box open (Figure 10).

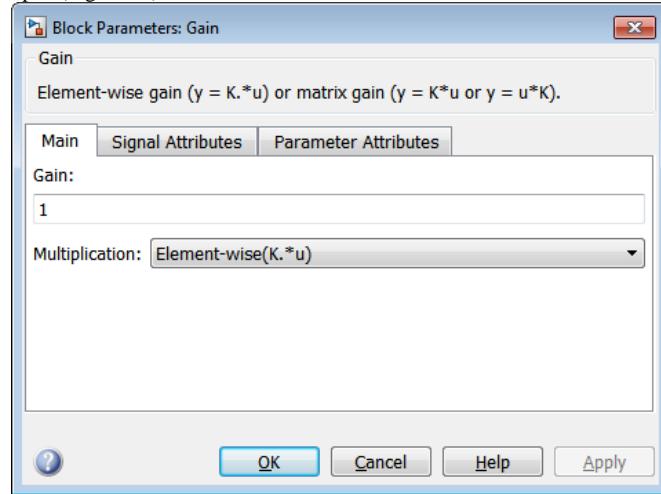


Figure 10

The gain default value is 'Gain: 1', i.e.,  $K_p=1$ . After entering a new gain value, press 'Apply'. Keep the dialog box open for the next gain value. (simulation runs OK with the dialog box open.)

Try a number of different gains to see their effect on the response. Gains to be tried are:

- $K_p=1$
- $K_p=10$
- $K_p=0.1$
- $K_p=0.01$

#### 4.3.1 Kp=1 Response

Put 'Gain = 1' in 'Block Parameters: Gain' dialog box and press 'Apply'. The response is shown in Figure 11.

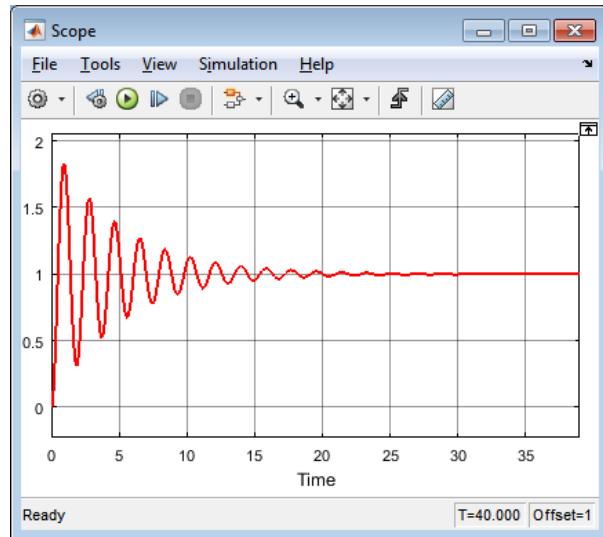


Figure 11

This  $K_p=1$  case corresponds to Unit FB which we have already studied. It is unsatisfactory because

- many oscillations until it settles down to  $x_{ss} = 1$
- large overshoot ( $x_p \approx 1.8$ ,  $M_p \approx 80\%$ )

#### 4.3.2 K<sub>p</sub>=10 Response

Put ‘Gain = 10’ in ‘Block Parameters: Gain’ dialog box and press ‘Apply’. The response is shown in Figure 12.

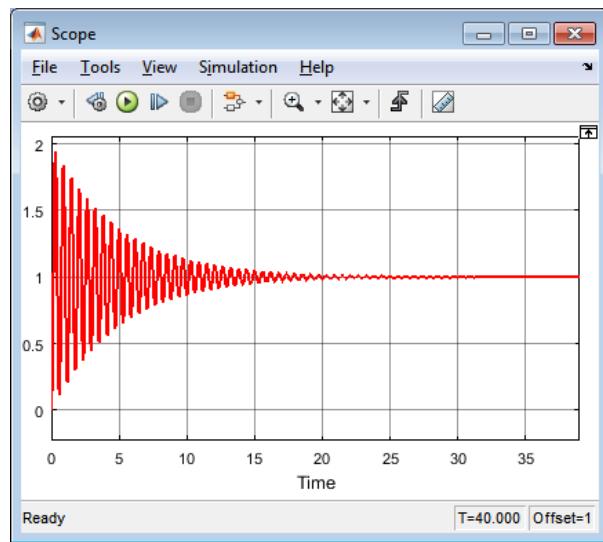


Figure 12

This K<sub>p</sub>=10 response is unsatisfactory because

- very many oscillations until it settles down to  $x_{ss} = 1$
- large overshoot ( $x_p \approx 2$ ,  $M_p \approx 100\%$ )

#### 4.3.3 Kp=0.1 Response

Put ‘Gain = 0.1’ in ‘Block Parameters: Gain’ dialog box and press ‘Apply’. The response is shown in Figure 13.

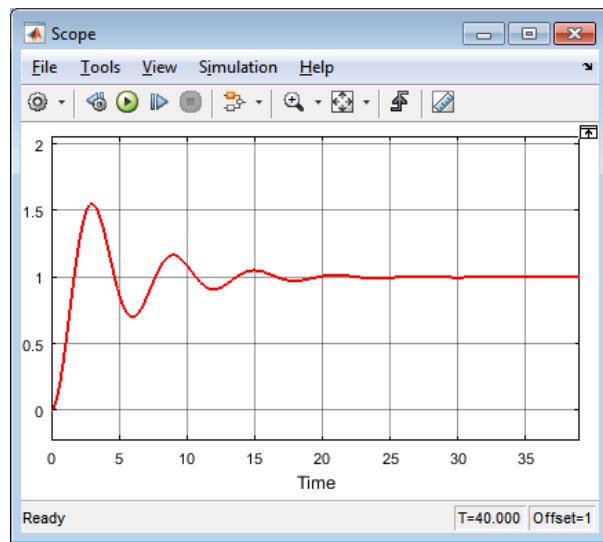


Figure 13

The response for  $K_p=0.1$  seems better because:

- fewer oscillations until it settles down to  $x_{ss} = 1$
- smaller overshoot ( $x_p \approx 1.55$ ,  $M_p \approx 55\%$ )

#### 4.3.4 Kp=0.01 Response

Put ‘Gain = 0.01’ in ‘Block Parameters: Gain’ dialog box and press ‘Apply’. The response is shown in Figure 14.

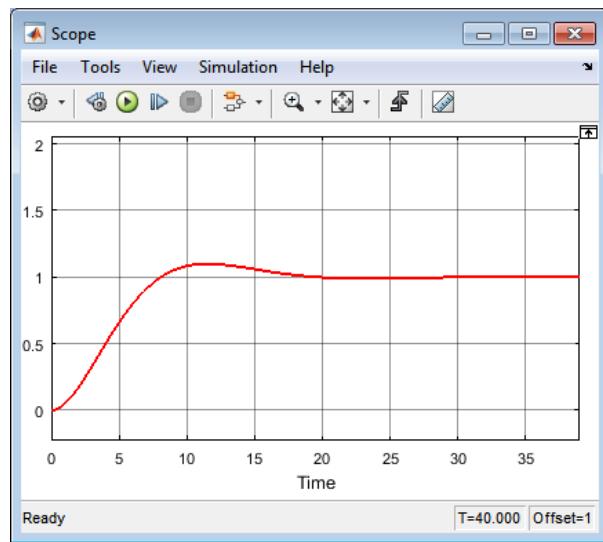


Figure 14

The response for  $K_p=0.01$  seems even better because:

- almost no oscillation until it settles down to  $x_{ss} = 1$
- much smaller overshoot ( $x_p \approx 1.1$ ,  $M_p \approx 10\%$ )

However, the rise time is much longer than before ( $t_r \approx 8$  sec). This is unacceptable because it makes the aircraft very sluggish.

We need to examine the mathematics of P-control. Recall:

$$\omega_n = \sqrt{\frac{K}{J}}, \zeta = \frac{c}{2\sqrt{JK}}$$

where K is the overall forward gain ( $K=K_p * K_1$ , with  $K_1=114$  for our aircraft roll model). Hence we can calculate:

- rise time:  $t_r = \frac{\pi - \varphi}{\omega_n \sqrt{1 - \zeta^2}}$  where  $\varphi = \sin^{-1} \sqrt{1 - \zeta^2}$
- overshoot:  $M_p = e^{-\pi \frac{\zeta}{\sqrt{1 - \zeta^2}}}$

A plot of these values is given in Figure 15.

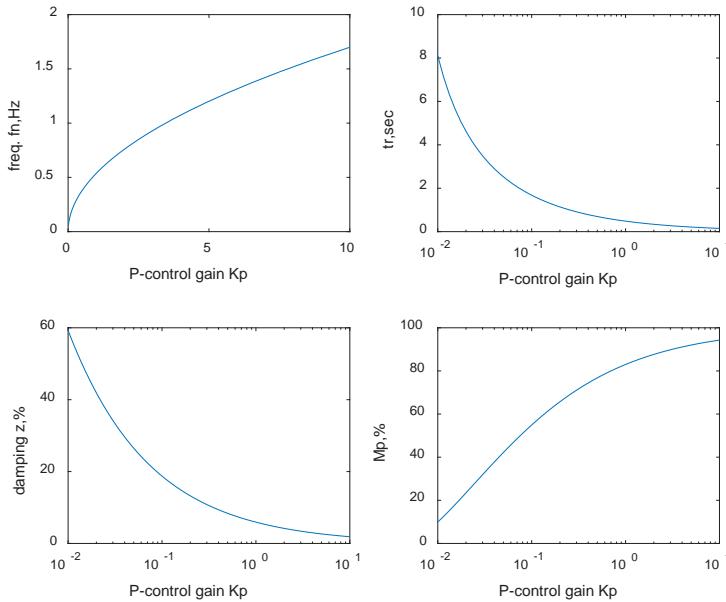


Figure 15

The above plot shows that small overshoot  $M_p$  happens at low gain values. However, at these low gain values, the rise time  $t_r$  becomes large. Hence, we conclude that **P-control gain has an opposite effect on the two performance indicators,  $M_p$  and  $t_r$  considered in this controller design**. We need to explore other control options.

## 5 PI CONTROL

PI control stands for ‘proportional + integrative control’.

### 5.1 PI CONTROL SETUP

The aircraft model with PI control has two branches, one for the P block (which is just a simple gain) and the other for the I block which consists of a gain followed by an integrator. The PI controlled aircraft model looks as shown in Figure 16.

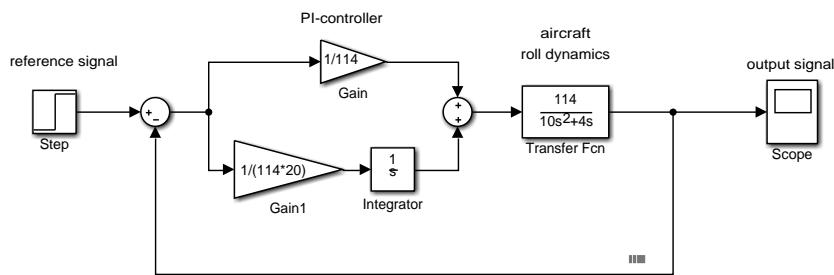


Figure 16

Construct this model and save it as ‘SIMULINK\_aircraft\_roll\_PI\_control\_Kp\_Ki’

Observe that the model has two gains:

- Proportional gain,  $K_p=1/114$
- Integrative gain  $K_i=1/(114*20)$

The resulting PI controller has the expression:

$$G_{PI}(s) = K_p + K_i \frac{1}{s}$$

## 5.2 AIRCRAFT ROLL RESPONSE WITH PI CONTROL

Run the PI control model. The resulting response will show up in the ‘Scope’ window (Figure 17):

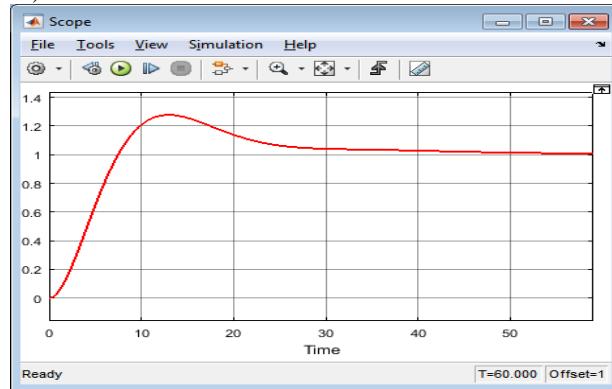


Figure 17

Press the ‘Cursor Measurements’ button and set the cursors to measure rise time and overshoot (Figure 18).

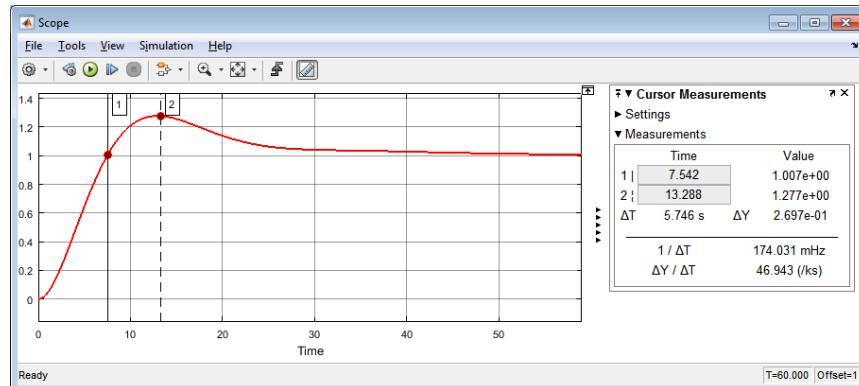


Figure 18

The results are:

- rise time  $t_r = 7.542$  sec
- overshoot  $M_p = 27.7\%$

These values are better than the values obtained with P-control. However, they are still below the target values. Need to try something else.

## 6 PD CONTROL

PD control stands for ‘proportional + derivative control’.

### 6.1 PD CONTROL SETUP

The aircraft model with PI control has two branches, one for the proportional P-block (which is just a simple gain) and the other for the derivative D-block. The PD controlled aircraft model looks as shown in Figure 19.

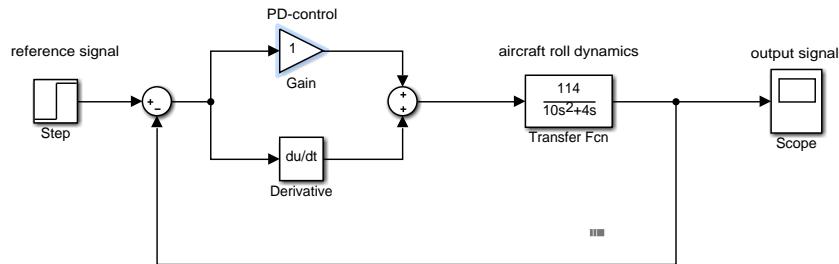


Figure 19

Construct this model and save it as ‘SIMULINK\_aircraft\_roll\_PD\_control’

### 6.2 AIRCRAFT ROLL RESPONSE WITH PD CONTROL

Set the model run time to 20 sec and run it. The response is as shown in Figure 20.

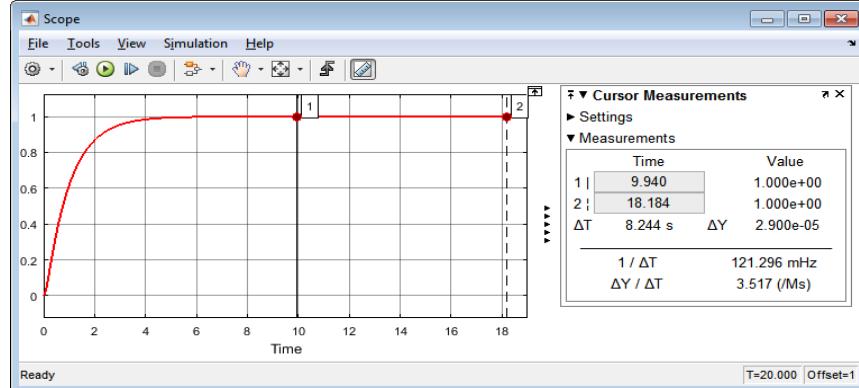


Figure 20

The response converges continuously towards the target value 1, but the rise time is around 10 sec.

This is unacceptable. But we can now try different value of the P gain  $K_p$ .

Increase P control gain to  $K_p=10$ . The model looks as shown in Figure 21.

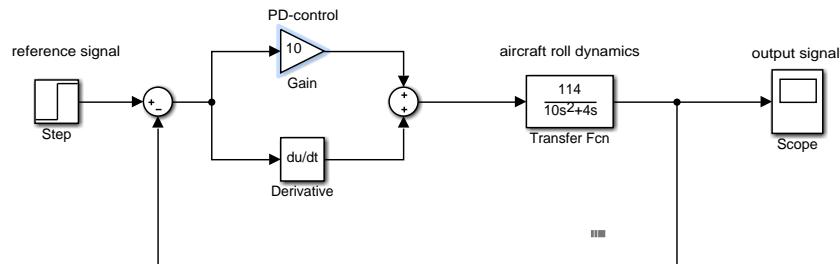


Figure 21

The response of this model with  $K_p=10$  is as shown in Figure 22.

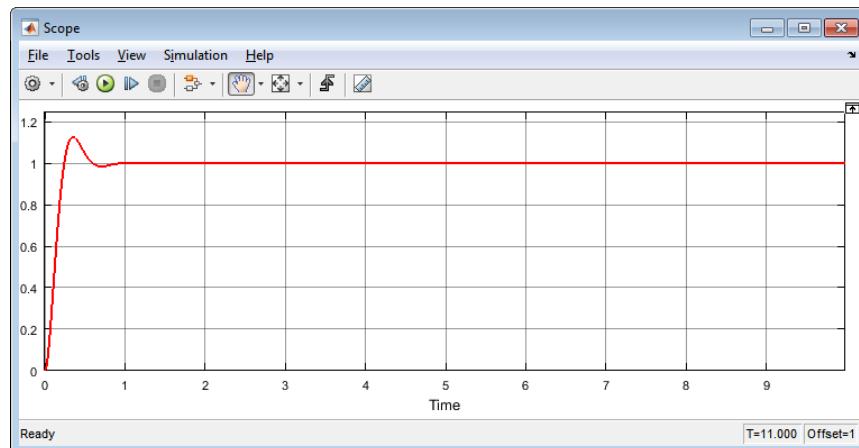


Figure 22

Note that the response has become much crispier but overshoot is now present.

Press the ‘Cursor Measurements’ button to get measurements (Figure 23).

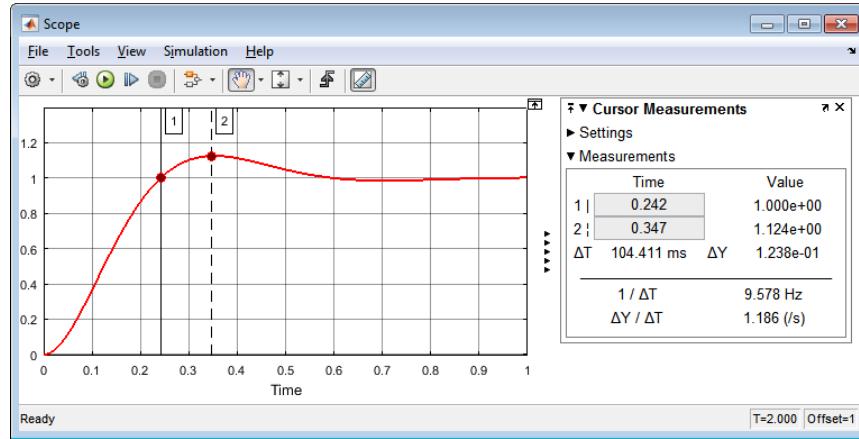


Figure 23

The cursor measurements allow us to determine  $t_r = 0.242$  sec and  $M_p = 12.4\%$ .

#### Conclusion:

PD control with  $K_p=10$  seems to give a reasonable response:

- short rise time  $t_r = 0.242$  sec
- small overshoot  $M_p = 12.4\%$

Note of caution: I cannot reproduce these results exactly in MATLAB control systems toolbox. It is perhaps because PD control is not a stable control. In practice, the D part of a PD controller is difficult to implement because differentiation is not easy to obtain with electrical system; therefore, the D part of the controller is implemented as a compensator  $s / (T_1 s + 1)$ .

## 7 PID CONTROL

PID control stands for ‘proportional + integrative + derivative control’.

### 7.1 PID CONTROL SETUP

The aircraft model with PID control has two branches, uses the PID block which can be found in ‘Continuous’ part of the library. The PI controlled aircraft model looks as shown in Figure 24.

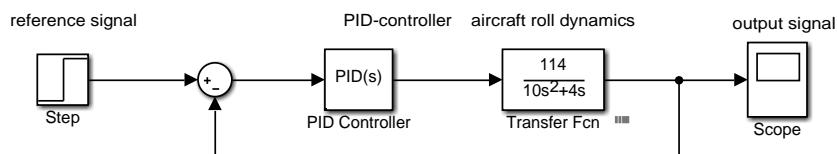


Figure 24

Construct this model and save it as ‘SIMULINK\_aircraft\_roll\_PID\_control’.

The PID block represents the following controller transfer function:

$$G_c(s) = K_p + \frac{K_I}{s} + K_D s$$

The default values at startup are  $K_p = 1$ ,  $K_I = 1$ ,  $K_D = 0$ . The response with these default values is as shown in Figure 25.

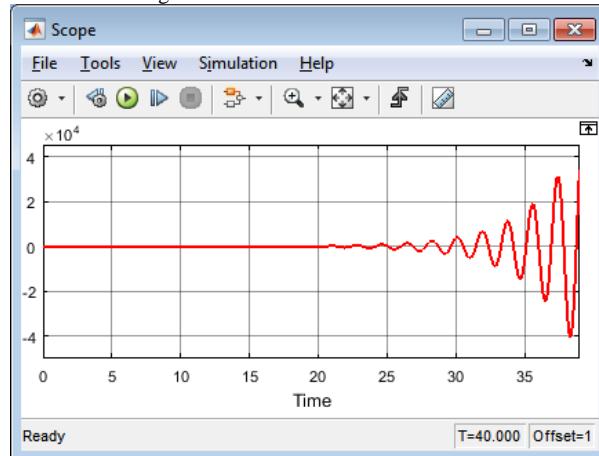


Figure 25

This is an unstable response and needs to be corrected! We will adjust the PID controller parameters to obtain a stable response which meets the design specifications. This process of adjustment is called ‘tuning’.

## 7.2 AUTOMATIC TUNING OF THE PID CONTROLLER

Double click the PID box and open it (Figure 26).

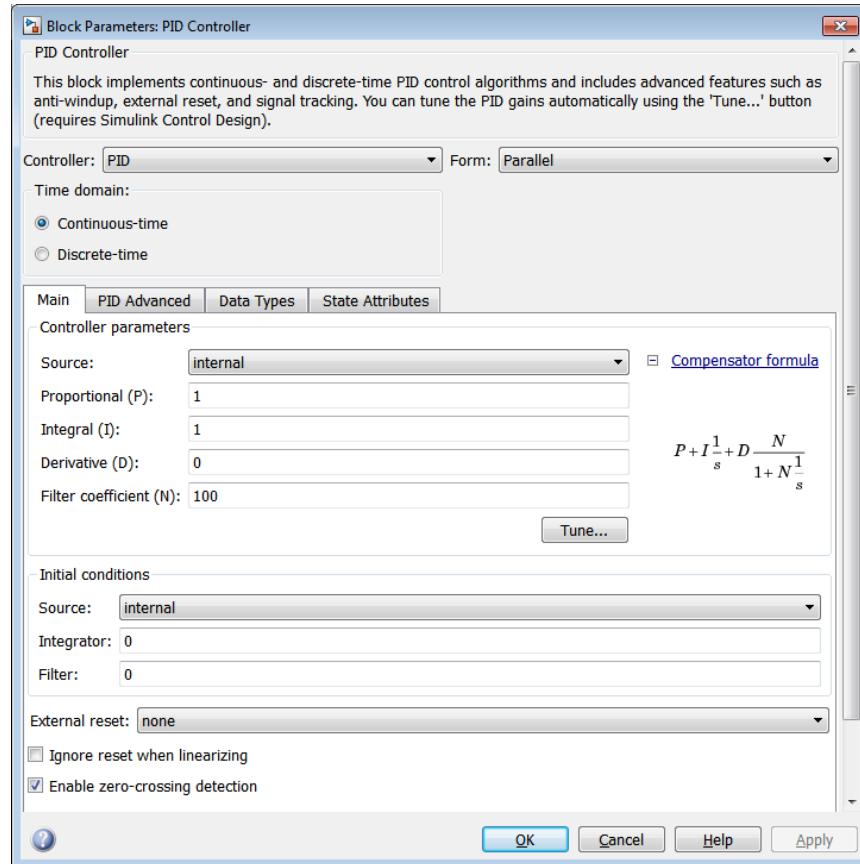


Figure 26

Next, press the ‘Tune...’ button. This will open a separate window for tuning, i.e., the ‘PID Tuner’ window shown in Figure 27.

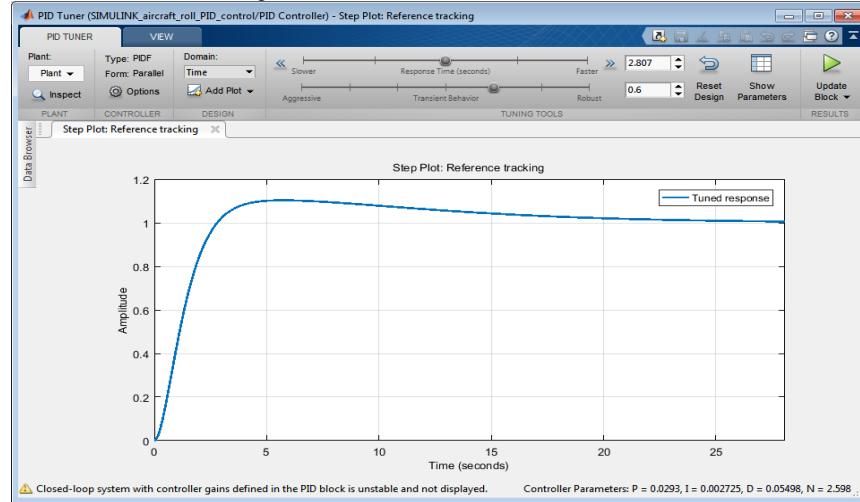


Figure 27

Note the message at the bottom of the window in the left corner: ‘Closed-loop system with controller gains defined in the PID block is unstable and not displayed’.

In the right bottom corner, we read: ‘Controller Parameters: P=0.0293, I=0.002725, D=0.05498, N=2.598’. These controller parameters have been obtained by the tuner through an internal algorithm.

Press the ‘Show parameters’ button in the upper right corner to open Figure 28.

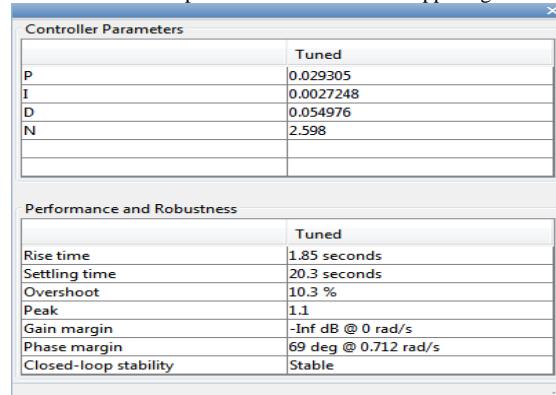


Figure 28

In this window, we read Rise time = 1.85 sec and Overshoot = 10.3%. Note that MATLAB defines the rise time in the 10%--90% range To evaluate the 0--100% rise time  $t_r$ , read the upper-right side window next the 'Response Time (seconds) Faster'. Here, we read  $t_r = 2.807$  sec.

It is apparent that the overshoot is within specifications but the rise time is too long. Need to do some tuning adjustments manually.

### 7.3 MANUAL ADJUSTMENT OF THE PID TUNER

To tune the system manually, use the 'Response time (seconds)' slider. When the slider is moved to the right, the response time becomes faster.

Under 'Options', check the box 'Show Block Response' to display the original response ('Block response') besides the 'Tuned response'.

Move the slider until the time  $t_r$  shown in the small window on the right is less than 1.5 sec (Figure 29).

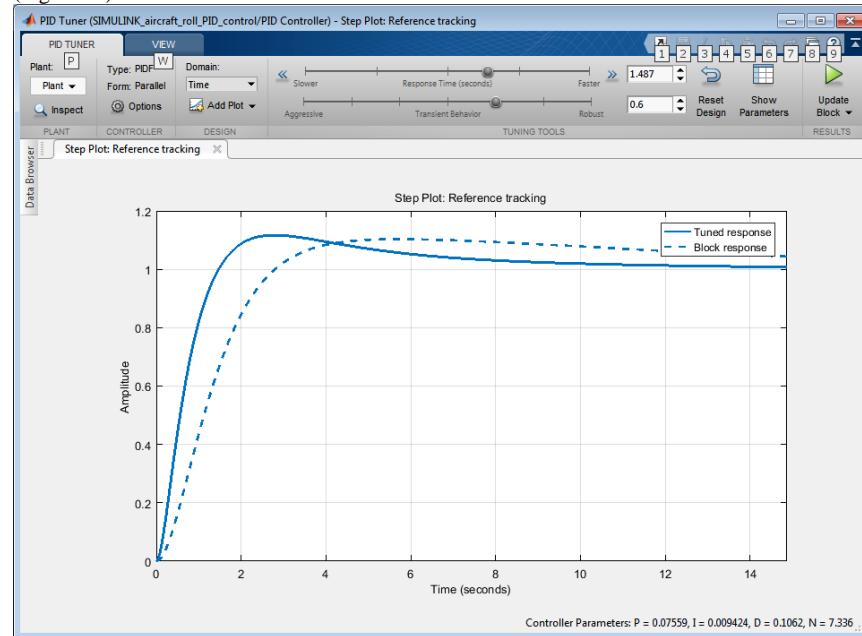


Figure 29

We read  $t_r = 1.487$  sec

At this moment, the ‘Block Parameters PID’ window looks as shown in Figure 30.

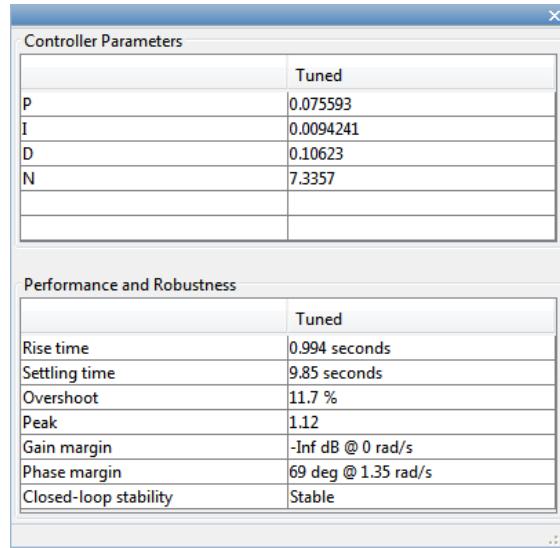


Figure 30

In this window, we read ‘Overshoot = 11.7%’ which means  $M_p = 11.7\%$  which is less than the required 20%.

It is now apparent that the design specification have been met. The final PID settings are read from the ‘Block Parameters PID’ window as:

- ‘P= 0.075593’, i.e.,  $K_p = 0.075593$
- ‘I = 0.0094241’, i.e.,  $K_i = 0.0094241$
- ‘D = 0.10623’,  $K_d = 0.10623$
- N = 7.3357

The aircraft roll response now meets the design specifications, i.e.,

- DS1: Fast response time as measured by rise time  
 $t_r = 1.487 < 1.5$  sec
- DS2: maximum percentage overshoot for step input less than 20%  
 $M_p = 11.7 < 20\%$

#### 7.4 VERIFICATION OF THE PID TUNING PROCESS

In the ‘PID Tuner’ window, press the ‘Update Block’ green arrow. Look in the ‘Block Parameters: PID Controller’ window to verify that the PID parameters have been updated. The window should look like Figure 31.

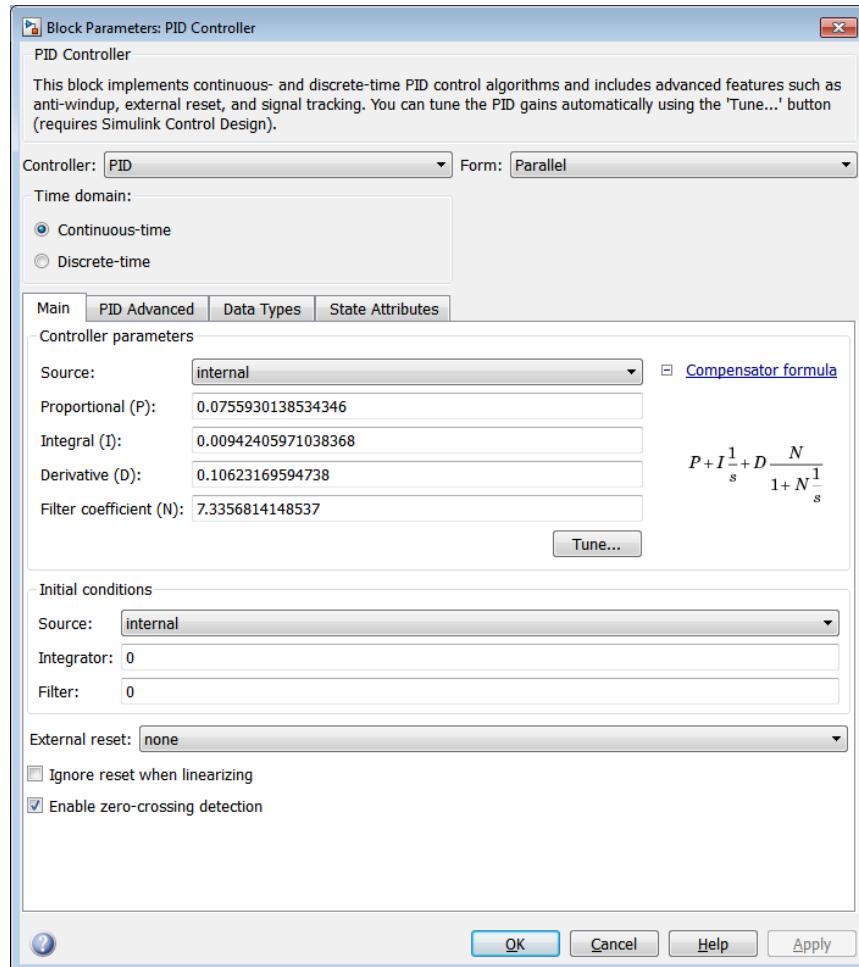


Figure 31

Now, reduce the SIMULINK run time to 20 sec and run the model (Figure 32).

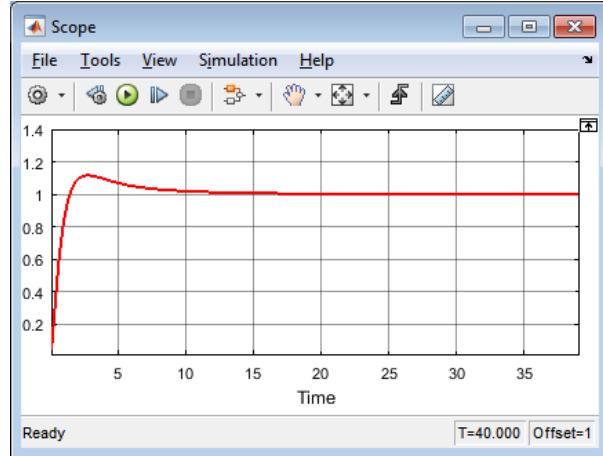


Figure 32

Press the ‘Cursor Measurements’ button. The window should look like Figure 33.

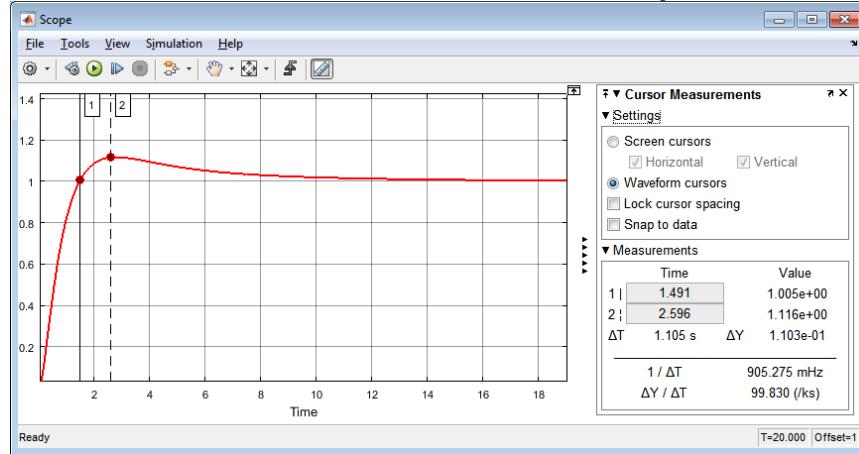


Figure 33

Measure the rise time and overshoot.

- $t_r = 1.483 \leq 1.5$  sec
- $M_p = 11.6 \leq 20\%$

It appears that the two design specifications are satisfied.

To verify the MATLAB 10%--90% rise time of 0.994 seconds indicated in the previously shown ‘Controller Parameters’ window, try to measure this rise time in the ‘Scope’ window.

To do this measurement, we need higher resolution; hence, reduce the run time to 5 sec and run again. The Scope window should look like Figure 34.

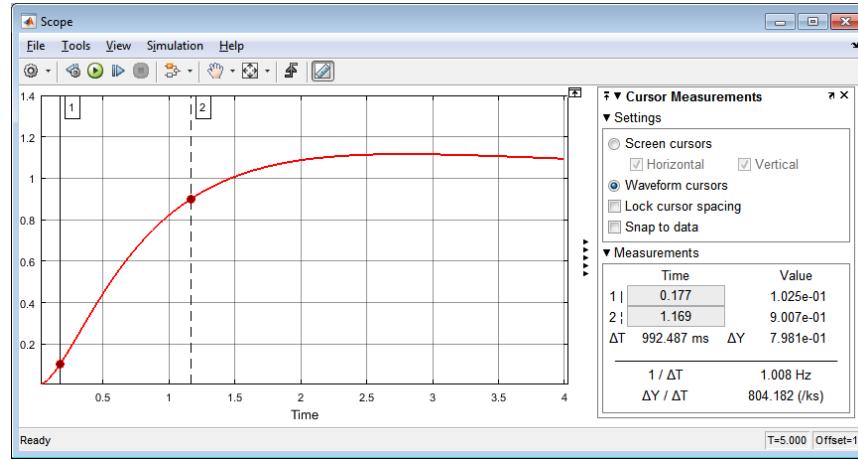
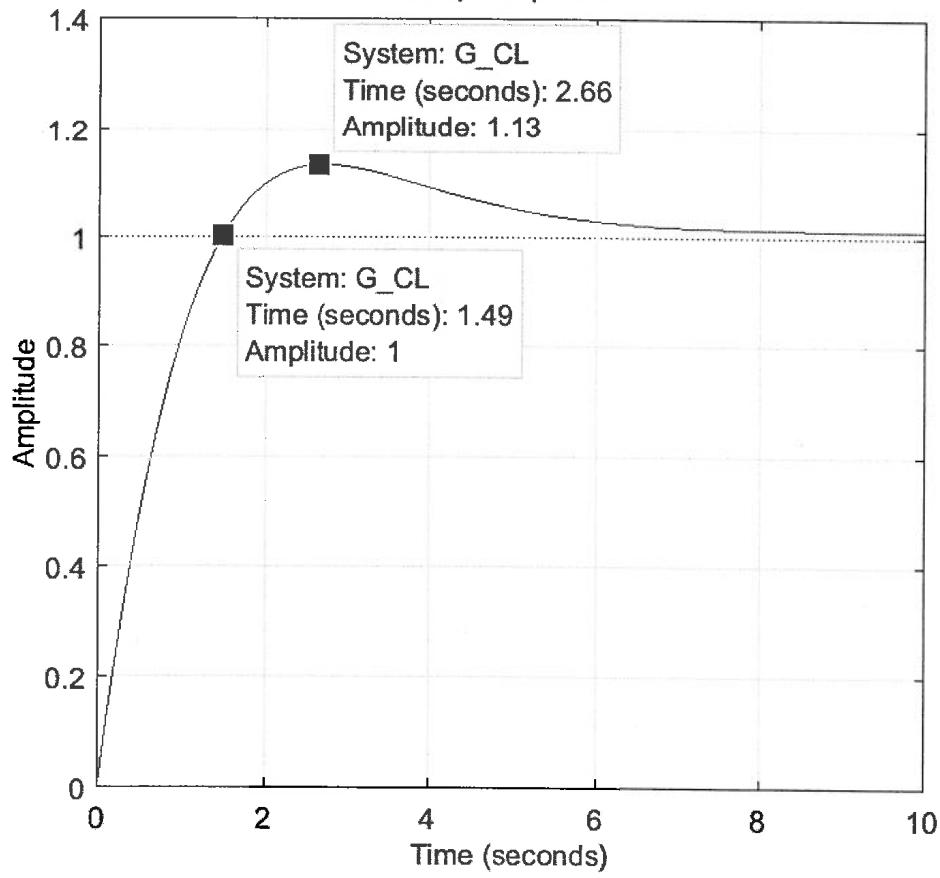
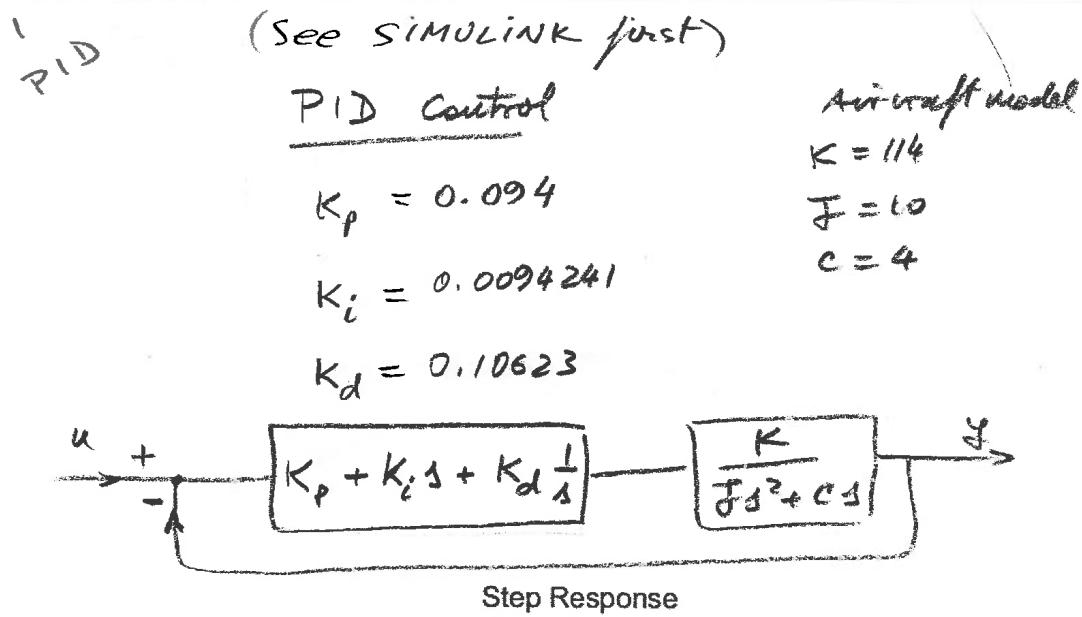


Figure 34

The cursors 1 and 2 are used to identify the 10% and 90% points on the curve. To determine the rise time from this image, read the time difference between the two cursors, i.e.,  $\Delta T = 992.487$  ms. This is the rise time between 10% and 90%. It agrees with the 0.994 seconds indicated in the previously shown ‘Controller Parameters’ window.

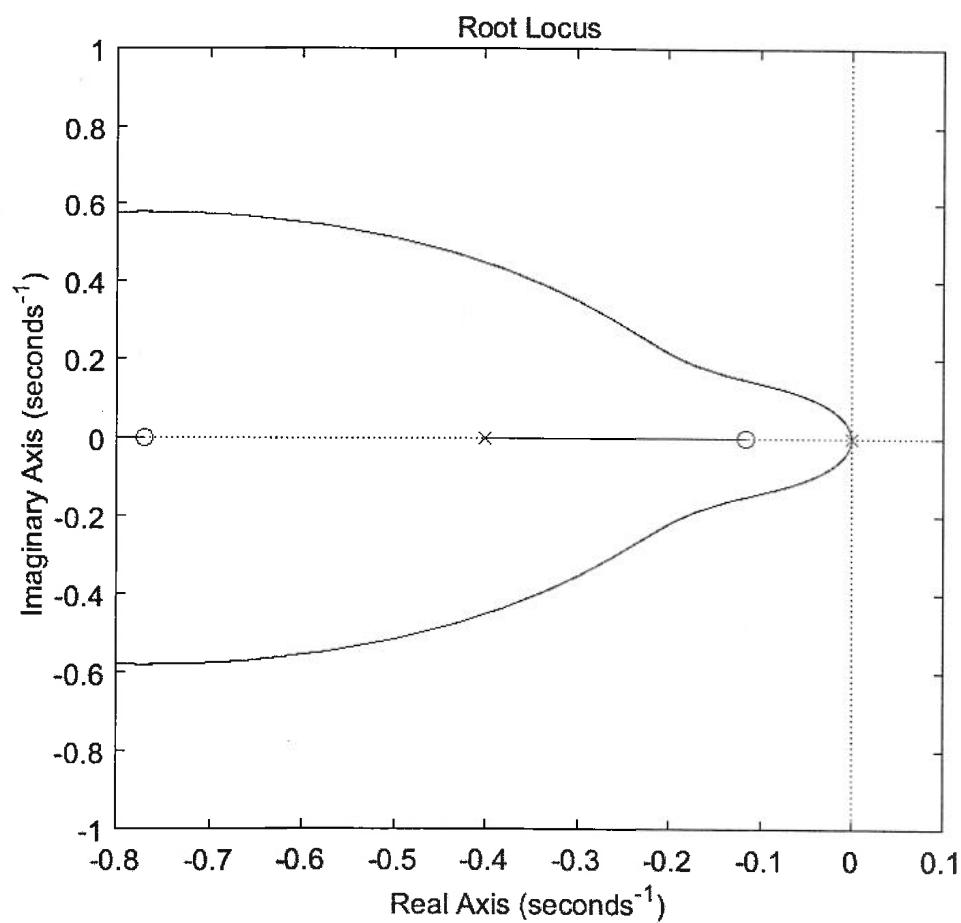
## 7.8 Feedback Controllers (continued)



2  
PID

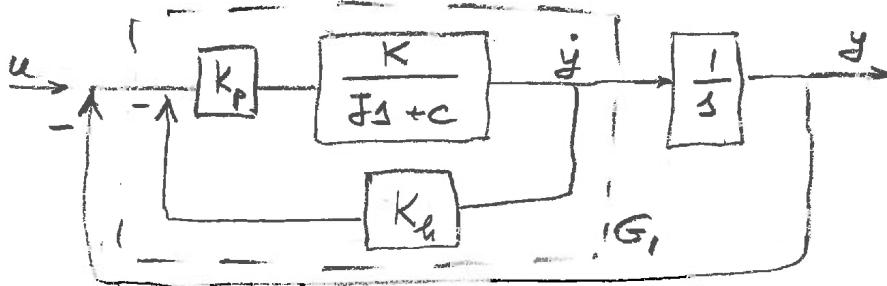
PID control

Root locus



## Velocity Feedback P-Controller

$$G(s) = \frac{K}{Js + c} = \frac{K}{Js + c} \cdot \frac{1}{1}$$



$$G_1(s) = \frac{\frac{K_p K}{Js + c}}{1 + \frac{K_p K}{Js + c} K_h} = \frac{K_p K}{Js + c + K_p K K_h} \quad (1)$$



$$G_2 = G_1(s) \frac{1}{s} = \frac{K_p K}{Js^2 + (c + K_p K K_h)s} \quad (2)$$



$$G_{CL} = \frac{G_2}{1 + G_2} = \frac{K_p K}{Js^2 + (c + K_p K K_h)s + K_p K} \quad c^*$$

$$G_{CL} = \frac{K_p K}{Js^2 + c^* s + K_p K} \quad (3)$$

2

$$c^* = c + K_p K K_h \quad (4)$$

$$G_{CL} = \frac{\frac{K_p K}{J}}{s^2 + \frac{c^*}{J}s + \frac{K_p K}{J}} = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

$$\omega_n^2 = \frac{K_p K}{J} \rightarrow \omega_n = \sqrt{\frac{K_p K}{J}} \quad (5)$$

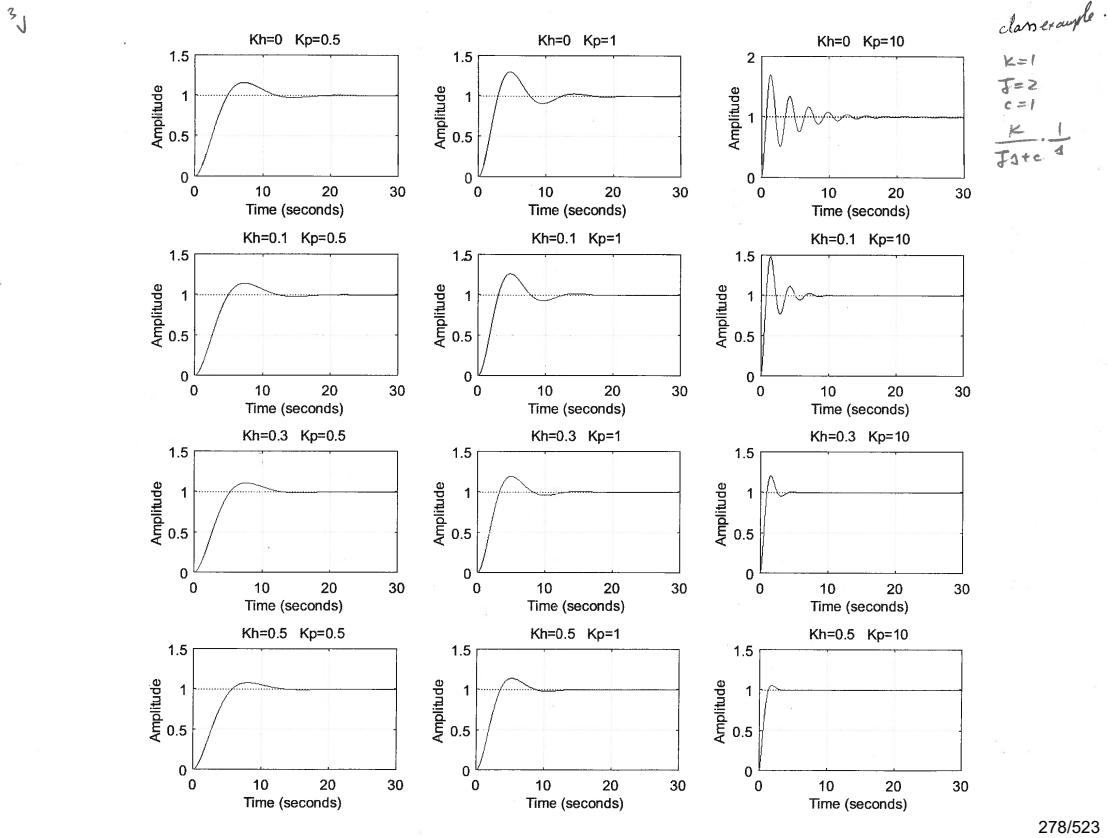
$$2\zeta\omega_n = \frac{c^*}{J} \rightarrow \zeta = \frac{c^*}{2\omega_n J} = \frac{c^*}{2\sqrt{K_p K J}}$$

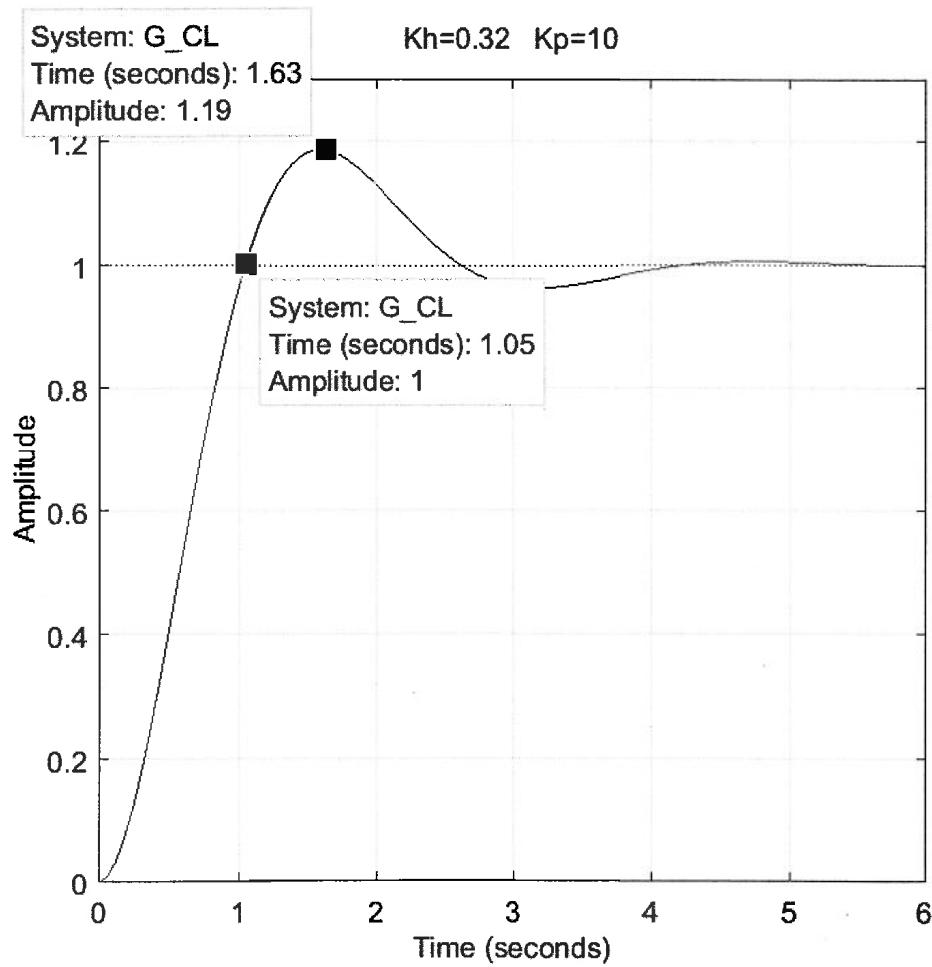
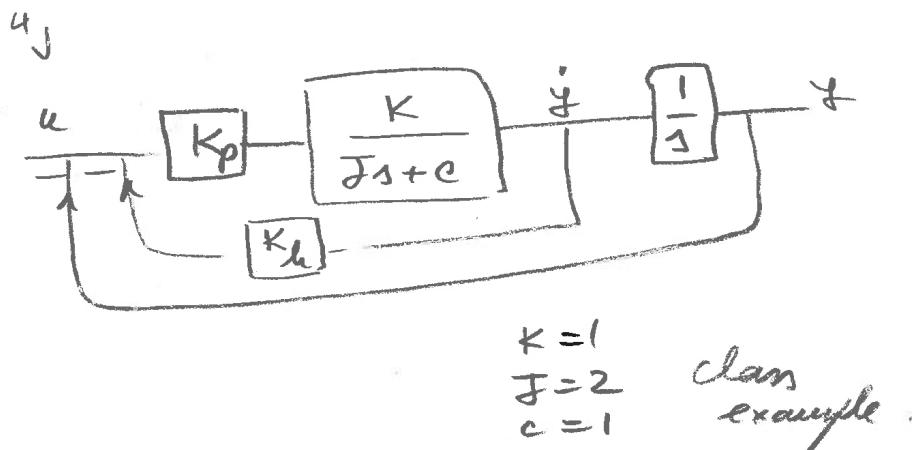
$$\zeta = \frac{c + K_p K K_h}{2\sqrt{K_p K J}}$$

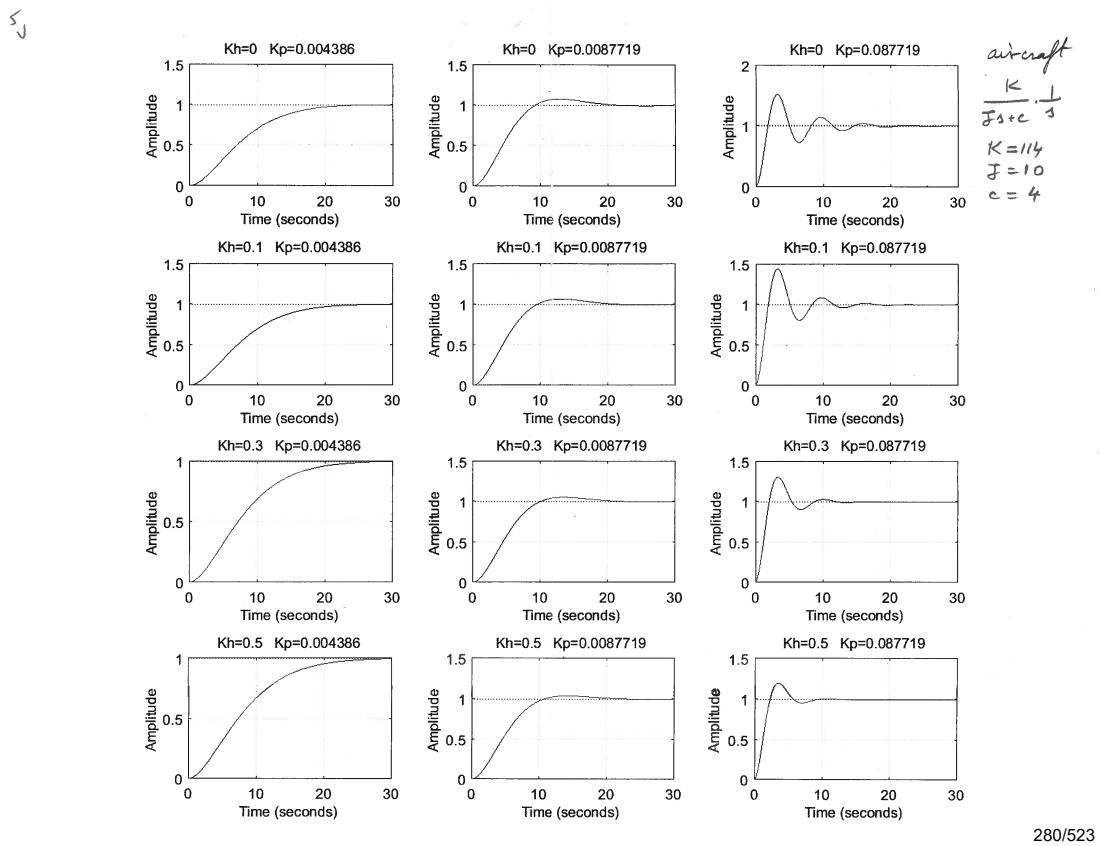
- Velocity feedback gain  $K_h$  increases  $\zeta$  (damping)
- P-control gain  $K_p$  modifies both freq.  $\omega_n$  and damping  $\zeta$

### Strategy

- modify frequency with  $K_p$
- add more damping with  $K_h$





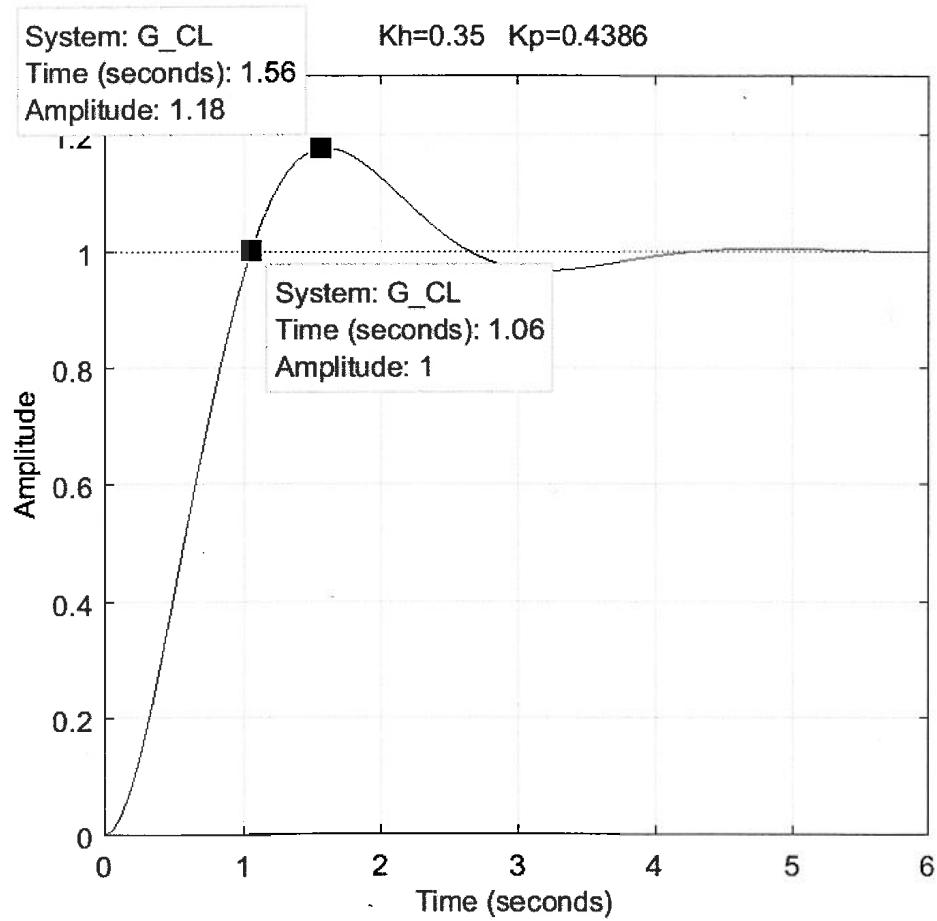


6J

velocity feedback

$$\frac{K}{Js+C} \cdot \frac{1}{s}$$

$K = 114$  aircraft  
 $J = 10$  roll  
 $C = 4$  model



281/523

## 7.9 SIMULINK Aircraft Roll Motion (continued)

## 8 VELOCITY FEEDBACK CONTROL

Velocity feedback control concept consists of two controls:

- P-control with gain  $K_p$
- velocity control with gain  $K_h$

The advantage of this control combination is that one can use the P-control gain  $K_p$  to improve the response time by increasing the natural frequency and then use the velocity feedback gain  $K_h$  to reduce the overshoot by increasing damping.

The initial values for the model are  $K_p=1/114$ ,  $K_h=0$ . With these values, the SIMULINK model for velocity feedback looks like Figure 35.

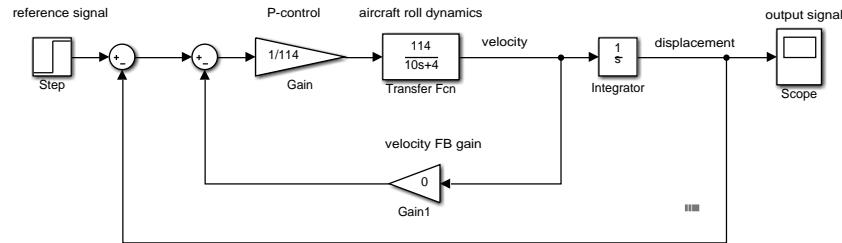


Figure 35

Note: when inserting 'Gain1' for velocity FB, use right click to flip the box to point backward.

The aircraft response is sluggish with  $t_p \approx 9$  sec , as shown in Figure 36.

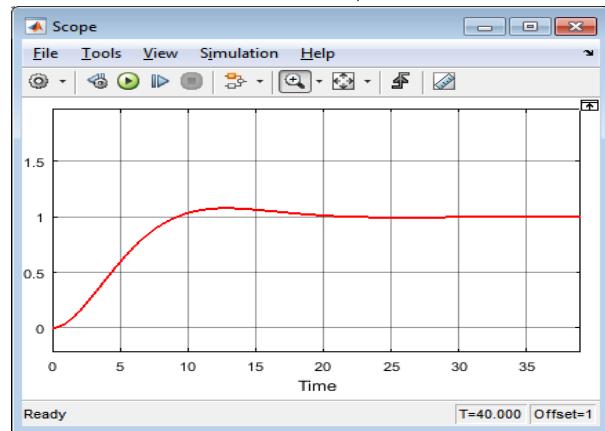


Figure 36

To accelerate the aircraft response, increase the P-control gain to  $K_p=50/114$  (Figure 37).

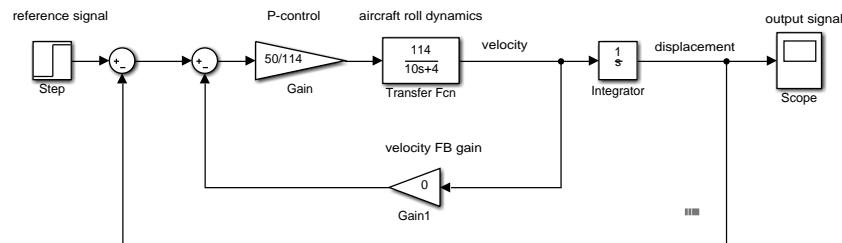


Figure 37

The aircraft response now looks like Figure 38.

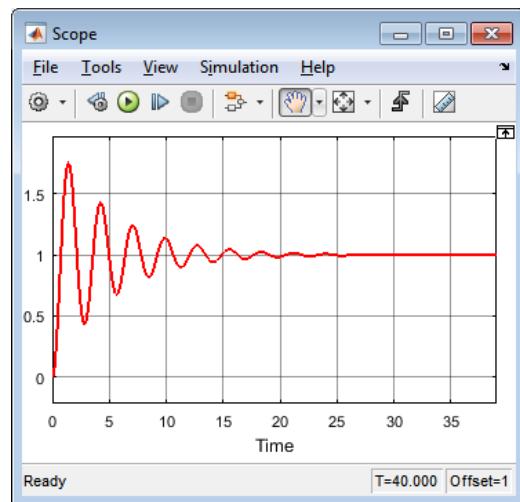


Figure 38

Notice that the aircraft response is much more rapid now. However, it has a large overshoot which must be reduced.

To reduce the overshoot, increase the velocity feedback gain to  $K_h=0.2$  (Figure 39).

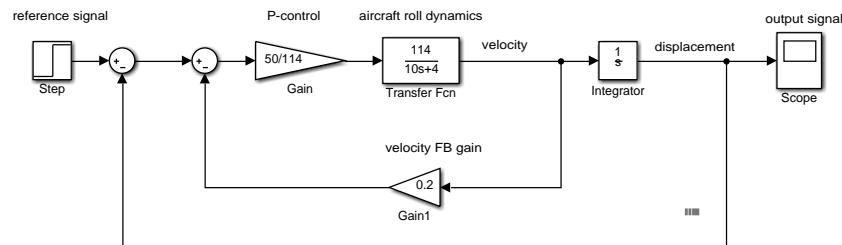


Figure 39

The response has become less oscillatory while remaining relatively fast (Figure 40).

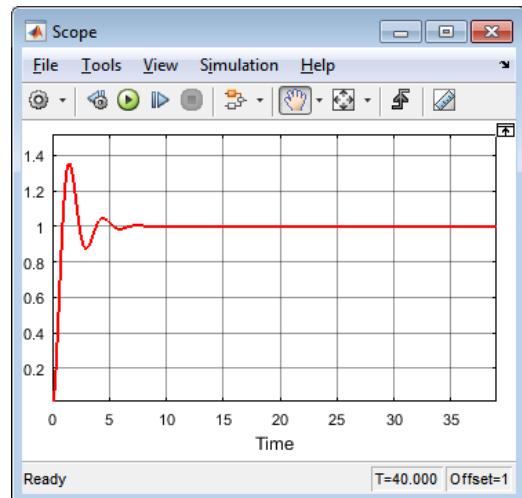


Figure 40

Further adjustments of  $K_h$  can ensure that the design specifications are met.

Reduce the computation time to 6 sec in order to expand the initial response zone and get a better reading of rise time and overshoot.

Increase  $K_h$  to various values until a satisfactory reduction of overshoot is obtained. For  $K_h=0.35$ , we get Figure 41:

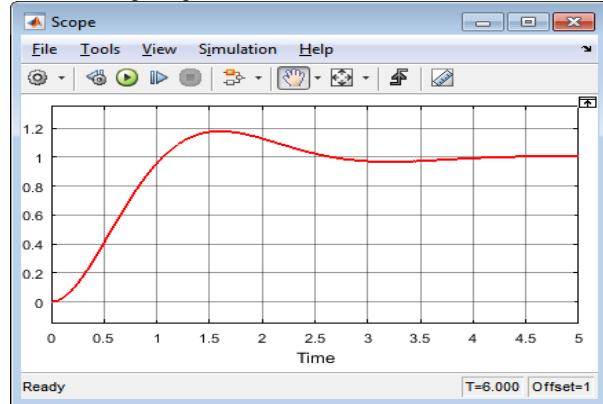


Figure 41

Open the cursors to read the rise time and overshoot(Figure 42).

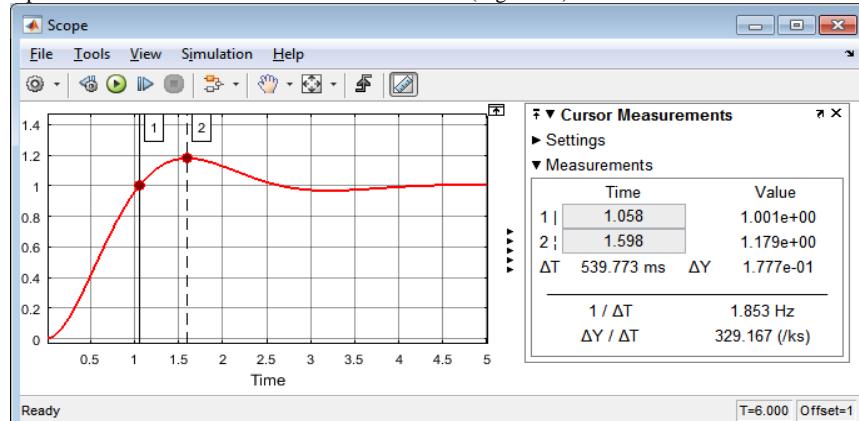


Figure 42

The readings indicate satisfactory results, i.e.,

- $t_r = 1.058 < 1.5$  sec
- $M_p = 17.9 < 20\%$

## 7.10 Instability Suppression with velocity feedback

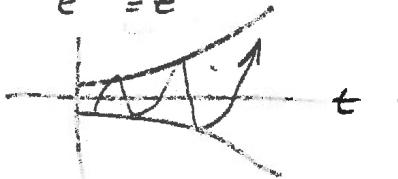
VFB  
2016/06/13

## INSTABILITY SUPPRESSION WITH VELOCITY FEEDBACK

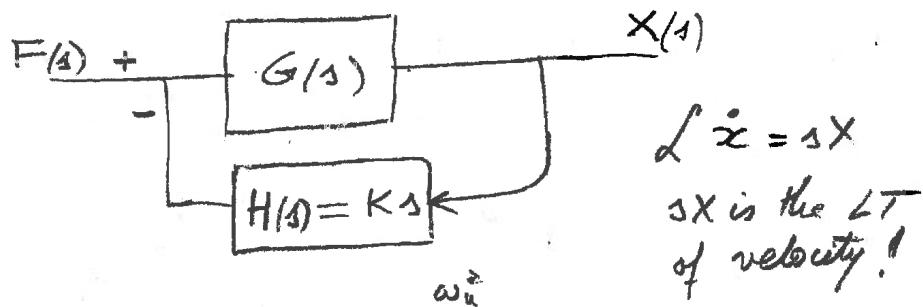
Assume a 2nd order system with negative damping,  $\zeta < 0$ . Hence, the response will be unstable:  $e^{\zeta \omega_n t} = e^{15t}$

$$\zeta < 0$$

$$-x_{st} = |\zeta| \omega_n$$



We can use FB to suppress this instability.



$$G_{CL} = \frac{G}{1+GH} = \frac{\frac{s^2 + 2\zeta\omega_n s + \omega_n^2}{\omega_n^2}}{1 + \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} Ks}$$

$$= \frac{\omega_n^2}{s^2 + (2\zeta\omega_n + K\omega_n^2)s + \omega_n^2} \quad (1)$$

$$G_{CL} = \frac{\omega_n^2}{s^2 + 2\zeta_{CL}\omega_n s + \omega_n^2} \quad (2)$$

*FB<sup>u</sup>  
2018/022 (1)&(2)*

$$2\zeta_{CL} = 2\zeta + K\omega_n$$

$$\zeta_{CL} = \zeta + \frac{K\omega_n}{2} \quad (3)$$

- 1) Velocity feedback increases damping.
- 2) If the initial system is unstable because it has -ve damping ( $\zeta < 0$ ), then velocity feedback can give the CL system a positive damping,  $\zeta_{CL} > 0$ .

Solve Eq. (3) to get

$$K = \frac{2}{\omega_n} (\zeta_{CL} - \zeta) \quad (4)$$

Critical gain  $K_{cr}$  is the gain that makes  $\zeta_{CL} = 0$ .

Recall Eq (3) and set it to zero to get

$$\zeta + \frac{K\omega_n}{2} = 0 \rightarrow K_{cr} = -\frac{2\zeta}{\omega_n} \quad (5)$$

Example:

$$\zeta = -5\% = -0.05$$

$$f_u = 4 \text{ Hz}$$

$$\omega_n = 2\pi \times 4 \approx 8\pi \text{ rad/s}$$

$$K_{cr} = -\frac{2 \times (-0.05)}{8\pi} \approx 0.004$$

For  $K > K_{cr}$ , the system is stable.

$\text{FB gain}$   
 $20/8 \cdot 10^{-2}$

### Calculation of FB gain K

Recall (4)  $K = \frac{2}{\omega_n} (\zeta_{cl} - \zeta)$

Factor out  $-\zeta$  to get  $K = -\frac{2\zeta}{\omega_n} \left(1 - \frac{\zeta_{cl}}{\zeta}\right)$

Recall (5) and write

$$K = K_{cr} \left(1 - \frac{\zeta_{cl}}{\zeta}\right) \quad (6)$$

Define  $\zeta_{ratio}$  as

$$\zeta_{ratio} = \frac{\zeta_{cl}}{\zeta} \quad (7)$$

(7)  $\rightarrow$  (6):

$$K = K_{cr} (1 - \zeta_{ratio}) \quad (8)$$

Eq (8) gives the FB gain K as a function of  $\zeta_{ratio}$

### Gain ratio $K_{ratio}$

Divide Eq (8) by  $K_{cr}$  to get

$$K_{ratio} = 1 - \zeta_{ratio} \quad (9)$$

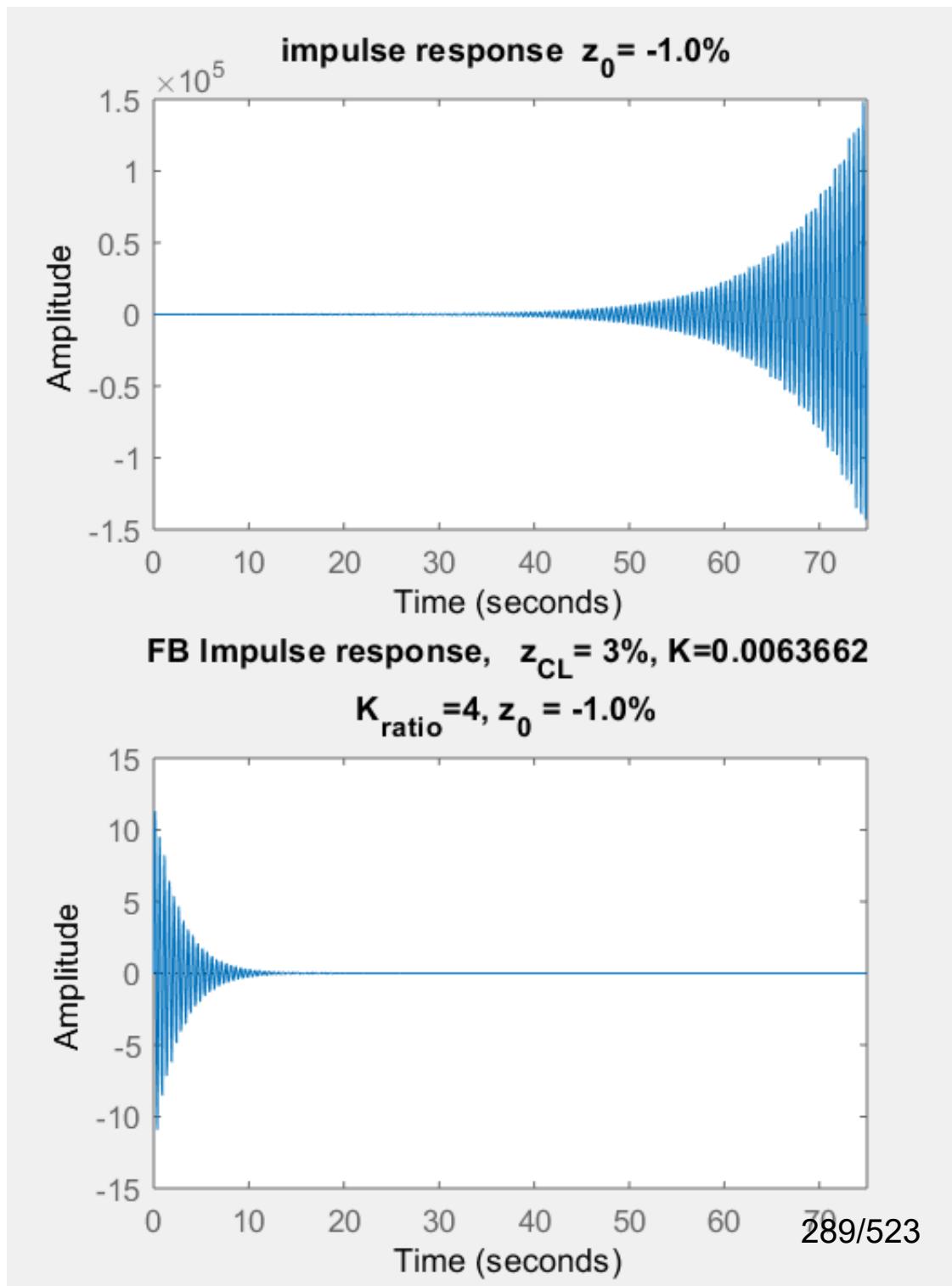
### Discussion of Eq. (8)

(a) if  $\zeta < 0$ , then  $\zeta_{ratio} < 0$ ,  $K_{cr} > 0 \rightarrow K > 0$

(b) if  $0 < \zeta < \zeta_{cl}$ , then  $K_{cr} < 0$ ,  $\zeta_{ratio} > 1$ ,  $(1 - \zeta_{ratio}) < 0$ ,  $K > 0$

(c)  $\zeta > \zeta_{cl}$ , then  $K_{cr} < 0$ ,  $\zeta_{ratio} < 1$ ,  $1 - \zeta_{ratio} < 0$ ,  $K < 0$   
 reduce  $\zeta_{cl}$

$\frac{K < 0}{288/523}$



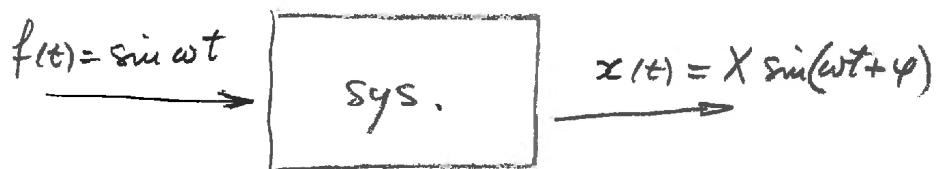
## 8 Frequency Analysis

### 8.1 Time Response Under Harmonic Excitations

2

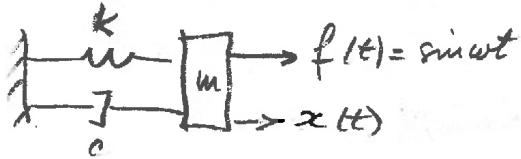
F

### Time Response to harmonic excitation



One is interested in finding how the system responds to harmonic excitation of different frequency values,  $\omega = 2\pi f$ .

Ex: For a 2<sup>nd</sup> order system, one is concerned with avoiding resonance where the response may become very large.

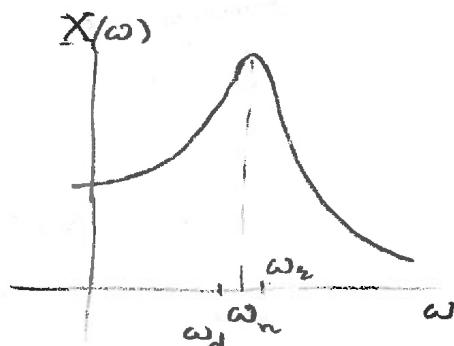


$$x(t) = X(\omega) \sin(\omega t + \varphi(\omega))$$

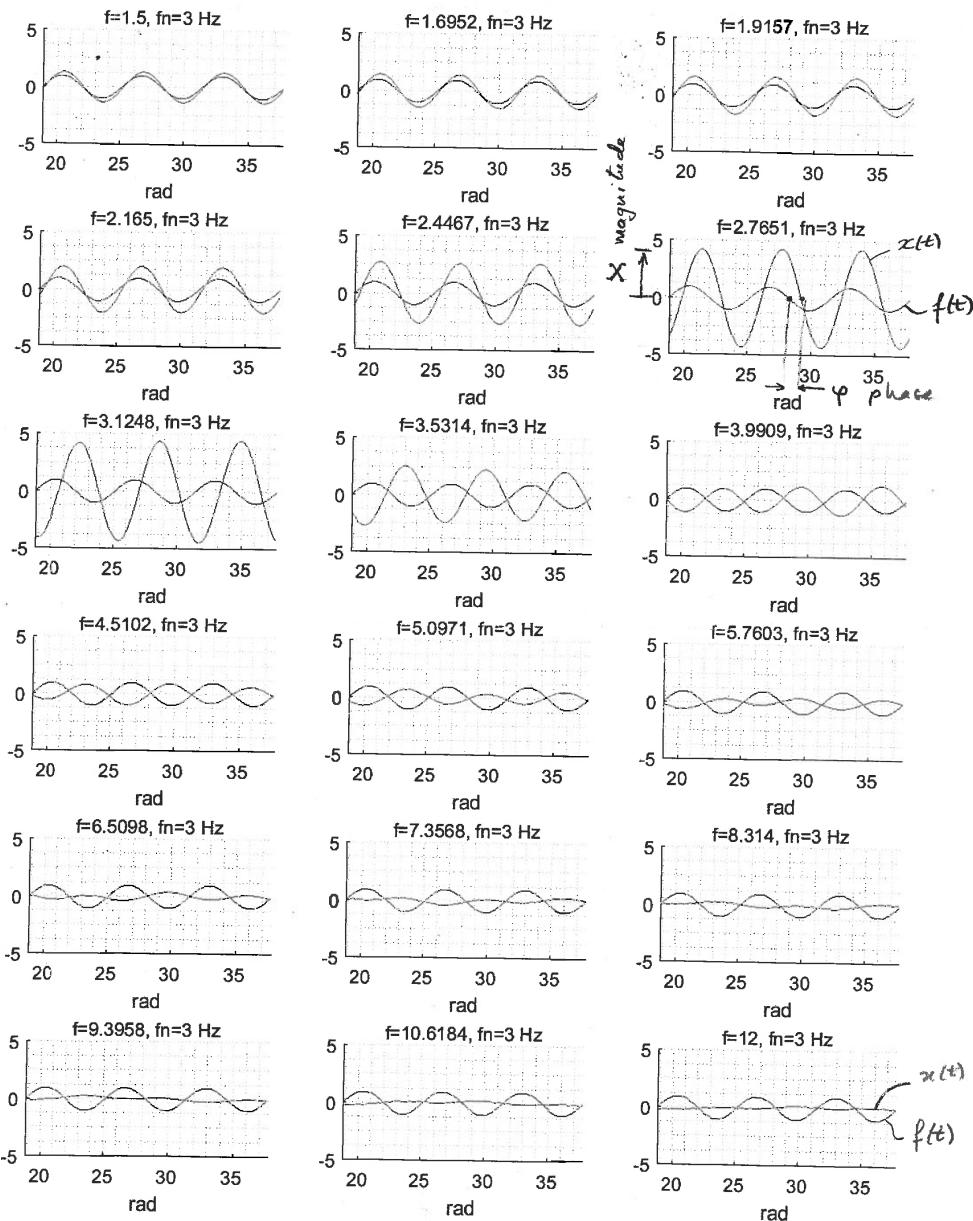
$$\zeta = \frac{c}{2\sqrt{km}}, \quad \omega_n = \sqrt{\frac{k}{m}}$$

$$\omega_d = \omega_n \sqrt{1 - \zeta^2}$$

We notice that the amplitude and phase of the time response varies with excitation freq.  $\omega$



$$f(x) = \sin(\omega t), \quad \omega = 2\pi f \quad x(t) = 2^{\text{nd}} \text{ order system response to } f(x)$$



<sup>a</sup> F

$$f(x) = \sin \omega t$$

$$x(t) = X(\omega) \sin[\omega t + \varphi(\omega)]$$

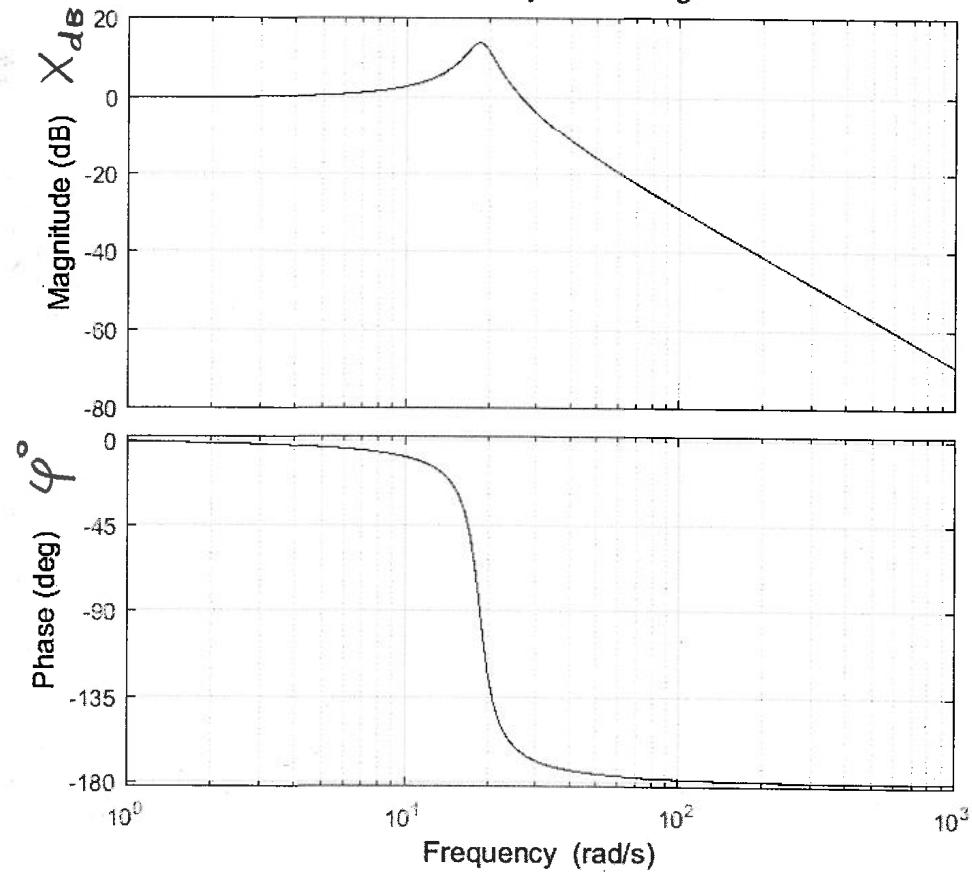
$X(\omega)$  = magnitude of response

$\varphi(\omega)$  = phase of response

Magnitude  $X$  and phase  $\varphi$  vary with excitation freq.  $\omega$ .

Bode diagram plots: variation of  $X(\omega)$  &  $\varphi(\omega)$

2nd order sys Bode diag.



Note that  $\varphi$  is -ve, i.e. the response lags behind the excitation.

5  
4

C:\Mydata\1\_USC\c...\FRF\_2ndOrderSys\_20161214.m Page 1

```
1 % Amplitude_and_phase_2ndOrderSys.m
2 % AMPLITUDE AND PHASE IN FREQUENCY RESPONSE
3 clc
4 clear
5 close all
6 s=tf('s');
7 % f=input('f=');
8 figure (1);
9 %% 2nd order system
10 fn=3; z=10e-2; wn=2*pi*fn; G2=wn^2/(s^2+2*z*wn*s+wn^2);
11 M=6; N=3; Nplots=M*N;
12 fmin=fn/2; fmax=fn*4;
13 a=log10(fmin); b=log10(fmax); f=logspace(a,b,Nplots);
14 %% plotting setup
15 Na=1e3; amax=10*2*pi; da=amax/Na; angle=0:da:amax;
16 xmin=0.3*Na*da; xmax=0.6*Na*da;
17 %% plot response at various frequencies
18 for i=1:Nplots
19 w=2*pi*f(i); % excitation frequency
20 t=angle/w; % time steps at this excitation freq.
21 A=1; % forcing function amplitude
22 F=A*w/(s^2+w^2); % Laplace transform of sine forcing function
23 fe=impulse(F,t); % time response of forcing function
24 X2=G2*F; % Laplace transform of 2nd order system response
25 subplot(M,N,i);
26 x2=impulse(X2,t); % time response of 2nd order system
27 plot(angle,fe,angle,x2); hold on
28 title(['f=' num2str(f(i)) ', fn=' num2str(fn) ' Hz' ],...
29 'FontSize', 10,'FontWeight','normal')
30 xlabel('rad'); xlim([xmin xmax]); ylim([-5*A 5*A]);
31 % grid on
32 grid minor; box off
```

6  
4

C:\Mydata\1 USC\c...\FRF\_2ndOrderSys\_20161214.m Page 2

```
33 end
34 %% FRF Bode plots
35 figure(2)
36 bode(G2); grid on; title('2nd order sys Bode diag.');
37 aw=log10(2*pi*fn/2); bw=log10(2*pi*fn*2); N=1e4; wBode=logspace(aw,bw,N);
38 figure(3)
39 bode(G2,wBode); grid on; title('zoom 2nd order sys Bode diag.');
40
41
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54
```

## 8.2 Frequency Response Function (FRF)

FRF

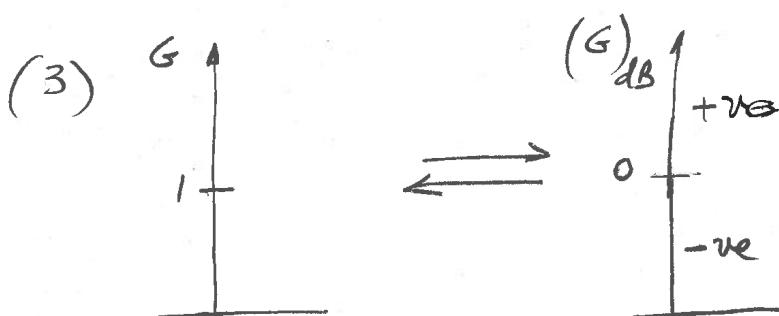
dB scaledecibell  
 $\frac{1}{10}$ dB scale is a  $\log_{10}$  scale magnified by 20

$$(G)_{dB} = 20 \log_{10} G$$

Properties

- (1) Only +ve numbers have dB value  
(cannot take log of -ve numbers!)

(2)  $(\frac{1}{G})_{dB} = -(G)_{dB}$  (reciprocal numbers)



$$G > 1$$

$(G)_{dB}$  is +ve

$$G = 1$$

$$(G)_{dB} = 0$$

$$G < 1$$

$(G)_{dB}$  is -ve

FRF

(4) Scale up or scale down in physical values means shift up or shift down in dB

→ (a) double (octave up) = +6 dB

(b) "half" (octave down) = -6 dB

(c) "ten times" (decade) = +20 dB

octave =  $\frac{A_B C D E F G}{8^{\text{musical note}}}$  has double the frequency  $f_2 = 2f_1$

Proof

$$(a) \log_{10} 2 = 0.303 ; 20 \log_{10} 2 \approx 6$$

$$G_2 = 2G_1$$

$$(G_2)_{dB} = 20 \log_{10} (2G_1) = 20 \log_{10} 2 + 20 \log_{10} G_1$$

$$= \frac{20 \times 0.303}{\approx 6} + (G_1)_{dB}$$

>> mag 2 dB (2)

ans 6.0206

$$(G_2)_{dB} = (G_1)_{dB} + 6 dB$$

$$(b) G_2 = \frac{1}{2} G_1$$

$$(G_2)_{dB} = (G_1)_{dB} + 20 \log_{10} \left(\frac{1}{2}\right)$$

$$= (G_1)_{dB} + 20 \log_{10} 2$$

$$= (G_1)_{dB} - 6 dB$$

<sup>3</sup>  
FRF

(C)  $G_2 = 10 G_1$

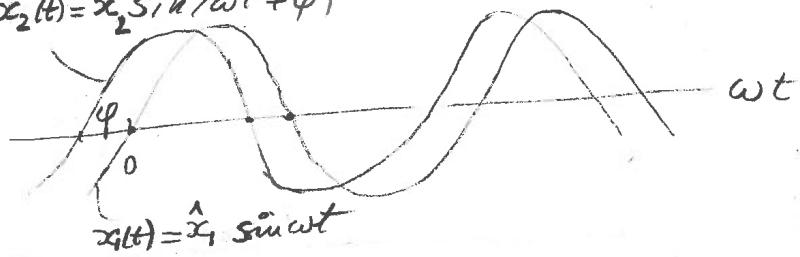
$$(G_2)_{dB} = (G_1)_{dB} + 20 \log_{10}^{=1}$$
$$= (G_1)_{dB} + 20 dB$$

FRFPhase

$$x_1(t) = \hat{x}_1 \sin \omega t \quad \text{reference signal}$$

$$x_2(t) = \hat{x}_2 \sin(\omega t + \varphi)$$

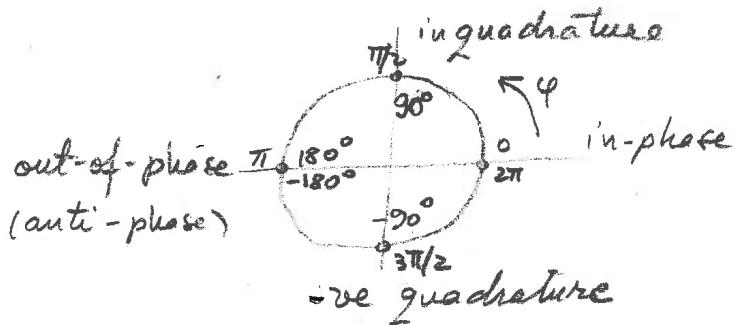
$$x_2(t) = \hat{x}_2 \sin(\omega t + \varphi)$$



$$\psi_1 = \omega t \quad \text{phase of } x_1 \text{ (reference)}$$

$$\psi_2 = \omega t + \varphi \quad \text{phase of } x_2$$

$$\varphi = \psi_2 - \psi_1 \quad \begin{array}{l} \text{phase difference} \\ \text{x}_2 \text{ leads by } \varphi \end{array}$$

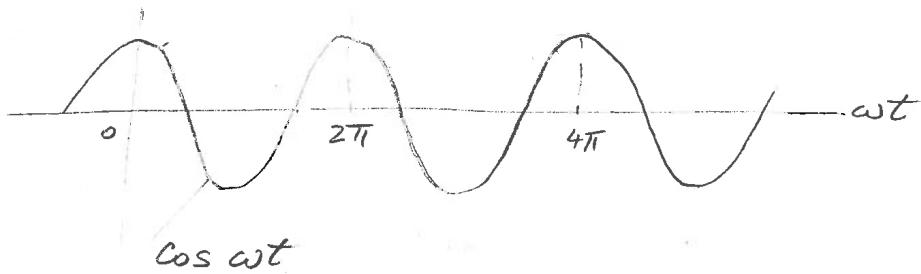
Periodic signals

$$\left\{ \begin{array}{l} 2\pi \\ 360^\circ \end{array} \right. \text{ periodicity: } \bullet x(\omega t + 2\pi) = x(\omega t)$$

$$\bullet x(\omega t + \pi) = x(\omega t - \pi)$$

5  
FRF

## COMPLEX NUMBER REPRESENTATION OF HARMONIC SIGNALS



$$e^{i\omega t} = \cos \omega t + i \sin \omega t \quad (\text{Euler})$$

$$\left\{ \begin{array}{l} x(t) = \hat{x} \cos(\omega t + \varphi) \\ x(t) = \operatorname{Re}[\hat{x} e^{i(\omega t + \varphi)}] \end{array} \right.$$

$$\left\{ \begin{array}{l} x(t) = \hat{x} \sin(\omega t + \varphi) \\ x(t) = \operatorname{Im}[\hat{x} e^{i(\omega t + \varphi)}] \end{array} \right.$$

In general  $x(t) = \hat{x} e^{i(\omega t + \varphi)}$

$$\hat{x} e^{i(\omega t + \varphi)} = \hat{x} e^{i\varphi} e^{i\omega t} = X e^{i\omega t}$$

$$X = \hat{x} e^{i\varphi} \quad (\text{phasor})$$

$$|X| = \hat{x}$$

magnitude of  $x(t)$

$$|X| = \operatorname{abs}(X)$$

$$\angle X = \varphi$$

phase of  $x(t)$

$$\angle X = \operatorname{angle}(X), \text{ rad}$$

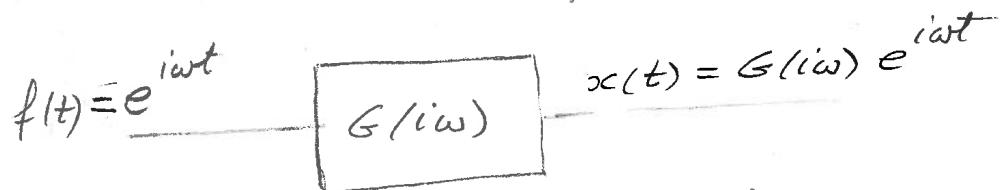
$$2 \text{ rad} \equiv \text{deg}(\pi) = 180^\circ$$

302/523

<sup>b</sup>  
FRFFrequency response function (FRF)

- $G(s)$  TF: Transfer function in Laplace domain  $s$

- $G(i\omega)$  FRF: Transfer function in frequency domain  $\omega$



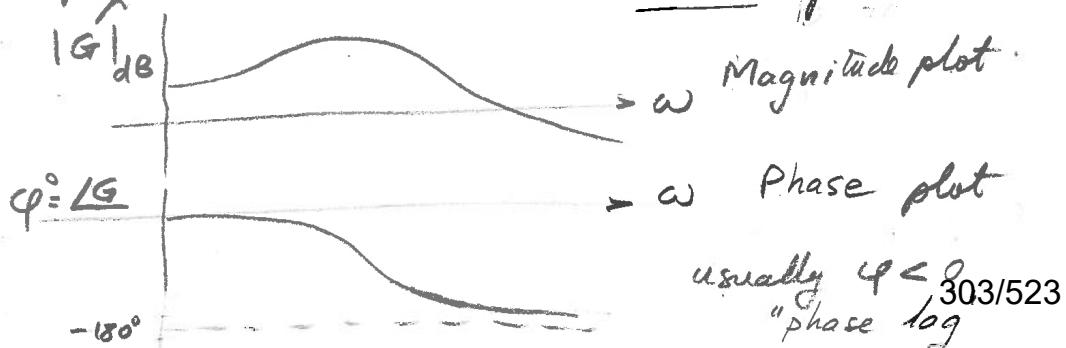
$G(i\omega)$  is a complex number

$|G(i\omega)|$  = magnitude, dB

$\angle G(i\omega)$  = phase, deg

Frequency response function (FRF) is  
 $G(i\omega)$  measured over a range of  
frequencies  $\omega$

Bode diagram



*2018/2019* Harmonic Response via Laplace Transform

$$f(t) = e^{i\omega t} \xrightarrow{\mathcal{L}} F(s) = \frac{1}{s - i\omega}$$

$$X(s) = G(s) F(s) = G(s) \frac{1}{s - i\omega} \quad (1)$$

$$\text{Assume: } G(s) = K \frac{(s - z_1)(s - z_2) \dots (s - z_m)}{(s + p_1)(s + p_2) \dots (s + p_n)} \quad (2)$$

(1) & (2) yields the PFE:

$$X(s) = \frac{a}{s - i\omega} + \frac{b_1}{s + p_1} + \frac{b_2}{s + p_2} + \dots + \frac{b_n}{s + p_n} \quad (3)$$

$$x(t) = a e^{i\omega t} + b_1 e^{-p_1 t} + b_2 e^{-p_2 t} + \dots + b_n e^{-p_n t} \quad (4)$$

↑  
steady state  
response.

Transient response  $\xrightarrow[t \rightarrow \infty]{\longrightarrow} 0$

$$x_{ss}(t) = a e^{i\omega t} \quad (5)$$

To find  $a$ , multiply (3) by  $s - i\omega$  and make

$s = i\omega$  to get:

$$\left[ (s - i\omega) X(s) \right]_{s=i\omega} = \left[ a + \left[ \frac{b_1}{s + p_1} + \frac{b_2}{s + p_2} + \dots + \frac{b_n}{s + p_n} \right] (s = i\omega) \right]_{s=i\omega} \quad (6)$$

$$a = \left[ (s - i\omega) X(s) \right]_{s=i\omega} \quad (6)$$

$$= \left[ (s + i\omega) G(s) \frac{1}{(s - i\omega)} \right]_{s=i\omega} = G(i\omega).$$

$$(6) \rightarrow (5): \quad x_{ss}(t) = \boxed{G(i\omega)} e^{i\omega t}$$

complex amplitude

$$\text{FRF} = G(i\omega) = |G(i\omega)| e^{i\varphi}, \quad \varphi = \angle G(i\omega)$$

304/523

FR14/

## Harmonic Response via Laplace Transform

$$G(s) = \frac{1}{Ts+1} , f(t) = e^{i\omega t} \longrightarrow F(s) = \frac{1}{s-i\omega}$$

$$X(s) = G(s) F(s) = \frac{1}{Ts+1} \cdot \frac{1}{s-i\omega} = \frac{1}{T} \cdot \frac{1}{s+\frac{1}{T}} \cdot \frac{1}{s-i\omega}$$

PFE, p 30, Eq (2.6), modified:  $s_1 = -\frac{1}{T}$   $s_2 = i\omega$

$$X(s) = \frac{a_1}{s-s_1} + \frac{a_2}{s-s_2} + \dots + \frac{a_k}{s-s_k} + \dots$$

$$a_k = \left[ (s-s_k) X(s) \right]_{s=s_k}$$

$$a_1 = \left[ \cancel{\left( s + \frac{1}{T} \right)} \frac{1}{T} \frac{1}{s+\frac{1}{T}} \frac{1}{s-i\omega} \right]_{s=-\frac{1}{T}} = \frac{1}{T} \frac{1}{-\frac{1}{T}-i\omega} = -\frac{1}{i\omega T + 1}$$

$$a_2 = \left[ \cancel{(s+i\omega)} \frac{1}{T} \frac{1}{s+\frac{1}{T}} \frac{1}{s-i\omega} \right]_{s=i\omega} = \frac{1}{T} \frac{1}{i\omega + \frac{1}{T}} = \frac{1}{i\omega T + 1}$$

$$X(s) = -\frac{1}{i\omega T + 1} \cdot \frac{1}{s+\frac{1}{T}} + \frac{1}{i\omega T + 1} \cdot \frac{1}{s-i\omega}$$

$$x(t) = -\frac{1}{i\omega T + 1} e^{-t/T} + \frac{1}{i\omega T + 1} e^{i\omega t}$$

| transient | steady state |

$$x_{ss}(t) = \frac{1}{i\omega T + 1} e^{i\omega t} = G(i\omega) e^{i\omega t}$$

$$FRF = G(i\omega) = |G(i\omega)| e^{i\varphi}, \varphi = \underline{|G(i\omega)|}$$

True for stable systems (i.e., transients vanish). 305/523

FR15

Harmonic response via Laplace Transform  
for sine excitation

$$G(s) = \frac{1}{Ts+1} \quad \rightarrow f(t) = \sin \omega t \rightarrow F(s) = \frac{\omega}{s^2 + \omega^2}$$

$$X(s) = G(s) F(s) = \frac{1}{Ts+1} \cdot \frac{\omega}{s^2 + \omega^2} = \frac{1}{T} \frac{1}{s + \frac{1}{T}} \cdot \frac{\omega}{(s - i\omega)(s + i\omega)}$$

Poles:  $s_1 = -\frac{1}{T}$   $s_{2,3} = \pm i\omega$

PFE, p.30, Eq. (2.6), modified:

$$X(s) = \frac{a_1}{s - s_1} + \frac{a_2}{s - s_2} + \dots + \frac{a_k}{s - s_k} + \dots$$

$$a_k = \left[ (s - s_k) X(s) \right]_{s=s_k}$$

$$X(s) = \frac{a_1}{s + \frac{1}{T}} + \frac{a_2}{s - i\omega} + \frac{a_3}{s + i\omega}$$

$$a_1 = (s + \frac{1}{T}) G(s) \Big|_{s=s_1} = (s + \frac{1}{T}) \frac{\frac{1}{T}}{\cancel{s + \frac{1}{T}}} \frac{\omega}{\cancel{s^2 + \omega^2}} \Big|_{s=-\frac{1}{T}}$$

$$a_1 = \frac{1}{T} \frac{\omega}{\frac{1}{T^2} + \omega^2} = \frac{\omega T}{\omega^2 T^2 + 1}$$

$$a_2 = (s - i\omega) G(s) \Big|_{s=i\omega} = (s - i\omega) \frac{1}{Ts+1} \frac{\omega}{(s - i\omega)(s + i\omega)} \Big|_{s=i\omega}$$

$$a_2 = \frac{1}{i\omega T + 1} \cdot \frac{\omega}{i\omega + i\omega} = \frac{1}{i\omega T + 1} \cdot \frac{1}{2i} = G(i\omega) \frac{1}{2i}$$

FR16

$$a_3 = (s + i\omega) G(s) \Big|_{s=-i\omega} = (s+i\omega) \frac{1}{T_s+1} \frac{\omega}{(s-i\omega)(s+i\omega)} \Big|_{s=-i\omega}$$

$$= \frac{1}{-i\omega T+1} \cdot \frac{1}{-2i} = -G(-i\omega) \frac{1}{2i}$$

$$X(s) = \frac{\omega T}{\omega^2 T^2 + 1} \frac{1}{s + \frac{1}{T}} + \frac{1}{2i} \left[ \frac{G(i\omega)}{s - i\omega} - \frac{G(-i\omega)}{s + i\omega} \right]$$

— transient — | — steady state — |

$$x(t) = \frac{\omega T}{\omega^2 T^2 + 1} e^{-t/T} + \frac{1}{2i} \left[ G(i\omega) e^{i\omega t} - G(-i\omega) e^{-i\omega t} \right]$$

But  $G(i\omega) = |G(i\omega)| e^{i\varphi}$ ,  $\varphi = \angle G(i\omega)$

$$G(-i\omega) = |G(i\omega)| e^{-i\varphi}$$

Hence

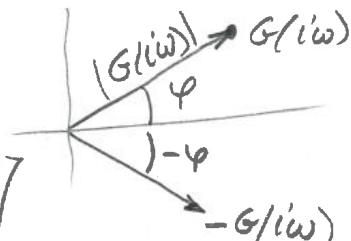
$$x_{ss}(t) = \frac{1}{2i} \left[ G(i\omega) e^{i\omega t} - G(-i\omega) e^{-i\omega t} \right]$$

$$= \frac{1}{2i} |G(i\omega)| \left( e^{i\varphi} e^{i\omega t} - e^{-i\varphi} e^{-i\omega t} \right), \varphi = \angle G(i\omega)$$

$$= |G(i\omega)| \frac{e^{i\varphi} - e^{-i\varphi}}{2i}, \varphi = \omega t + \varphi$$

$$= |G(i\omega)| \sin \varphi$$

$$x_{ss}(t) = |G(i\omega)| \sin(\omega t + \varphi), \varphi = \angle G(i\omega)$$



FR17Lemmas:

If  $G(s)$  is polynomial (or fraction of)

and  $G(i\omega) = |G(i\omega)| e^{i\varphi}$

Then  $G(-i\omega) = |G(i\omega)| e^{-i\varphi}$

Proof

$$(a) \quad G(s) = s + a$$

$$G(i\omega) = i\omega + a = \sqrt{a^2 + \omega^2} e^{i\varphi}, \quad \varphi = \tan^{-1} \frac{\omega}{a}$$

$$G(-i\omega) = -i\omega + a = \sqrt{a^2 + \omega^2} e^{i\varphi^*}, \quad \varphi^* = \tan^{-1} \frac{-\omega}{a} = -\varphi$$

$$(b) \quad G(s) = (s + a_1)(s + a_2)$$

$$G(i\omega) = (i\omega + a_1)(i\omega + a_2)$$

$$= \sqrt{a_1^2 + \omega^2} \sqrt{a_2^2 + \omega^2} e^{i\varphi_1} e^{i\varphi_2}, \quad \varphi_1 = \tan^{-1} \frac{\omega}{a_1}$$

$$\varphi_2 = \tan^{-1} \frac{\omega}{a_2}$$

$$G(-i\omega) = (-i\omega + a_1)(-i\omega + a_2)$$

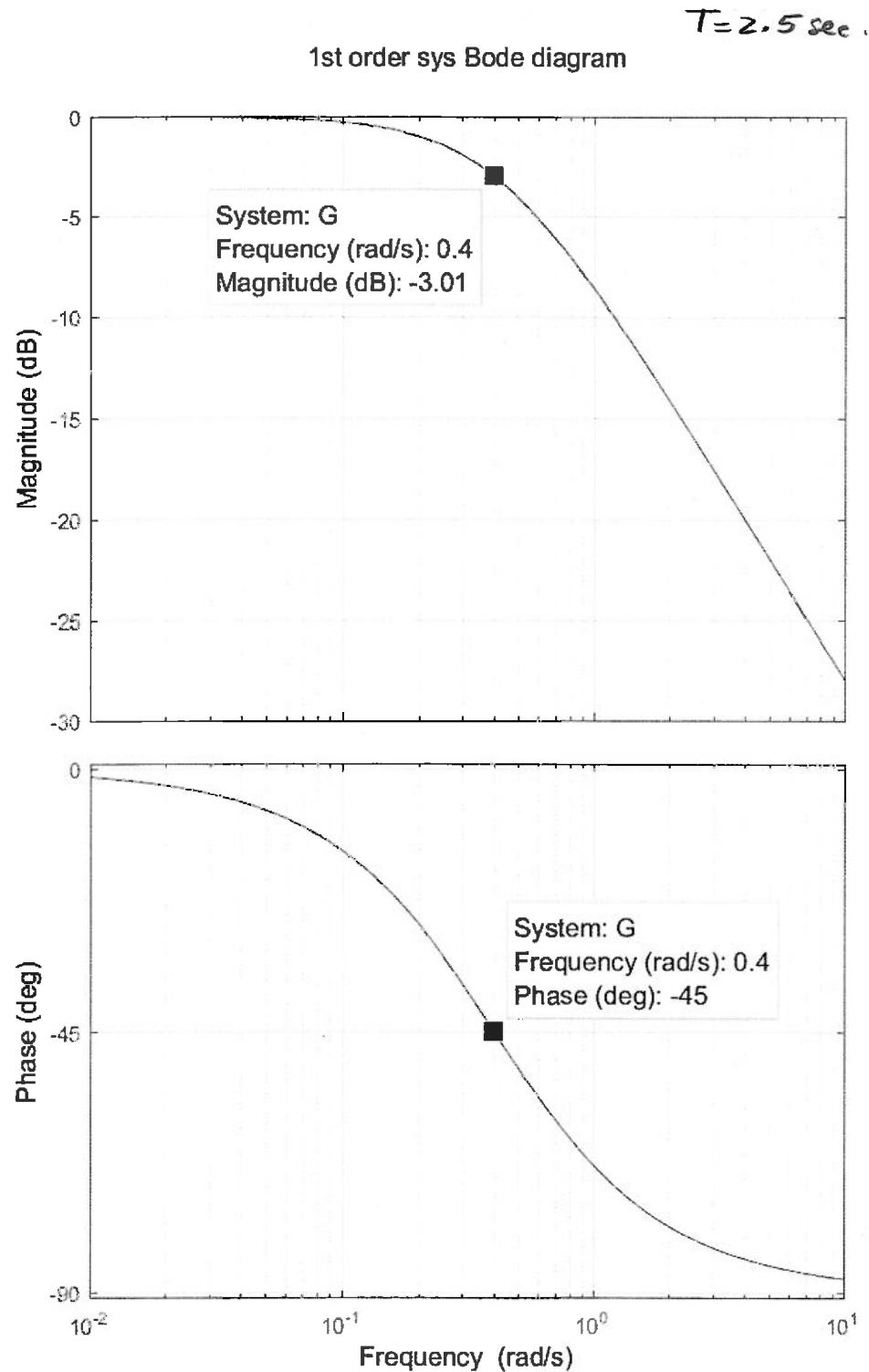
$$= \sqrt{a_1^2 + \omega^2} \sqrt{a_2^2 + \omega^2} e^{i\varphi_1^*} e^{i\varphi_2^*}$$

$$\varphi_1^* = \tan^{-1} \frac{-\omega}{a_1} = -\varphi_1$$

$$\varphi_2^* = \tan^{-1} \frac{-\omega}{a_2} = -\varphi_2$$

etc.

1<sup>st</sup> order system FRF



### 1<sup>st</sup> Order System FRF

$$G(s) = \frac{1}{Ts+1} \quad \text{transfer function (TF)}$$

$$G(i\omega) = \frac{1}{i\omega T + 1} \quad \text{frequency response function (FRF)}$$

Asymptotes

define  $\omega_c = 1/T$

Low freq. asymptote  $\omega \ll \omega_c$   $G(i\omega) \rightarrow G_{LF}$   
 $\omega T \ll 1$

$$G(i\omega) = \frac{1}{i\omega T + 1} \xrightarrow{\omega T \ll 1} \frac{1}{1} = 1$$

$$\therefore G_{LF}(i\omega) = 1, |G_{LF}|_{dB} = 0 \quad \angle G_{LF} = 0^\circ$$

High freq. asymptote  $\omega > \omega_c$   $\omega T \gg 1$   $G(i\omega) \rightarrow G_{HF}$

$$G(i\omega) = \frac{1}{i\omega T + 1} \xrightarrow{\omega T \gg 1} \frac{1}{i\omega T}$$

$$\therefore G_{HF}(i\omega) = \frac{1}{i\omega T} = \frac{1}{\omega T} e^{-i\frac{\pi}{2}}$$

$$|G_{HF}| = \frac{1}{\omega T} \quad \angle G_{HF} = -\frac{\pi}{2} = -90^\circ$$

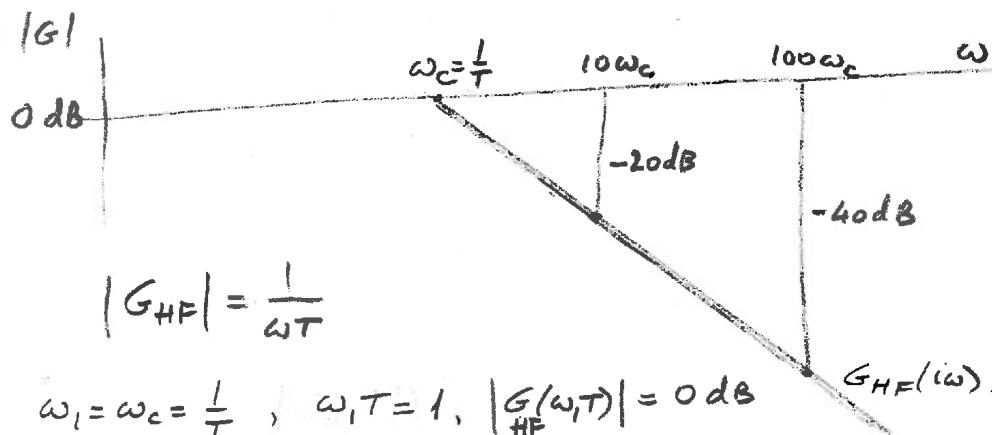
### Intersection of $G_{LF}$ & $G_{HF}$

$$1 = \frac{1}{\omega_c T} \quad \omega_c = \frac{1}{T} \quad \begin{array}{l} \text{cutoff frequency} \\ \text{decay starts at } \omega_c \end{array}$$

311/523

<sup>2</sup>  
<sup>3</sup>  
Slope of  $G_{HF}$

$$\left| \frac{G_{HF}}{dB} \right| = -20 \log_{10}(\omega T)$$



$$\left| G_{HF} \right| = \frac{1}{\omega T}$$

$$\omega_1 = \omega_c = \frac{1}{T}, \quad \omega_1 T = 1, \quad \left| G_{HF}(\omega_1 T) \right| = 0 \text{ dB}$$

$$\omega_2 = 10\omega_c, \quad \omega_2 T = 10, \quad \left| G_{HF}(\omega_2 T) \right| = \frac{1}{10} = -20 \text{ dB}$$

$$\omega_3 = 100\omega_c, \quad \omega_3 T = 100, \quad \left| G_{HF}(\omega_3 T) \right| = \frac{1}{100} = -40 \text{ dB}$$

$|G_{HF}|$  drops  $-20 \text{ dB/decade}$

For octave, take  $\omega_c, 2\omega_c, 4\omega_c$

$$\omega_1 \quad \omega_2 \quad \omega_3$$

$$1 \quad \frac{1}{2} \quad \frac{1}{4}$$

$$0 \text{ dB} \quad -6 \text{ dB} \quad -12 \text{ dB}$$

$|G_{HF}|$  drops  $-6 \text{ dB/octave}$

3

Exact amplitude and phase at  $\omega_c$ 

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

$$G(i\omega_c) = \frac{1}{i\omega_c T + 1} = \frac{1}{1+i} = \frac{1}{\sqrt{2}} e^{-i\frac{\pi}{4}}$$

$$|G(i\omega_c)| = \frac{1}{\sqrt{2}} = -3 \text{ dB}$$

mag  $2\text{dB}(1/\sqrt{2})$   
 ans  $3.0103$

$$\angle G(i\omega_c) = -\frac{\pi}{4} = -45^\circ$$

Error of using asymptotesMaximum error occurs at  $\omega_c$ 

Appox. value = 0 dB

Exact value = -3 dB

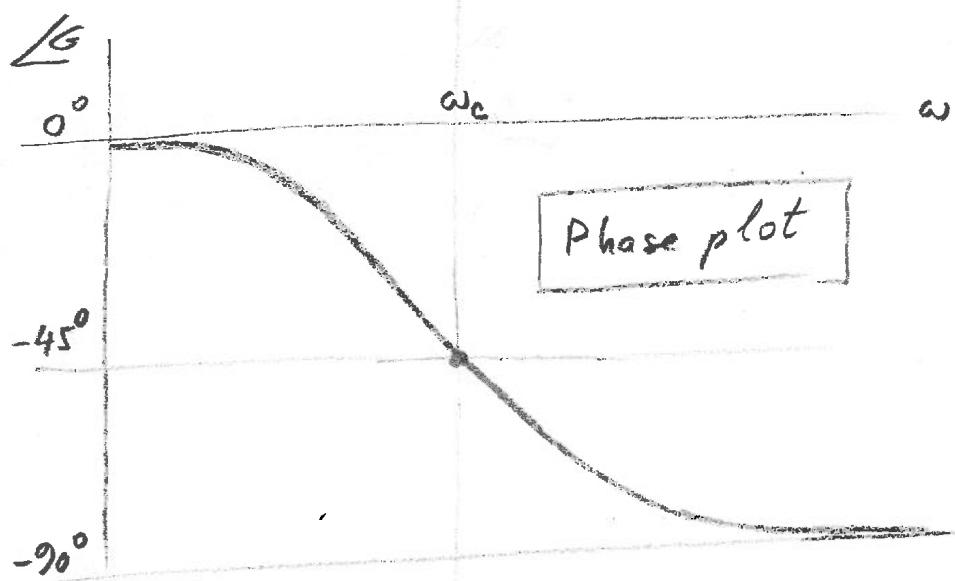
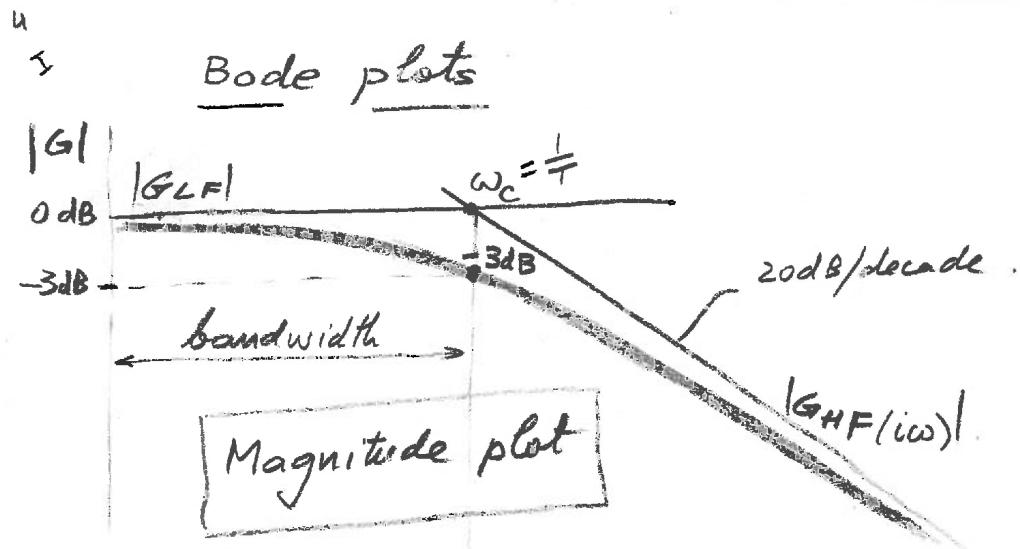
Error = 3 dB

Bandwidth  $\omega_B$ 

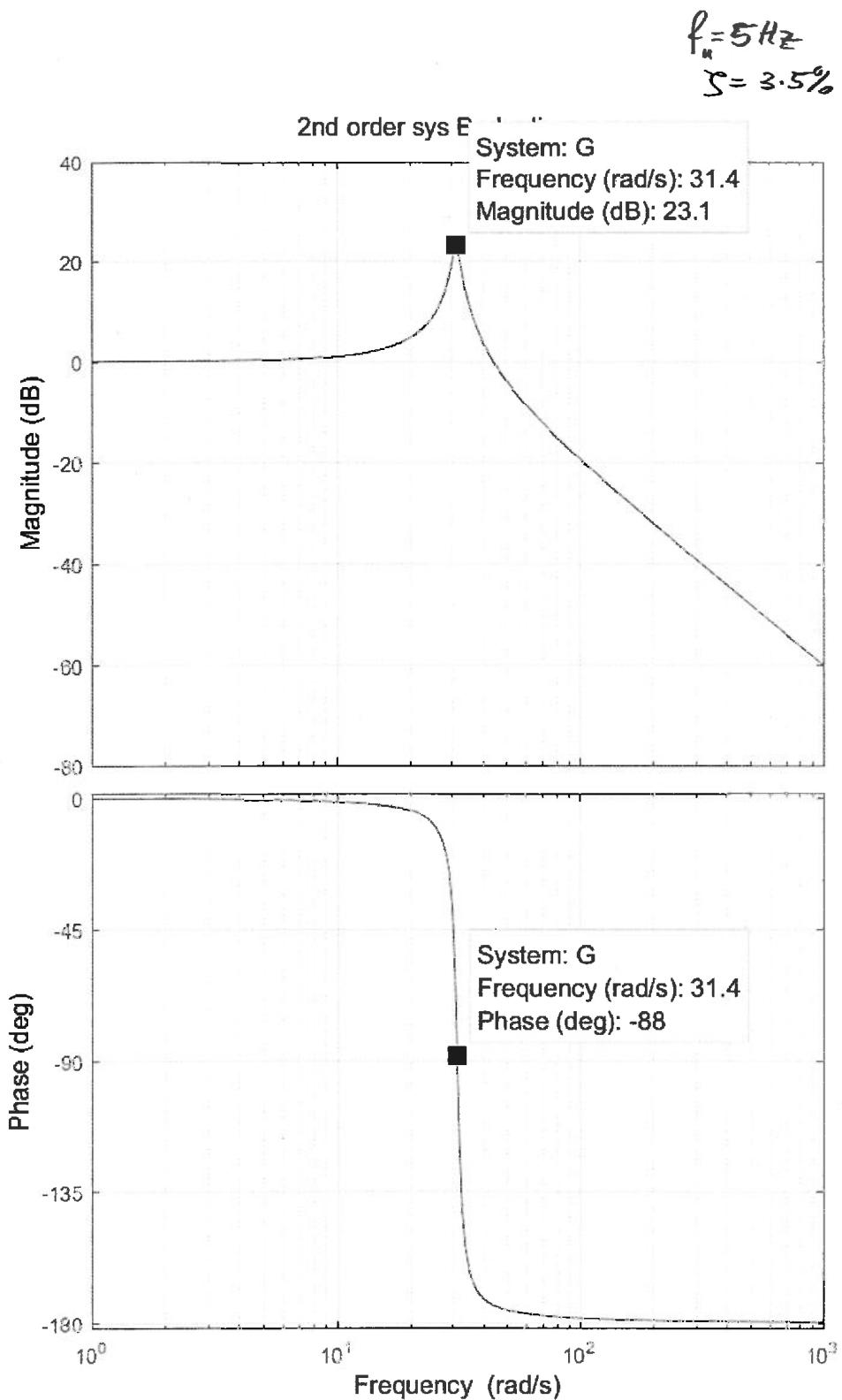
Bandwidth is defined as the frequency below which signal does not decrease more than 3 dB

For 1st order sys,  $\omega_B = \omega_c = \frac{1}{T}$ 

313/523



2<sup>nd</sup> order system FRF



## 2<sup>nd</sup> order syst. FRF

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

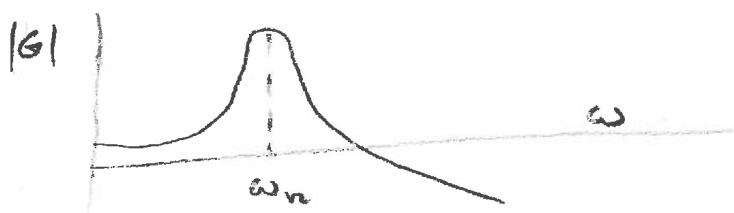
$$\begin{aligned} G(i\omega) &= \frac{\omega_n^2}{-\omega^2 + 2\zeta\omega_n\omega + \omega_n^2} \\ &= \frac{\omega_n^2}{(-\omega^2 + \omega_n^2) + i2\zeta\omega_n\omega} \end{aligned}$$

Phase resonance:  $\omega = \omega_n$  (natural freq.)

$$G(i\omega_n) = \frac{\omega_n^2}{(-\omega_n^2 + \omega_n^2) + i2\zeta\omega_n\omega_n} = \frac{1}{i2\zeta}$$

$$|G(i\omega_n)| = \frac{1}{2\zeta}$$

$$\angle G(i\omega_n) = \angle \frac{1}{i2\zeta} = -90^\circ$$



317/523

$\frac{1}{\omega}$ LF asymptote

$$G(i\omega) \xrightarrow{\omega \ll \omega_n} G_{LF}(i\omega)$$

$$G(i\omega) = \frac{\omega_n^2}{(-\omega^2 + \omega_n^2) + i2\zeta\omega\omega_n} \xrightarrow{\omega \ll \omega_n} \frac{\omega_n^2}{\omega^2} = 1$$

$$G_{LF}(i\omega) = 1 \quad |G_{LF}|_{dB} = 0 \text{ dB}$$

$$\angle G_{LF} = 0^\circ$$

HF asymptote

$$G(i\omega) \xrightarrow{\omega \gg \omega_n} G_{HF}(i\omega)$$

$$G(i\omega) = \frac{\omega_n^2}{(-\omega^2 + \omega_n^2) + i2\zeta\omega\omega_n} \xrightarrow{\omega \gg \omega_n} -\frac{\omega_n^2}{\omega^2}$$

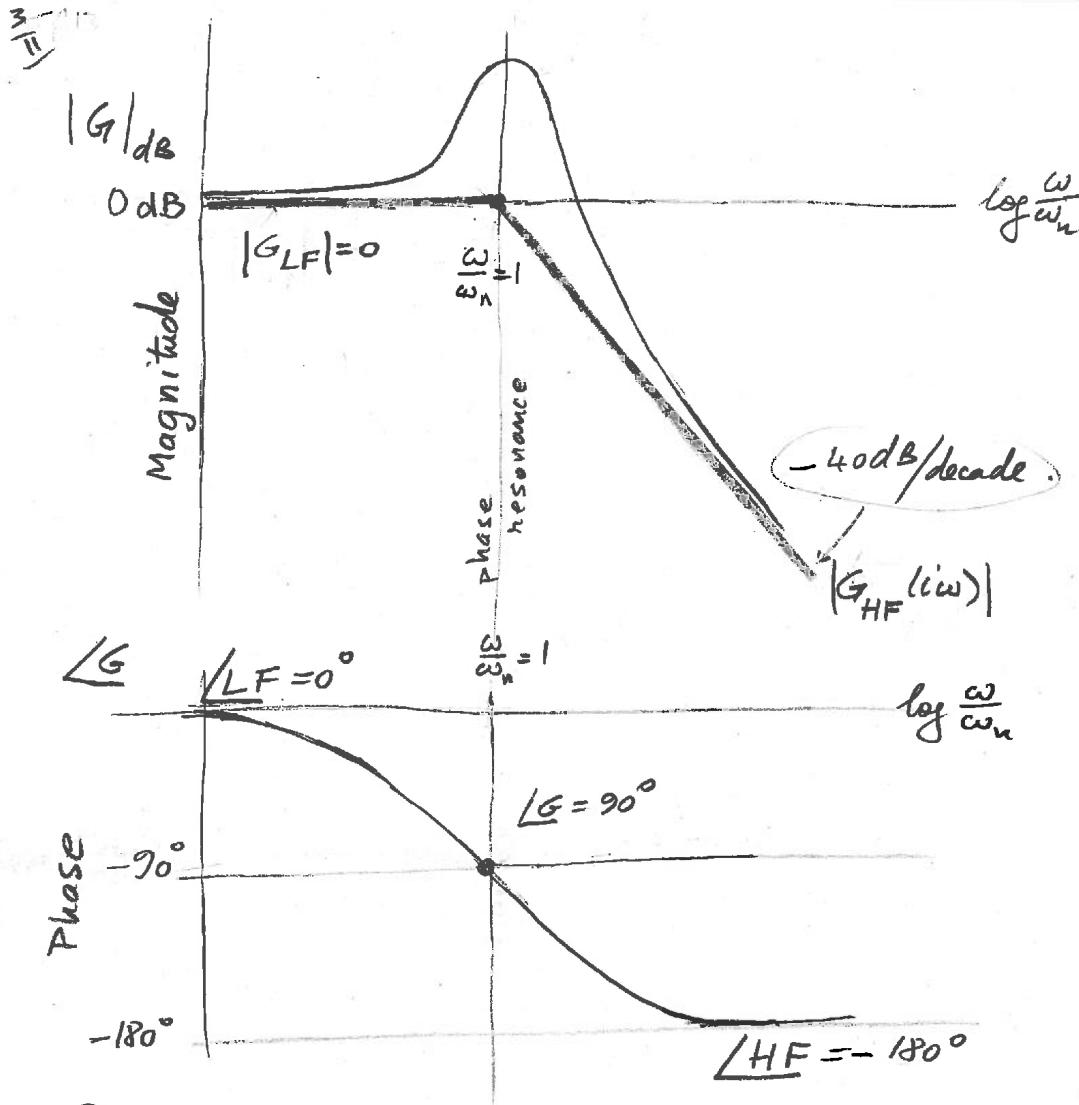
$$G_{HF}(i\omega) = -\frac{\omega_n^2}{\omega^2} = \frac{\omega_n^2}{\omega^2} e^{-i\pi}$$

$$|G_{HF}(i\omega)| = \frac{\omega_n^2}{\omega^2} = \left(\frac{\omega_n}{\omega}\right)^2 = 1/\left(\frac{\omega}{\omega_n}\right)^2$$

$$|G_{HF}(i\omega)|_{dB} = -40 \log_{10} \left(\frac{\omega}{\omega_n}\right)$$

$$\omega_2 = 10\omega_1 \\ \omega_2/\omega_1 = 10; \log_{10} \omega_2/\omega_1 = 1 \quad "40 \text{ dB/decade}"$$

$$\angle G_{HF}(i\omega) = -\pi = -180^\circ$$

Phase

$$\angle G = -\tan^{-1} \left( \frac{25 \omega \omega_n}{-\omega^2 + \omega_n^2} \right) = -\tan^{-1} \frac{25 \frac{\omega}{\omega_n}}{1 - \frac{\omega^2}{\omega_n^2}}$$

$$\frac{\omega}{\omega_n} \ll 1, \quad \angle G = 0^\circ \quad (\text{LF})$$

$$\frac{\omega}{\omega_n} = 1, \quad \angle G = -90^\circ \quad \text{'phase resonance'}. \\ (\tan(\infty) = 90^\circ)$$

$$\frac{\omega}{\omega_n} \gg 1, \quad \angle G = -180^\circ \quad (\text{HF})$$

319/523

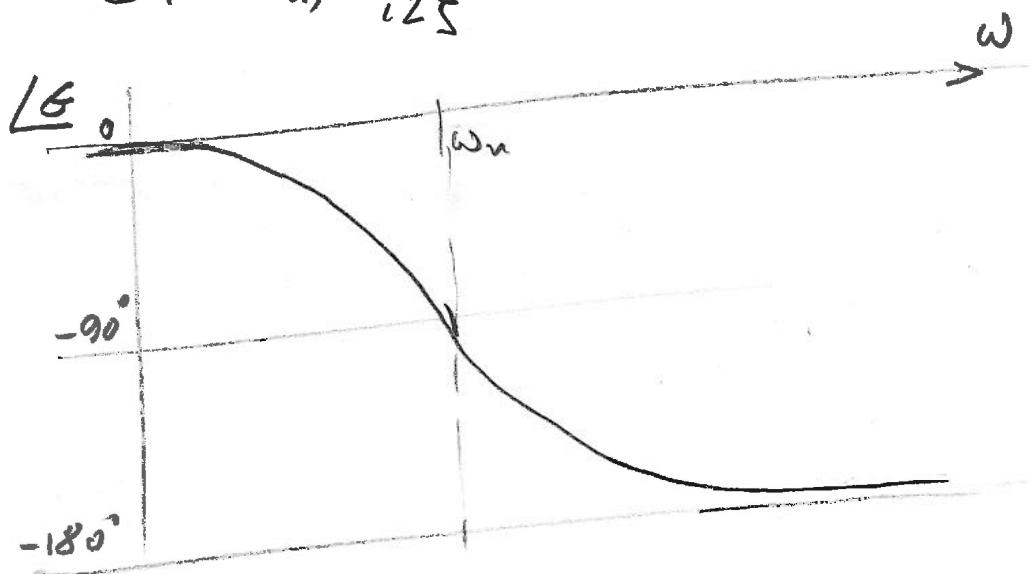
IV Phase diagram.

$$G(i\omega) = \frac{\omega_n^2}{(-\omega^2 + \omega_n^2) + i2\zeta\omega\omega_n}$$

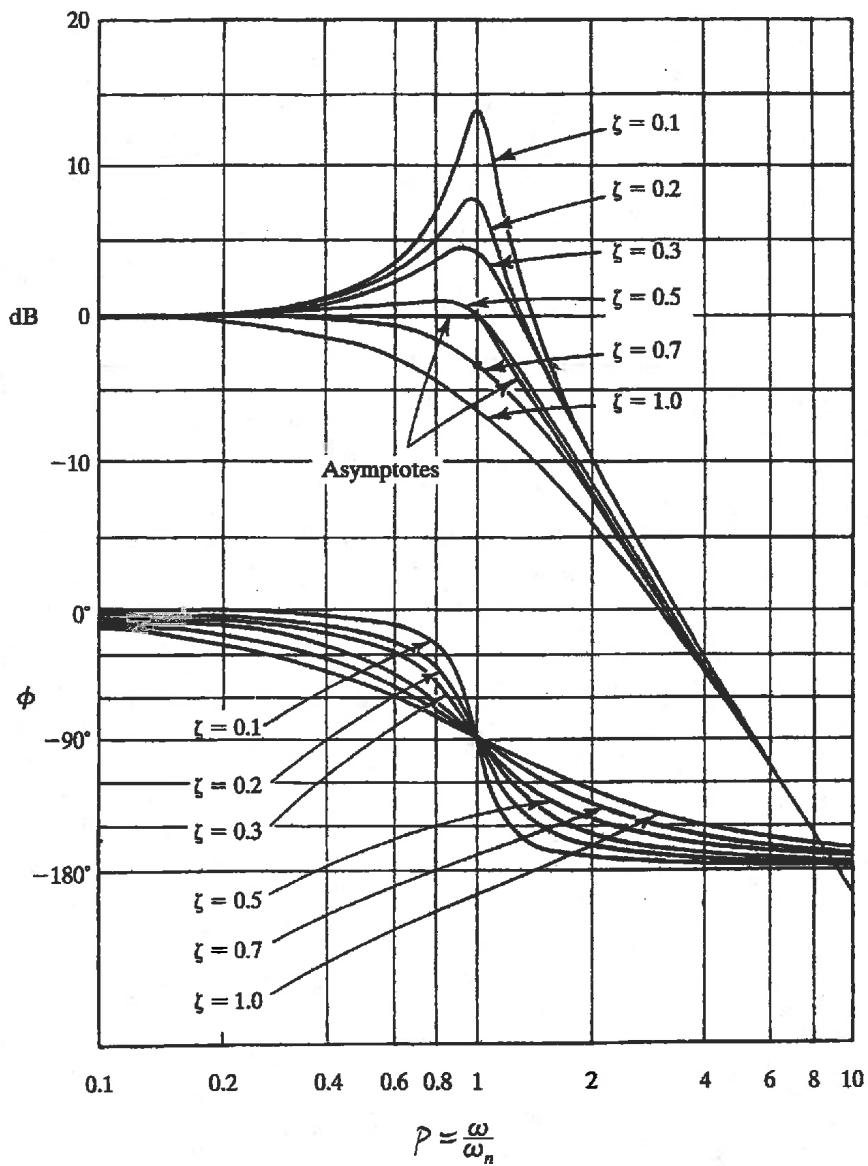
$$G_{LF}(\omega) = 1 \quad \angle 1 = 0^\circ$$

$$G_{HF}(\omega) = -\frac{\omega_n^2}{\omega^2} \quad \angle -1 = -180^\circ$$

$$G(\omega = \omega_n) = \frac{1}{i2\zeta} \quad \angle \frac{1}{i} = -90^\circ$$



### Bode Diagram Representation of the Frequency Response



Log-magnitude curves together with the asymptotes and phase-angle curves of the quadratic sinusoidal transfer function

MATLAB FD2

### 8.3 Performance Indicators in Frequency Domains

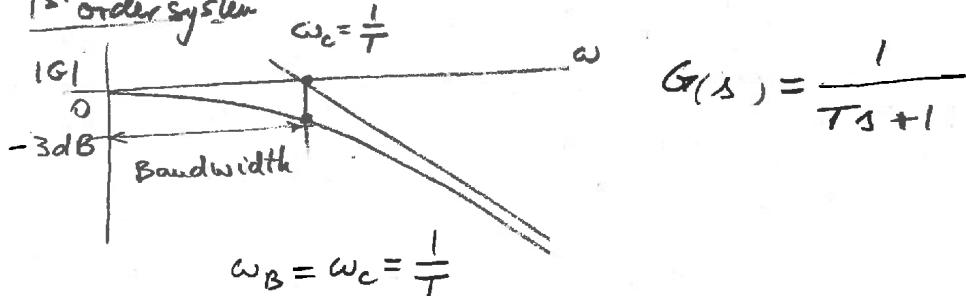
PIF

Generic Performance  
Indicators in Freq. Domain

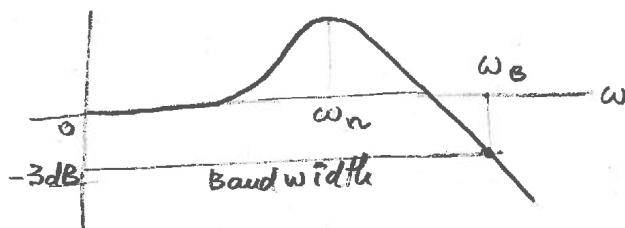
PF

Generic Performance Indicators in Freq. DomainBandwidth and cutoff frequency,  $\omega_B$ 

$$|G(j\omega_B)| = G(0) - 3 \text{ dB} \quad \text{has declined 3dB below LF value}$$

1<sup>st</sup> order system2<sup>nd</sup> order system

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

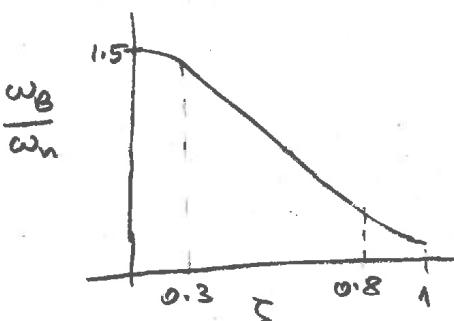


$$|G(j\omega)|_{dB} = -20 \log_{10} \sqrt{(1 - \frac{\omega^2}{\omega_n^2})^2 + (2\zeta \frac{\omega}{\omega_n})^2} = -3 \text{ dB}$$

$\omega_B$  is determined graphically or solved numerically

$$\frac{\omega_B}{\omega_n} \approx -1.19 \zeta + 1.85$$

$$0.3 < \zeta < 0.8$$



$\omega_B \uparrow \text{as } \zeta \downarrow$

<sup>3</sup><sub>PIF</sub> Ex. 11.2

Given: two systems:

$$G_1 = \frac{1}{s+1} ; G_2 = \frac{1}{3s+1}$$

1st order system performance

- Find: (a) bandwidth  
 (b) frequency response  
 (c) step response  
 (d) ramp response  
 (e) discuss results

Solution:  $G_1(i\omega) = \frac{1}{i\omega+1} = \frac{1}{i\omega T_1 + 1} \rightarrow T_1 = 1$

$$G_2(i\omega) = \frac{1}{3i\omega+1} = \frac{1}{i\omega T_2 + 1} \rightarrow T_2 = 3$$

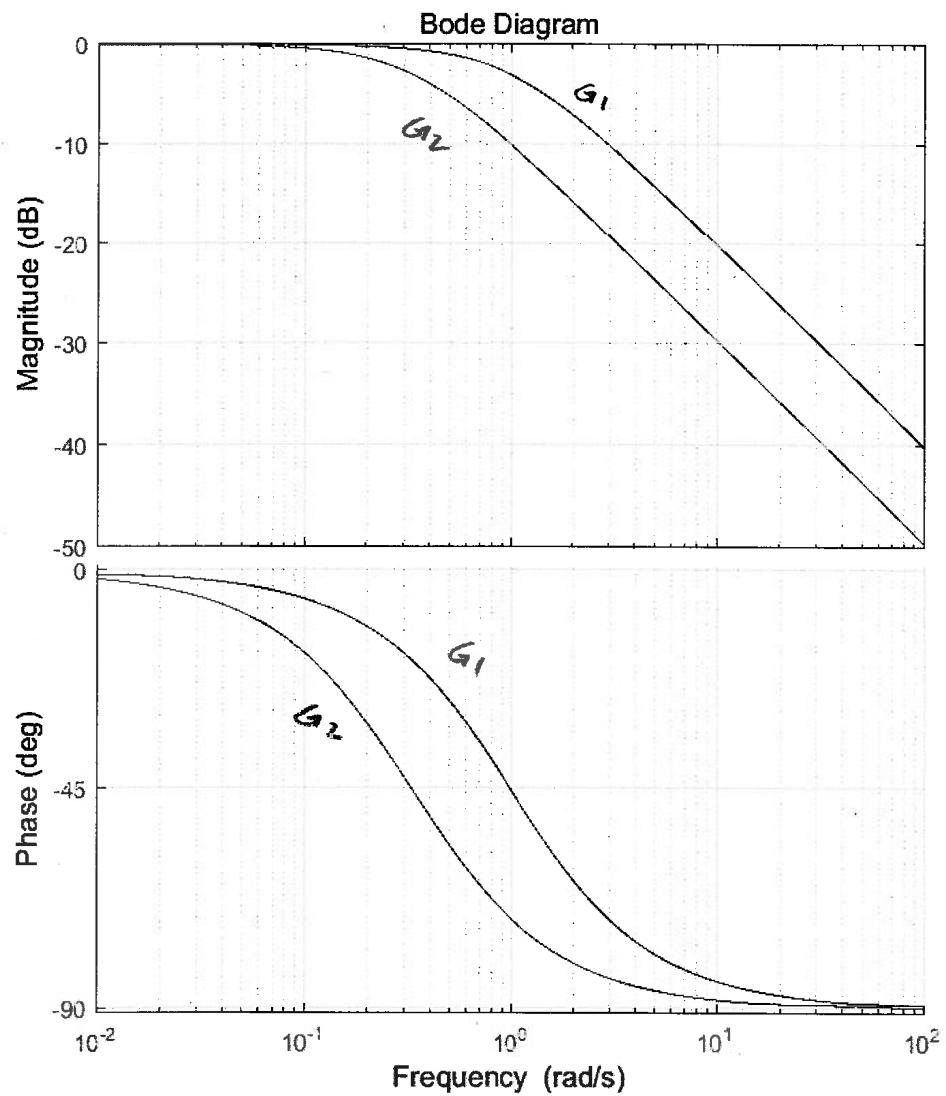
(a)  $\omega_B = \frac{1}{T}$        $\omega_1^B = \frac{1}{1} = 1 \text{ rad/sec}$        $\omega_2^B = \frac{1}{3} \text{ rad/sec}$

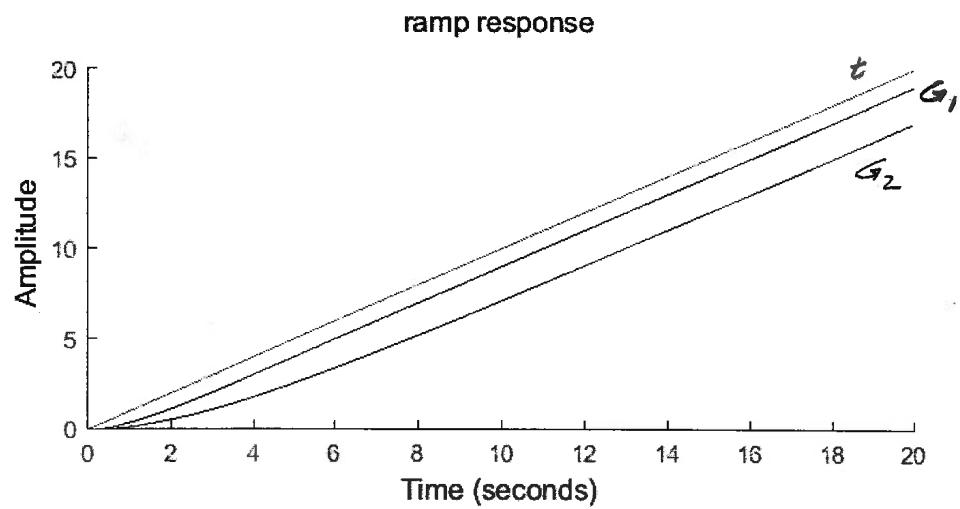
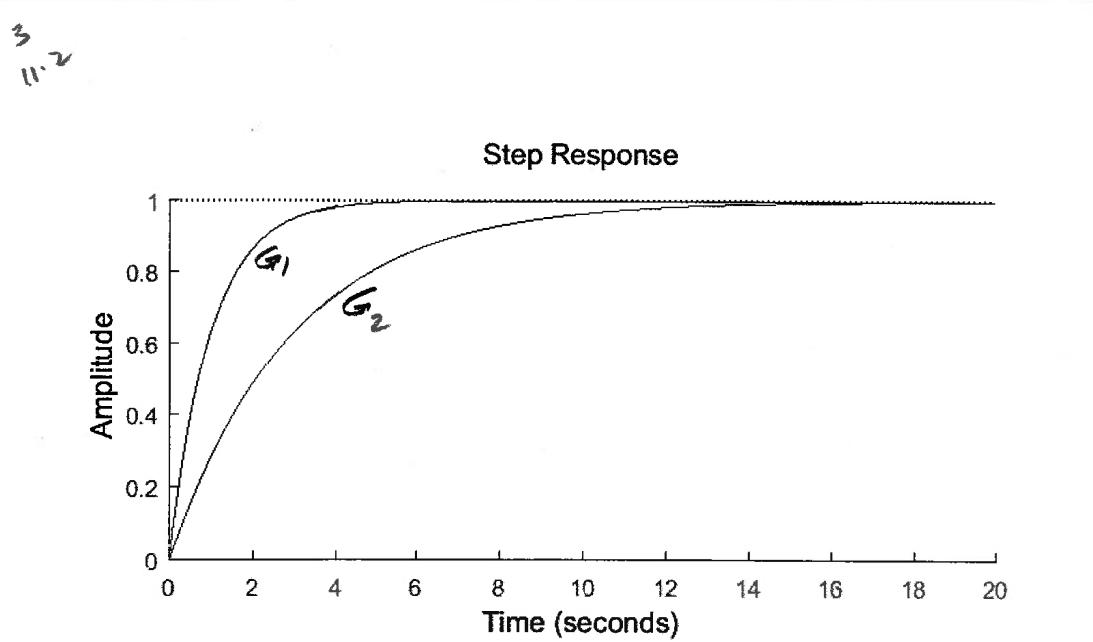
(b), (c), (d): see MATLAB Ex. 11.2

(e) System 1 has bandwidth three times larger than system 2 ( $\omega_1 = 1$  vs.  $\omega_2 = 1/3 \text{ rad/sec}$ ).

Sys. 1 has faster step response and follows the ramp input much better / smaller ramp error)

4  
PIF





QIF

C:\Mydata\1 USC...\Example11\_2\_p628 20161216.m Page 1

```
1 %{
2 EXAMiPLE 11.2
3 1st Order System analysis
4 %}
5 %% initialization
6 clc
7 clear all
8 % close all
9 s=tf('s');
10 %% system definition
11 G1=1/(s+1); G2=1/(3*s+1);
12 %% Bode plots
13 figure(1)
14 bode(G1,G2)
15 box off
16 grid on
17 %% time response
18 figure(2)
19 Tfinal=20;
20 dt=0.1; t=0:dt:Tfinal; u=1.*t;
21 subplot(2,1,1)
22 step(G1,G2,Tfinal)
23 box off
24 subplot(2,1,2)
25 lsim(G1,G2,u,t)
26 axis([0 Tfinal 0 Tfinal]); title( 'ramp response')
27 box off
28
29
30
31
32
```

Ex: 2<sup>nd</sup> order sys. Performance

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

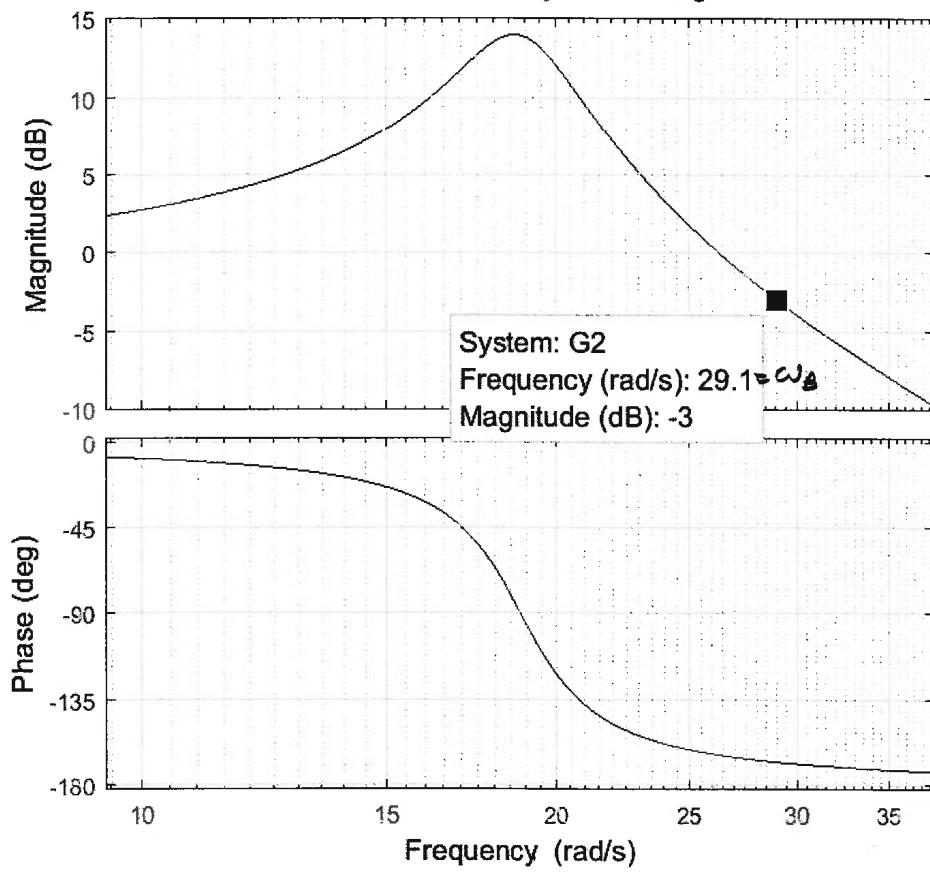
$$\omega_n = 2\pi f_n$$

$$f_n = 3 \text{ Hz}$$

$$\zeta = 10\%$$

$$\omega_B = 29.1 \text{ rad/s}$$

zoom 2nd order sys Bode diag.



7  
PIF

Specific Performance Indicators  
in Freq. Domain for  
2<sup>nd</sup> Order Systems

8  
PIF

Resonance: peak response  
max. response

$$(G(i\omega))^2 = \frac{1}{(1 - \frac{\omega^2}{\omega_n^2})^2 + (2\zeta \frac{\omega}{\omega_n})^2}$$

$$\frac{d}{d\omega} |G(i\omega)|^2 = 0 \text{ (for peak)}$$



$$\frac{d}{d\omega} \left[ (1 - \frac{\omega^2}{\omega_n^2})^2 + (2\zeta \frac{\omega}{\omega_n})^2 \right] = 0.$$

use auxiliary variable  $p = \frac{\omega}{\omega_n}$  and write

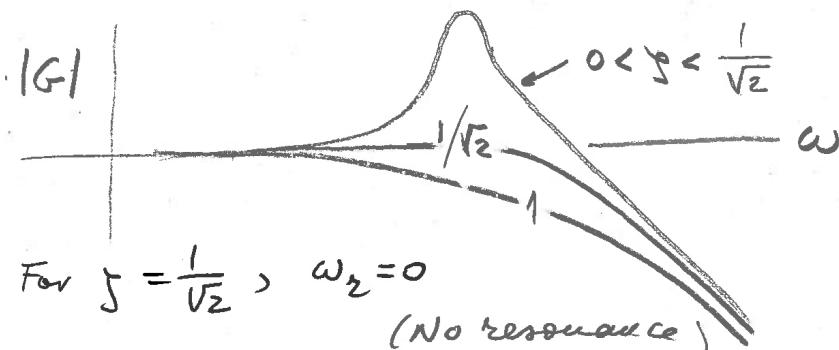
$$\frac{d}{dp} \left[ (1-p^2)^2 + (2\zeta p)^2 \right] = 0.$$

$$2(-2p)(1-p^2) + 2(2\zeta)(2\zeta p) = 0.$$

$$(1-p^2)p' = 2\zeta^2 p$$

$$p^2 = 1 - 2\zeta^2 \rightarrow p_2 = \sqrt{1 - 2\zeta^2}$$

$$\omega_2 = \omega_n \sqrt{1 - 2\zeta^2}$$



Peak exist only for  $0 < \zeta < \frac{1}{\sqrt{2}}$

Amplitude at resonance

$$M_2^2 = |G(i\omega_2)|^2 = \frac{1}{[1 - (1 - 2\zeta^2)]^2 + (2\zeta)^2(1 - 2\zeta^2)}$$

$$= \frac{1}{(1 - 1 + 2\zeta^2)^2 + 4\zeta^2(1 - 2\zeta^2)}$$

$$= \frac{1}{4\zeta^4 + 4\zeta^2 - 8\zeta^4} = \frac{1}{4\zeta^2(1 - \zeta^2)}$$

$$M_2 = \frac{1}{2\zeta\sqrt{1 - \zeta^2}}$$

$\zeta = \frac{1}{\sqrt{2}}$ 

$$\left| M_2 = \frac{1}{2 \frac{1}{\sqrt{2}} \sqrt{1 - \frac{1}{2}}} = 1 \right.$$

Phase at resonance,  $\varphi_2$ 

$$G(i\omega) = \frac{\omega_n^2}{(i\omega)^2 + 2i\zeta\omega_n\omega + \omega_n^2} = \frac{1}{(1 - \zeta^2) + 2i\zeta\omega}$$

$$\zeta = \sqrt{1 - 2\zeta^2}, \quad 1 - \zeta^2 = 1 - (1 - 2\zeta^2) = 2\zeta^2$$

$$G(i\omega_2) = \frac{1}{2\zeta^2 + 2i\zeta\sqrt{1 - 2\zeta^2}}$$

$$\varphi_2 = \angle G(i\omega_2) = -\tan^{-1} \frac{\sqrt{1 - 2\zeta^2}}{\zeta}$$

$$\varphi_2 = -90^\circ + \left( \sin^{-1} \frac{\zeta}{\sqrt{1 - \zeta^2}} \right) = -90^\circ + \varphi_2$$

Proof

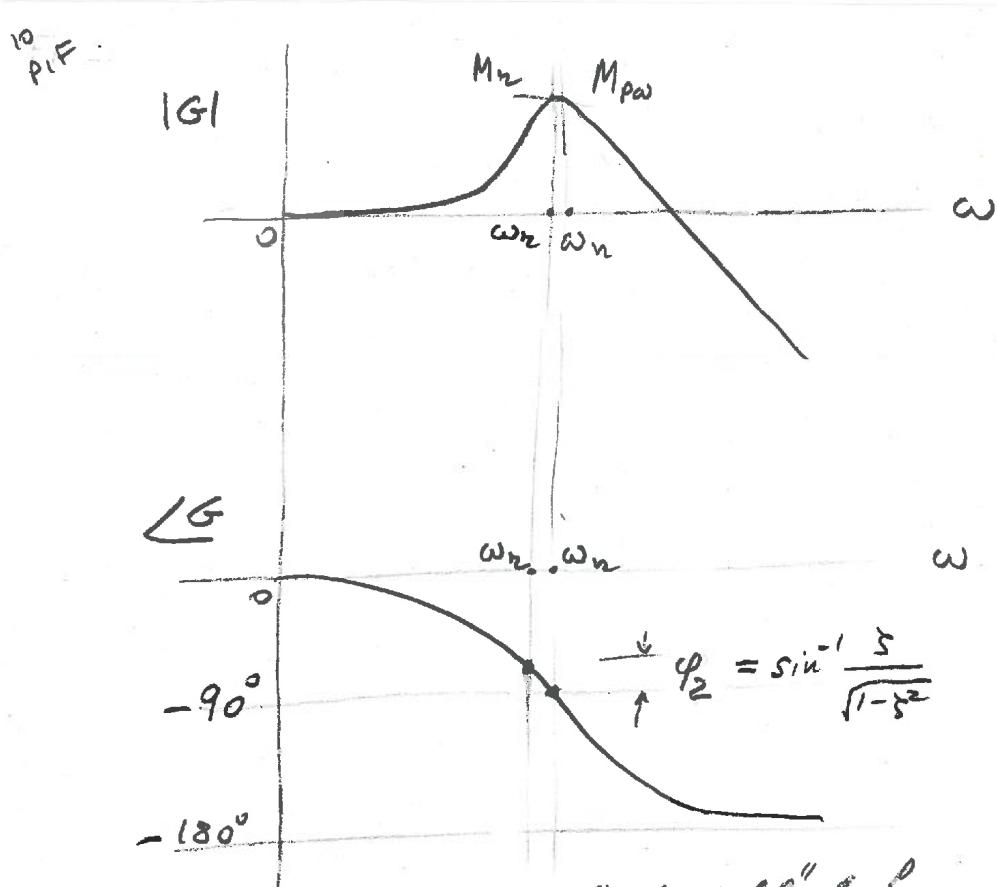
$$\tan(-90^\circ + \varphi_2) = -\cotan \varphi_2$$



$$\sin \varphi_2 = \frac{\zeta}{\sqrt{1 - \zeta^2}}$$

$$\cos^2 \varphi_2 = 1 - \sin^2 \varphi_2 = \frac{1 - 2\zeta^2}{1 - \zeta^2}$$

$$\cotan \varphi_2 = \frac{\cos \varphi_2}{\sin \varphi_2} = \frac{\sqrt{1 - 2\zeta^2}/\sqrt{1 - \zeta^2}}{\zeta/\sqrt{1 - \zeta^2}} = -\tan \varphi_2$$



Resonance happens "slightly" before  $\omega_n$   
 Two definitions of "resonance":

(1)  $90^\circ$  phase  $\Rightarrow \omega_n$ ,  $M_{\text{ph}} = \frac{1}{25}$   
 (phase resonance)

(2) peak value  $\Rightarrow \omega_r = \omega_n \sqrt{1-25^2}$

$$M_r = \frac{1}{25\sqrt{1-5^2}}$$

"RESONANCE"

$\zeta$	0.01	0.1	0.4	$1/\sqrt{2}$	0.8	0.99	1
$M_{pw} = \frac{1}{2\zeta}$	50	5	1.25	0.7	0.625	0.505	0.5
$M_n = \frac{1}{2\zeta\sqrt{1-\zeta^2}}$	50	5.025	1.36	1	—	—	—
$1+M_p = 1 + e^{-\frac{\zeta\pi}{\sqrt{1-\zeta^2}}}$	1.97	1.73	1.25	1.04	1.02	1.00	1
$\frac{\omega_r}{\omega_n} = \sqrt{1-2\zeta^2}$	1	0.99	0.825	0	—	—	—
$\frac{\omega_d}{\omega_n} = \sqrt{1-\zeta^2}$	1	0.995	0.917	$1/\sqrt{2}$	0.6	0.141	0
$\frac{t_3}{T_n} = \frac{2}{\pi\zeta}$	63	6.37	1.59	0.9	0.8	0.64	0.64
$\varphi = -\tan^{-1} \frac{\sqrt{1-2\zeta^2}}{\zeta}$	$-89^\circ$	$-84^\circ$	$-64^\circ$	$0^\circ$	—	—	—

$$\left. \begin{aligned} t_3 &= \frac{4}{\zeta \omega_n} \\ T_n &= \frac{1}{f_n} = \frac{2\pi}{\omega_n} \quad (\text{osc. period}) \end{aligned} \right\} \frac{t_3}{T_n} = \frac{4}{\zeta} \frac{1}{2\pi} = \frac{2}{\pi\zeta}$$

(2)  
PIFSpecific PIs vs  $\zeta$ 2<sup>nd</sup> order system

performance indicators comparison, 2nd order sys						
$z =$						
0.0100	0.1000	0.4000	0.7071	0.8000	0.9900	1.0000
$M_{pw} =$						
50.0000	5.0000	1.2500	0.7071	0.6250	0.5051	0.5000
$M_r =$						
50.0025	5.0252	1.3639	1.0000	0	0	NaN
$1+M_p =$						
1.9691	1.7292	1.2538	1.0432	1.0152	1.0000	1.0000
$w_r/w_n =$						
0.9999	0.9899	0.8246	0.0000	0	0	0
$w_d/w_n =$						
0.9999	0.9950	0.9165	0.7071	0.6000	0.1411	0
$t_s/t_n =$						
63.6620	6.3662	1.5915	0.9003	0.7958	0.6431	0.6366
$\phi_r =$						
-89.4270	-84.2318	-64.1233	0.0000	0	0	0

$$M_{pw} = \frac{1}{2\zeta}$$

magnitude at phase res.

$$M_r = \frac{1}{2\zeta\sqrt{1-\zeta^2}}$$

resonance peak

$$1+M_p = 1 + e^{-\frac{5\pi}{\sqrt{1-\zeta^2}}} \text{ step response peak}$$

$$\frac{\omega_r}{\omega_n} = \sqrt{1-2\zeta^2} \text{ resonance freq. ratio}$$

$$\frac{\omega_d}{\omega_n} = \sqrt{1-\zeta^2} \text{ damped freq / natural freq}$$

$$t_s/\tau_n = \frac{2}{\pi\zeta} \text{ settling time / osc. period}$$

$$\phi_r = -\tan^{-1} \frac{\sqrt{1-2\zeta^2}}{\zeta} \text{ phase at resonance}$$

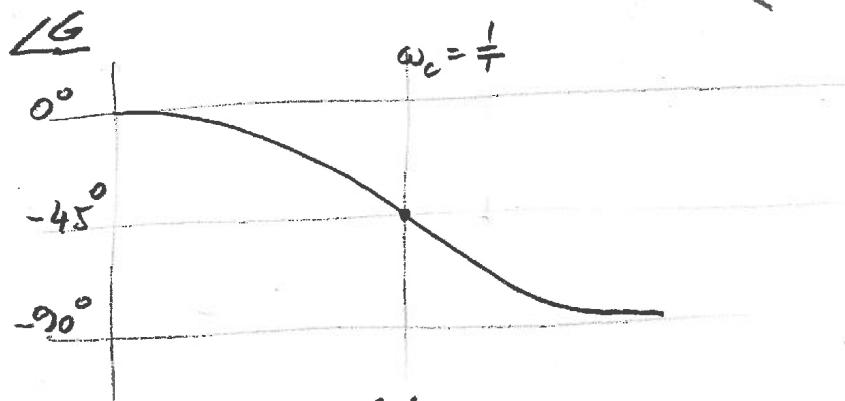
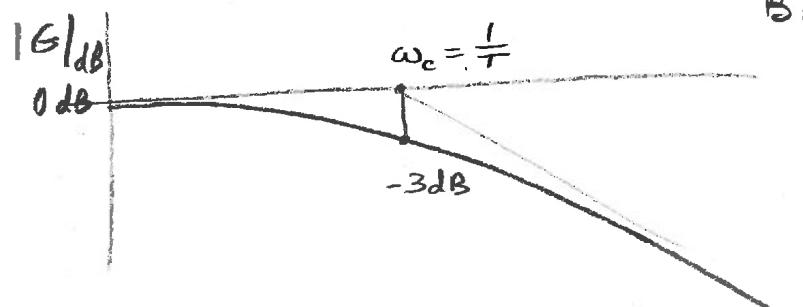
## 8.4 System Identification in Frequency Domains

IDF

1<sup>st</sup> order sys ID in Freq. Domain

$$G(i\omega) = \frac{1}{i\omega T + 1}$$

Bode plots



Given : Bode plots

Find :  $T$ 

Sol<sup>n</sup> : Read:  $-3dB$  point on  $|G|_{dB}$  plot  
 $45^\circ$  point on  $\angle G$  plot

Estimate  $\omega_c$ 

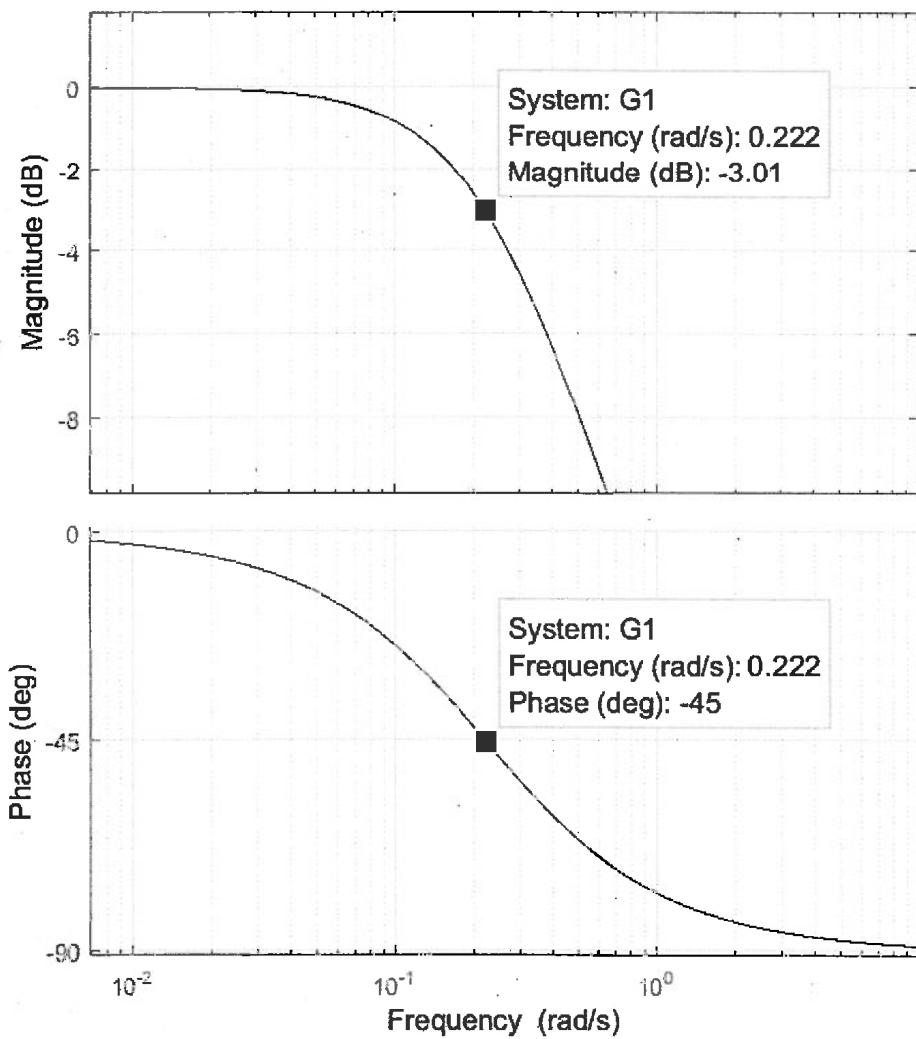
$$\text{Calculate } T = \frac{1}{\omega_c}$$

VDF

Ex: ID 1<sup>st</sup> order sys.

$$G(s) = \frac{1}{4.5s+1}$$

Bode Diagram

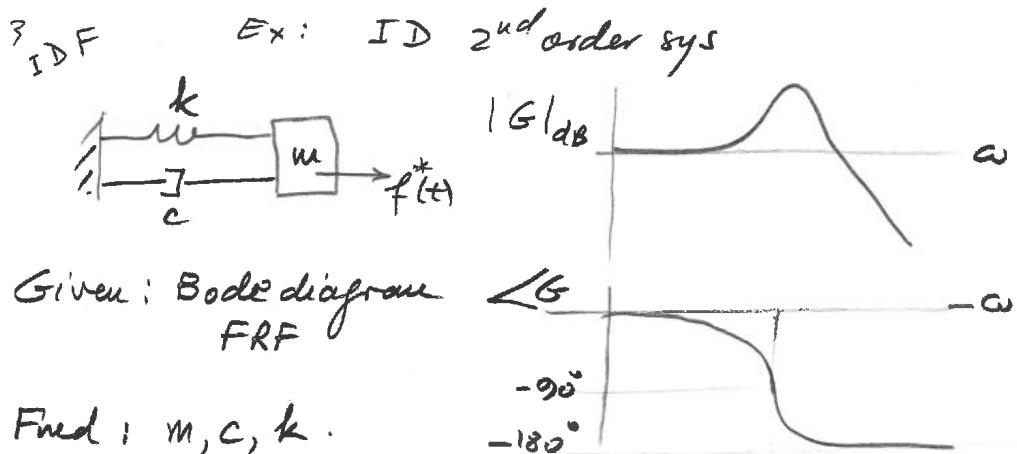


$$\omega_c = 0.222$$

$$T = \frac{1}{\omega_c} = 4.5045$$

Error: -0.1%

In practice, error may be larger due  
to noise.



Solution: recall FBD, EOM

$$m\ddot{x} + c\dot{x} + kx = f^*(t)$$

$$\mathcal{L}T \quad (m\Delta^2 + c\Delta + k)X(\Delta) = F^*(\Delta)$$

$$G(s) = \frac{X(s)}{F^*(s)} = \frac{1}{m\Delta^2 + c\Delta + k}$$

$$G(i\omega) = \frac{1}{-m\omega^2 + i\omega c + k} ; G(0) = \frac{1}{k}$$

$$\omega_n^2 = k/m \quad G(i\omega_n) = \frac{1}{i\omega_n c} ; |G(i\omega_n)| = \frac{1}{\omega_n c}$$

$$\text{Numerical example: } m=2\text{kg}, \quad c=4 \frac{\text{N}}{\text{m/s}}, \quad k=20 \frac{\text{N}}{\text{m}}$$

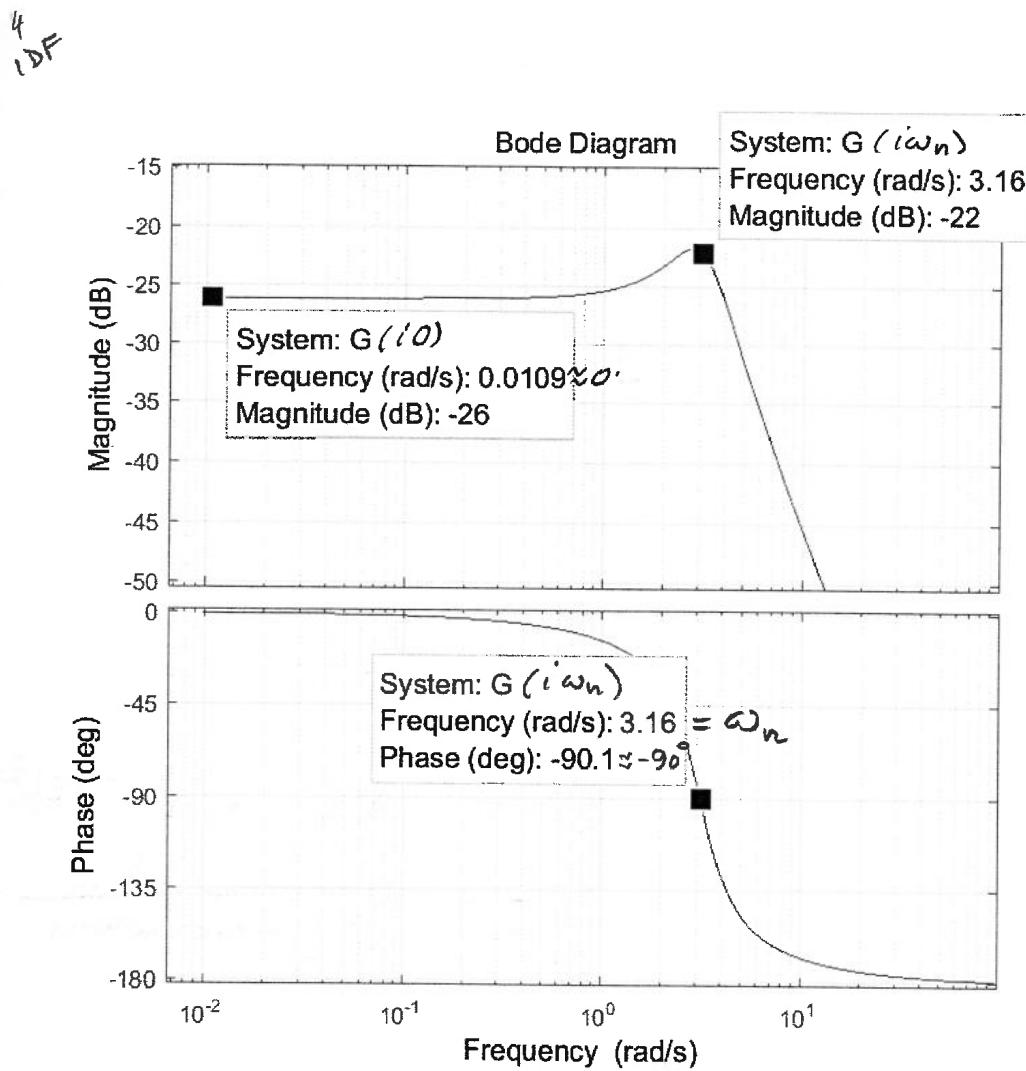
Run MATLAB sys-ID\_2ndOrder\_ExII-1

Read on Bode diagram:

$$|G(0)|_{\text{dB}} = -26 \text{dB}$$

$$\omega_n = 3.16 \text{ rad/sec}$$

$$\left| \frac{G(i\omega_n)}{\text{dB}} \right| = -22 \text{dB}$$



$$(M_{P\omega})_{dB} = |G(i\omega_n)|_{dB} - |G(i\omega)|_{dB}$$

$$= -22 dB - (-26 dB) = +4 dB$$

<sup>5</sup><sub>DF</sub> calculate  $k_1, \omega_1, c_1$

$$k_1 = \frac{1}{G(i\omega)} = 1/\text{dB}^2 \text{mag}(G(i\omega)_{\text{dB}})$$

$$= 19.9526 \text{ N/m} \approx 20 \text{ N/m} \quad \checkmark$$

$$\omega_n^2 = k/m$$

$$m_1 = \frac{k_1}{\omega_n^2} = 1.9981 \text{ kg} \approx 2 \text{ kg} \quad \checkmark$$

$$c_1 = \frac{1}{\omega_n |G(i\omega)|} = 1/\omega_n/\text{dB}^2 \text{mag}(G(i\omega_n)_{\text{dB}}) = 3.9839$$

or  $\approx 4 \frac{\text{N}}{\text{m/s}}$

$$(M_{p\omega})_{\text{dB}} = |G(i\omega_n)|_{\text{dB}} - |G(i\omega)|_{\text{dB}}$$

$$M_{p\omega} = \text{dB}^2 \text{mag}(|G(i\omega_n)|_{\text{dB}} - |G(i\omega)|_{\text{dB}})$$

$$M_{p\omega} = \frac{1}{2\zeta} \rightarrow \zeta = \frac{1}{2M_{p\omega}}$$

$$c_2 = 2\zeta \omega_n m_1 = 3.9920 \frac{\text{N}}{\text{m/s}} \approx 4 \frac{\text{N}}{\text{m/s}} \quad \checkmark$$

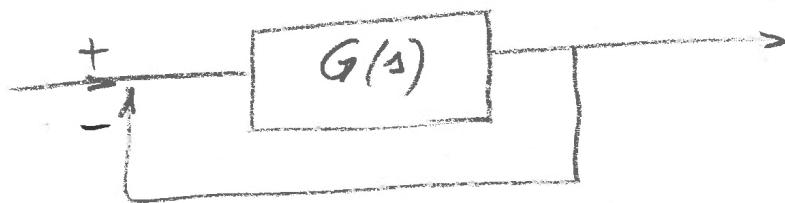
All three parameters of the system,  
 $k, m, c$  have been recovered with  
quite acceptable error.

## 8.5 Frequency Domain Analysis of Feedback System Stability

## STABILITY ANALYSIS

in FREQ. DOMAIN

Evaluate stability of CL system  
by analysing OL system in freq. domain



Two methods :

- Nyquist criterion
- Gain & phase margins

343/523

NYQUIST ANALYSIS  
OF FEEDBACK STABILITY.

Nyquist circuit

$$G(s) = \frac{1}{(s-p_1)(s-p_2)(s-p_3)}$$

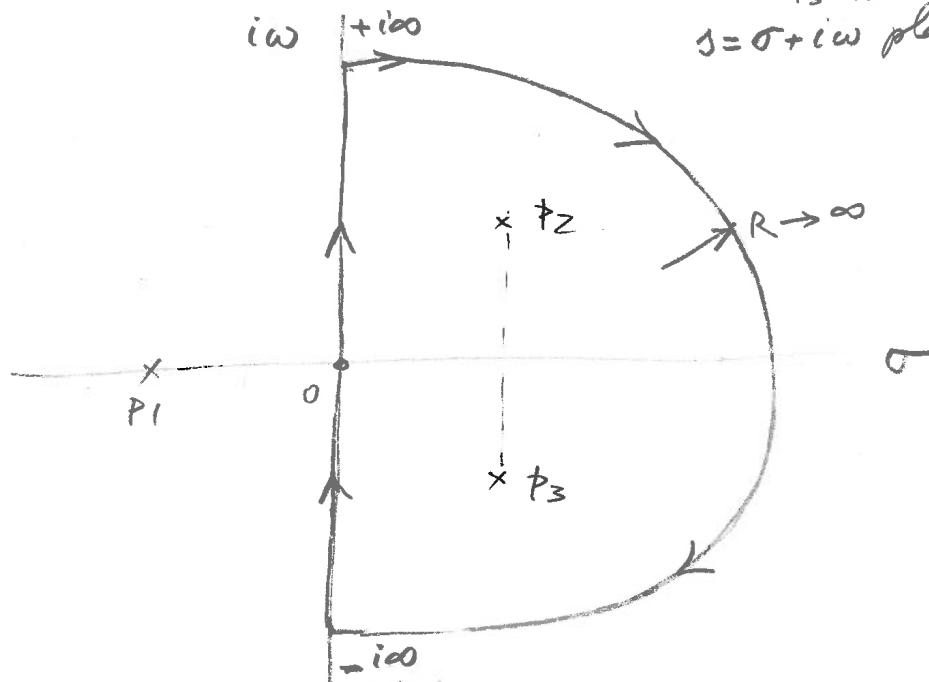
Assume

$$p_1 \in \mathbb{R}, p_1 < 0$$

$$p_2, p_3 \in \mathbb{C} \text{ in RHS}$$

$$p_3 = \bar{p}_2$$

$$s = \sigma + i\omega \text{ plane}$$

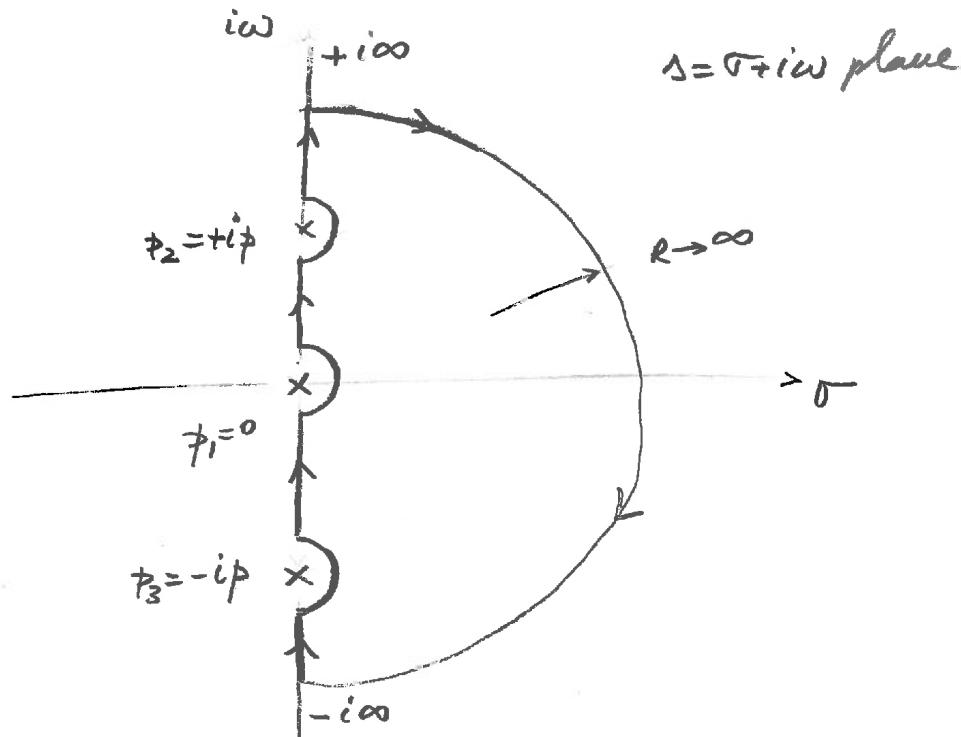


Nyquist circuit is :

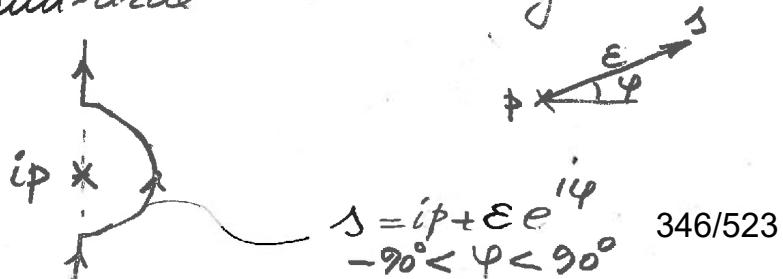
- semi-circle path with  $R \rightarrow \infty$
- in the RHS of  $s$ -plane
- along vertical axis from  $-i\infty$  to  $+i\infty$
- traveled clockwise (CW)

Nyquist circuit with poles  
on the vertical axis

$$G(s) = \frac{1}{s(s-i\omega)(s+i\omega)}$$



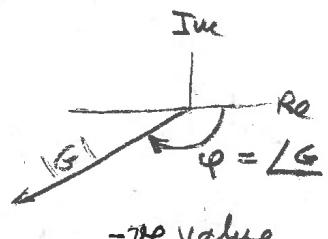
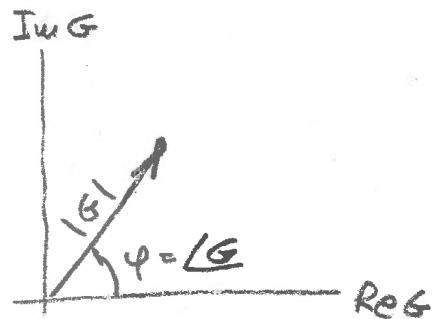
The poles on the vertical axis must be excluded. We travel around them in the RHS. To do so, we follow a small semi-circle with vanishing radius  $\epsilon$ .



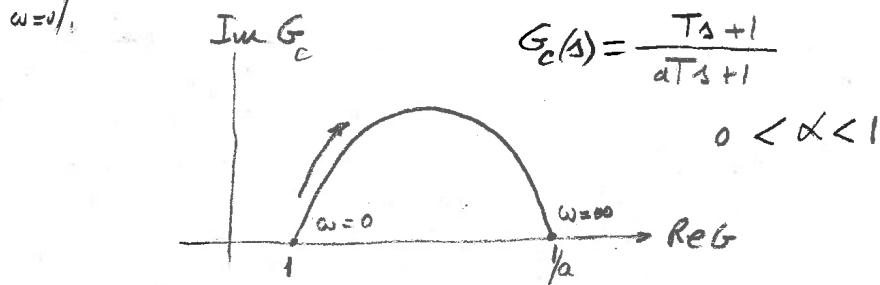
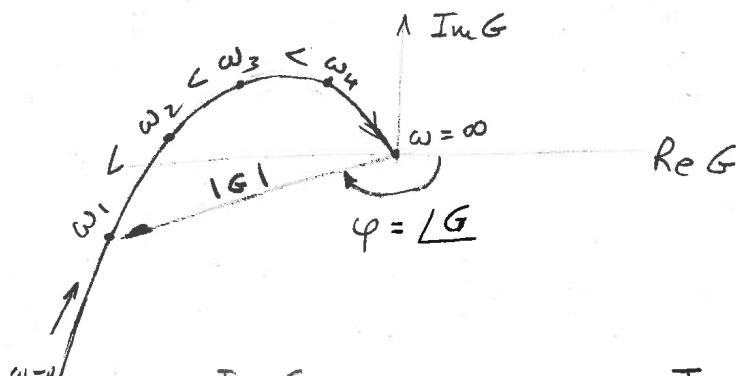
NI

## POLAR PLOTS

$$G(i\omega) = \operatorname{Re} G(i\omega) + i \operatorname{Im} G(i\omega)$$



$-\pi < \angle(G) < \pi$   
 $-180^\circ \quad 180^\circ$   
 (see MATLAB Help)



Run MATLAB examples.

NPP

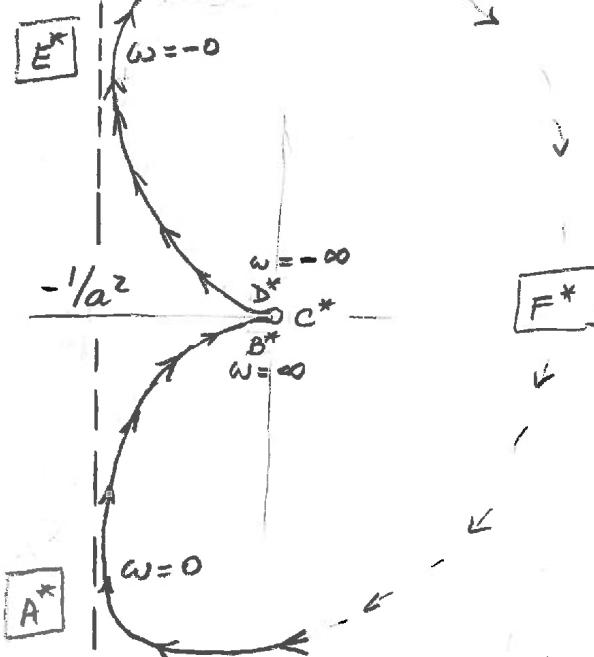
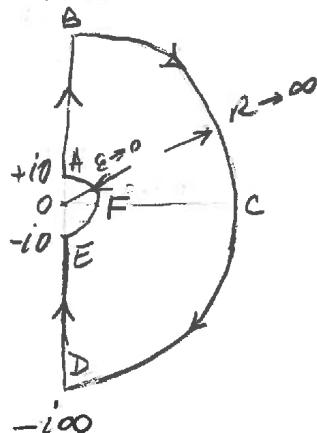
NYQUIST POLAR PLOTS

" $s$  follows  $N$ -circuit;  $G(s)$  follows  $N$ -polar plot"

Example:  $G(s) = \frac{1}{s(s+a)}$  Nyquist polar plot

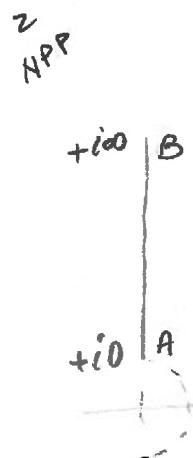
2 poles:  $p_1 = 0, p_2 = -a$

Nyquist circuit



- The variable  $s$  follows the Nyquist circuit A B C D E F A clockwise in the  $s$ -plane
- The function  $G(s)$  follows the resulting circuit in the  $G$ -plane

In this example, we distinguish 4 segments to be analyzed individually and then assembled in one continuous circuit



Segment  $(+i\epsilon, +iR)$

$$G(s) = \frac{1}{s(s+a)} = \frac{1}{as} - \frac{1}{a(s+a)}$$

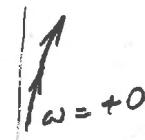
$$\boxed{A} G(i\epsilon) = \frac{1}{a i \epsilon} - \frac{1}{a(i\epsilon+a)} \quad \epsilon \ll a$$

$$G(i\epsilon) \approx -\frac{1}{a^2} - i \frac{1}{a\epsilon}$$

$$G(+i0) = \lim_{\epsilon \rightarrow 0} G(i\epsilon) = -\frac{1}{a^2} - i\infty$$

$$|G(+i0)| = \infty$$

$$\angle G(+i0) = \angle -i = -90^\circ$$



$$\boxed{B} G(iR) = \frac{1}{iR(iR+a)}$$

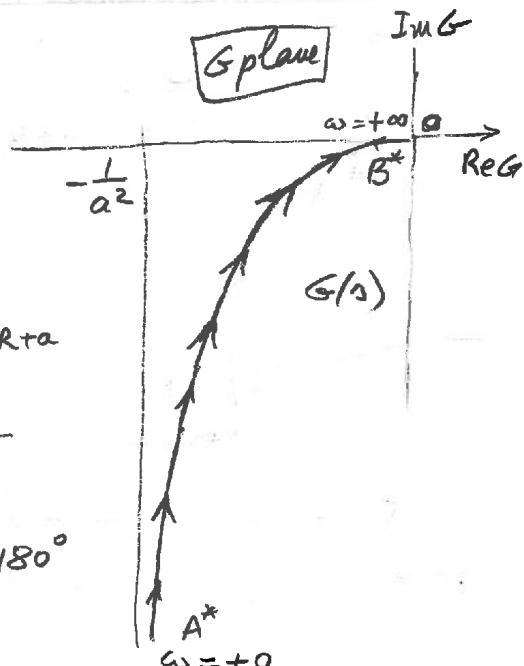
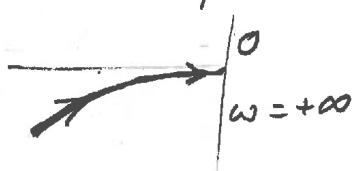
$$|G(iR)| \approx \frac{1}{R^2} \xrightarrow[R \gg a]{} 0$$

$$\angle G(iR) = -[\angle iR + \angle iR+a]$$

$iR$                      $iR+a$   
 $90^\circ$                      $\psi$

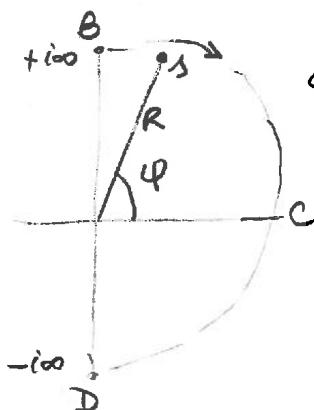
$$\varphi = \tan^{-1} \frac{a}{R}$$

$$\begin{aligned} \angle G(iR) &= -(90^\circ + 90^\circ - \psi) \\ &= -180^\circ + \psi \xrightarrow[R \rightarrow \infty]{\psi \rightarrow 0} -180^\circ \end{aligned}$$



<sup>3</sup> NPP Segment  $(iR, R, -iR)$  Big circle  
 $R \rightarrow \infty$

$$s = Re^{i\varphi}, \quad \varphi \in (90^\circ, 0, -90^\circ)$$

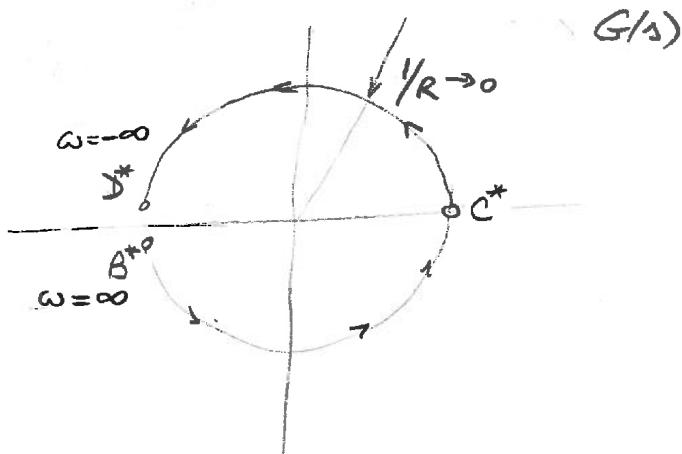


$$G(s) = \frac{1}{Re^{i\varphi}(R e^{i\varphi} + a)}$$

$$R \rightarrow \infty \quad \frac{1}{R e^{2i\varphi}} = \frac{1}{R^2} e^{-i2\varphi}$$

$$\angle G(s) = -2\varphi$$

$$\angle G(s) \in (-180^\circ, 0^\circ, 180^\circ)  
\text{B}^* \quad \text{C}^* \quad \text{D}^*$$



NPP

Segment  $(-iR, -i\varepsilon)$  $R \rightarrow \infty$      $\varepsilon \rightarrow 0$ 

s-plane

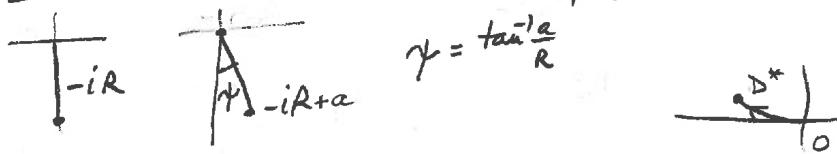
$$G(s) = \frac{1}{1/(1+a)} = \frac{1}{a^s} - \frac{1}{a(s+a)}$$

$G(s) \Big|_{s=-iR} = \frac{1}{-iR(-iR+a)}$

$$|G(s)| = \frac{1}{R^2} \xrightarrow[R \rightarrow \infty]{} 0$$

$$\angle G(s) = -(\angle -iR + \angle -iR+a)$$

$$= -[-90^\circ + (-90^\circ + \gamma)] = 180^\circ - \gamma \xrightarrow[\gamma \rightarrow 0]{} 180^\circ$$



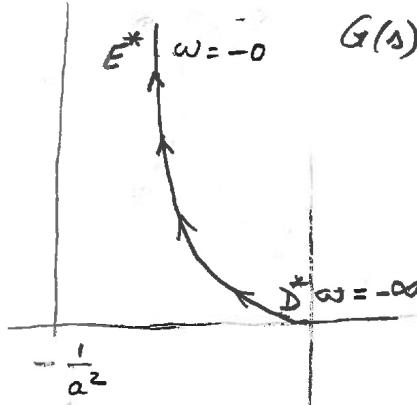
**E**  $G(-i\varepsilon) = -\frac{1}{a(-i\varepsilon)} - \frac{1}{a(-i\varepsilon+a)} = \frac{i}{a\varepsilon} - \frac{1}{a^2}$

$$G(-i0) = \lim_{\varepsilon \rightarrow 0} G(-i\varepsilon)$$

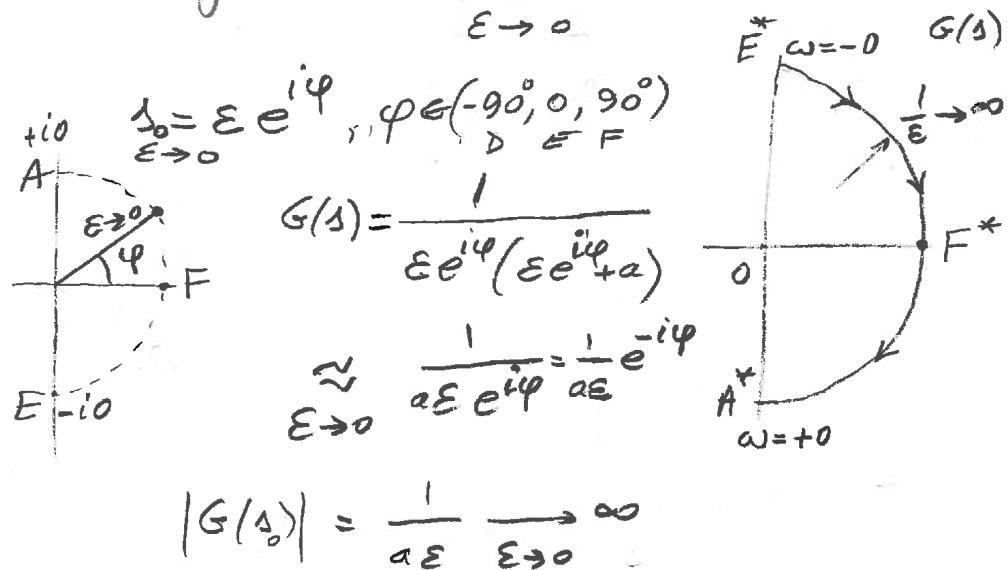
$$G(-i0) = -\frac{1}{a^2} + \frac{i}{a\varepsilon} \Big|_{\varepsilon \rightarrow 0} = -\frac{1}{a^2} + i\infty$$

$$|G(-i0)| = \infty$$

$$\angle G(-i0) = 90^\circ$$



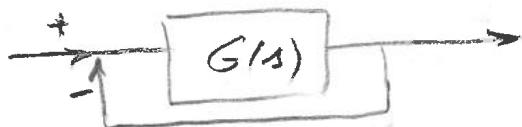
$\zeta \rightarrow 0$  Segment  $(-i\zeta, \zeta, i\zeta)$  small circle



$$\underline{G(\zeta_0)} = -\varphi \in \begin{pmatrix} 90^\circ & 0^\circ & -90^\circ \\ E^* & F^* & A^* \end{pmatrix}$$

$$\varphi = -90^\circ \quad \varphi = 0^\circ \quad \varphi = +90^\circ$$

### NYQUIST STABILITY CRITERION



Find stability of  $G_{CL} = \frac{G}{1+G}$  by  
analyzing the polar plot of  $G(s)$   
as we follow the Nyquist circuit.

$$G_{CL}(s) = \frac{G(s)}{1+G(s)}$$

Poles $G_{CL}(s)$	LHS	iω	RHS
X	+	1	> 0

Poles  $G_{CL}(s) \Rightarrow$  zeros of  $1+G(s)$

stable	$Z \leq 0$	unstable!	$Z > 0$
--------	------------	-----------	---------

Nyquist criterion:

$$Z = P + N \leq 0$$

$P$  = number of  $G(s)$  poles in RHS

$$G(s) = \frac{B(s)}{A(s)} \rightarrow P_1, P_2, \dots$$

$N$  = number of clockwise encirclements  
of the  $(-1, 0)$  point as we follow  
the Nyquist path (circuit)

$Z =$  number of zeros of  $1+G(s)$  in RHS

STABLE if  $Z \leq 0$

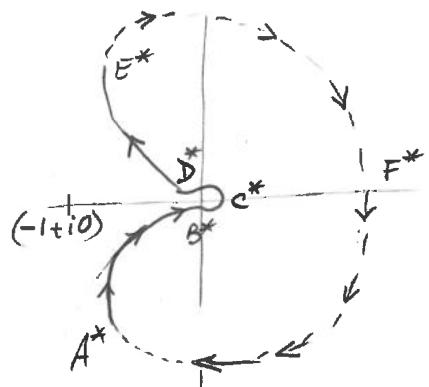
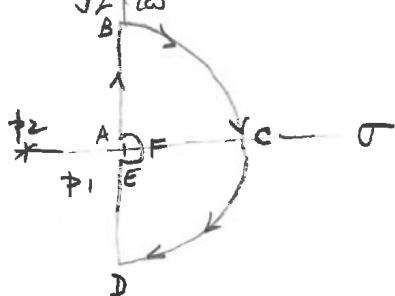
Example : aircraft roll model

$$G(s) = \frac{114}{10s^2 + 4s} = \frac{114}{s(10s+4)}$$

$$P_1 = 0, P_2 = -2/5 \therefore P = 0$$

Recall  $G(s) = \frac{1}{s(s+a)} = \frac{1}{s^2 + sa}$  no poles in RHS

Nyquist circuit



$$P = 0 \quad \text{no poles in RHS}$$

$$N = 0 \quad \text{no encirclements of } (-1+i0) \text{ point}$$

$$Z = N + P = 0 \quad \text{no CL poles in RHS}$$

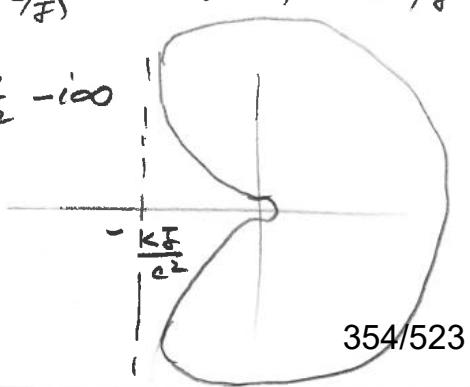
Conclusion : system is unconditionally stable

Note :  $G = \frac{K}{Js^2 + Cs} = \frac{K}{J} \frac{1}{s(s + c/J)} = \frac{B}{s(s+a)} \quad a = c/J$

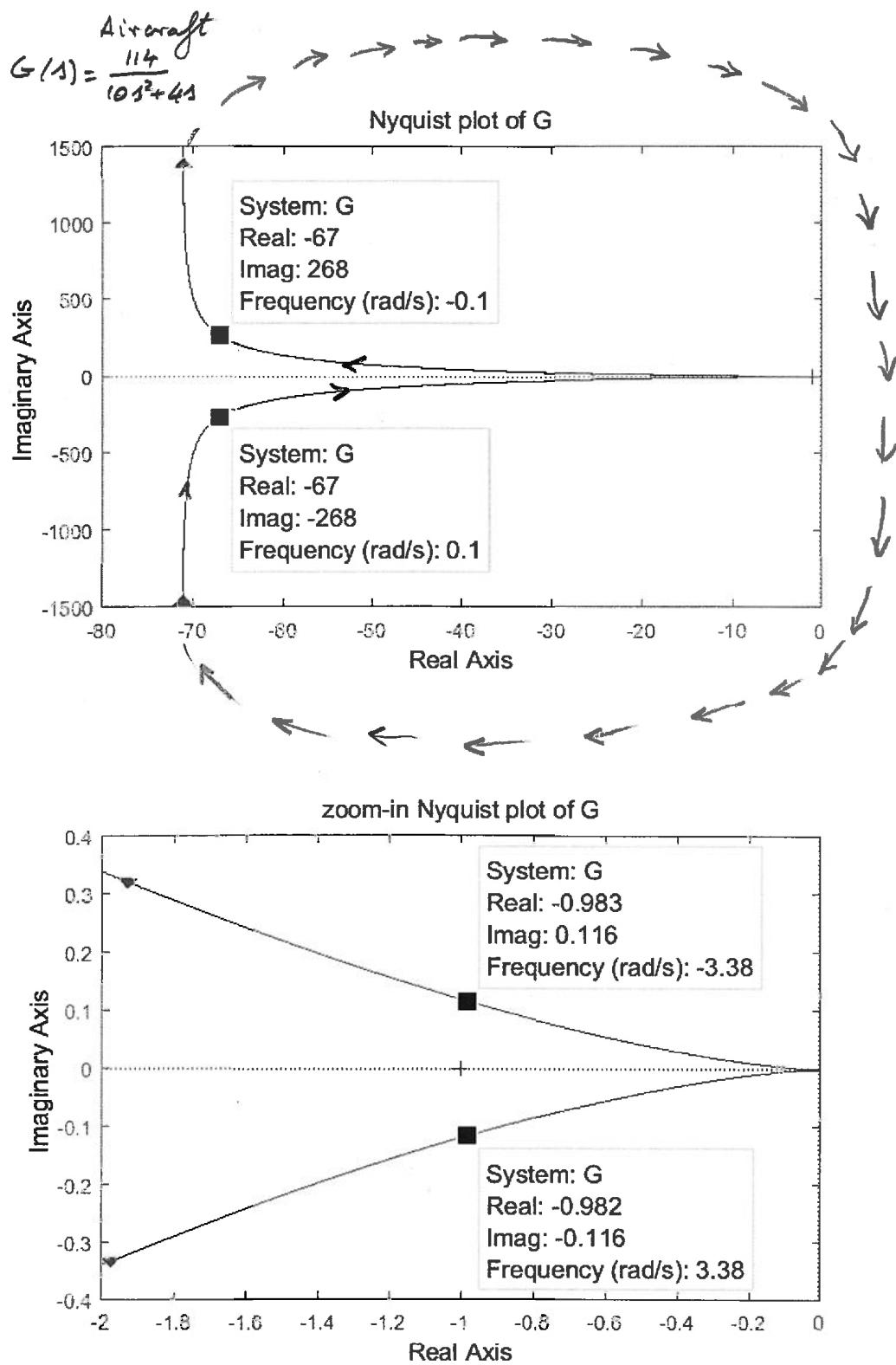
$$G(i\epsilon) = B\left(-\frac{1}{a^2} - i\infty\right) = -\frac{B}{a^2} - i\infty$$

$\epsilon \rightarrow 0$

$$\frac{B}{a^2} = \frac{K}{J} \frac{J^2}{C^2} = \frac{KJ}{C^2}$$



354/523



Example A11.8

Given:  $G = K \frac{1}{s-1}$



Note: pole in RHS,  $s=1$

Find: critical value of K for stability using Nyquist criterion

Solution: do Nyquist plot of  $G(s)$

$$G(0) = -\frac{K}{1} = -K$$

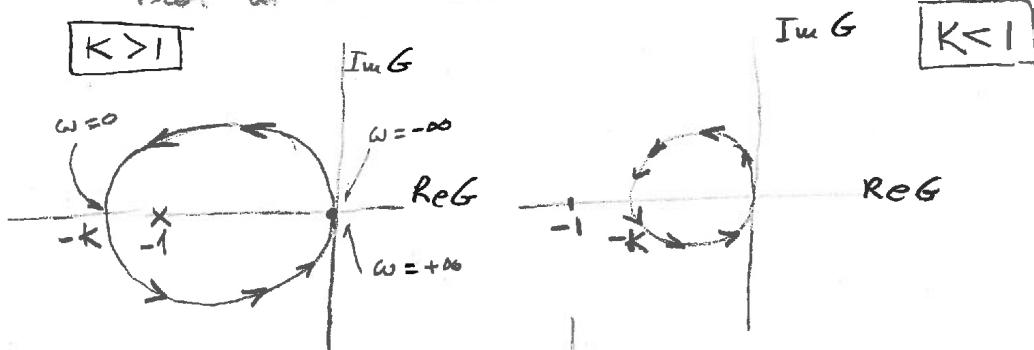
$$G(i\omega) = \lim_{\omega \rightarrow \infty} \frac{K}{i\omega - 1} \approx \frac{K}{i\omega}, |G(i\omega)| = 0$$

$\angle G(i\omega) = -90^\circ$

$$G(-i\omega) \rightarrow |G(-i\omega)| = 0$$

$\angle G(-i\omega) = 90^\circ$

Plot in MATLAB



$N = -1$  (counterclockwise)

$P = 1$  ( $s=1$  pole of  $G$  in RHS)

$$Z = -1 + 1 = 0$$

$N = 0$

$P = 1$

$$Z = 1$$

UNSTABLE for  $K < 1$

STABLE for  $K > 1$

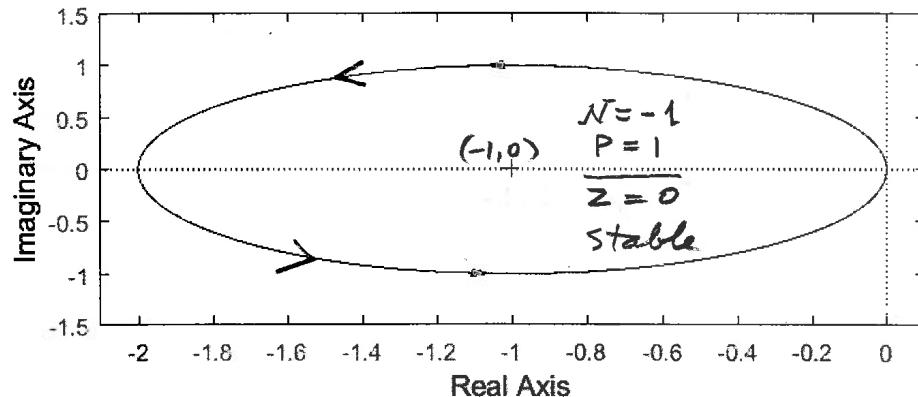
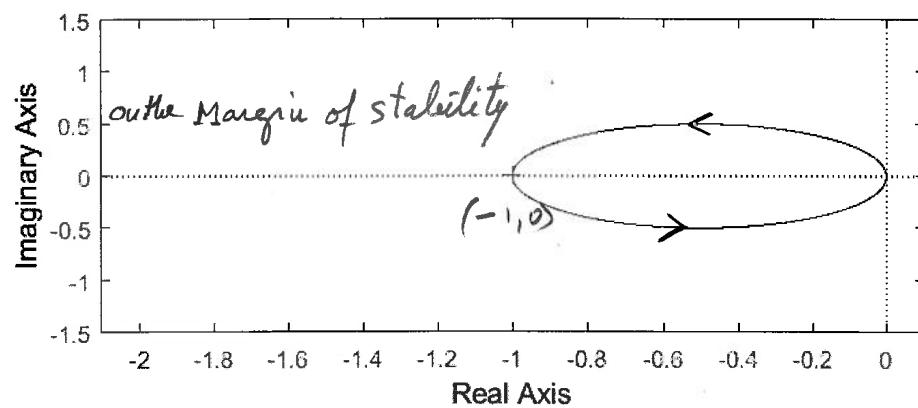
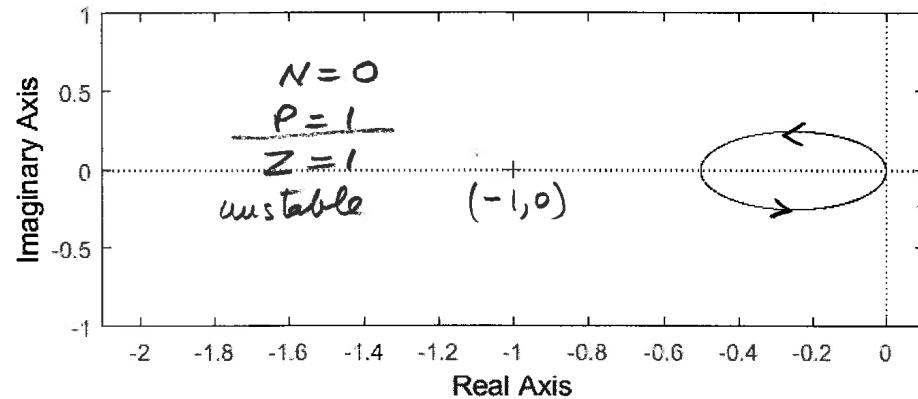
CRITICAL VALUE:  $K = 1$

356/523

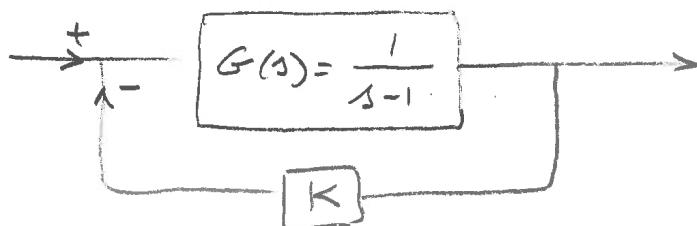
A 11.8

$$K=2 \text{ STABLE}$$

$$G = \frac{2}{s-1}$$

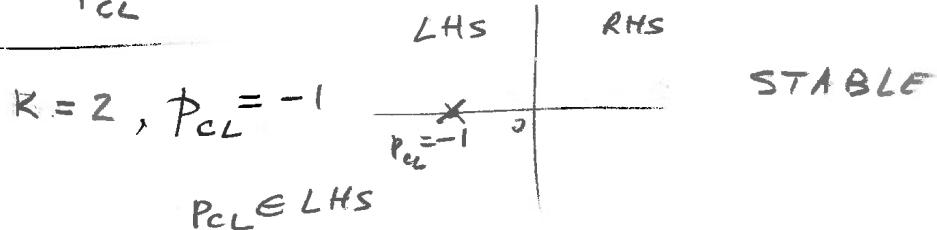
 $K=1$  $K=0.5 \text{ UNSTABLE; Ex.A11-8}$ 

A11.8

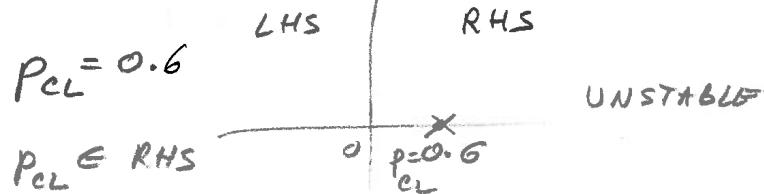


$$G_{CL} = \frac{G}{1+GH} = \frac{1}{s-1+K} = \frac{1}{s-(1-K)}$$

$$\rho_{CL} = 1-K$$

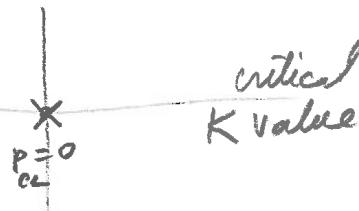


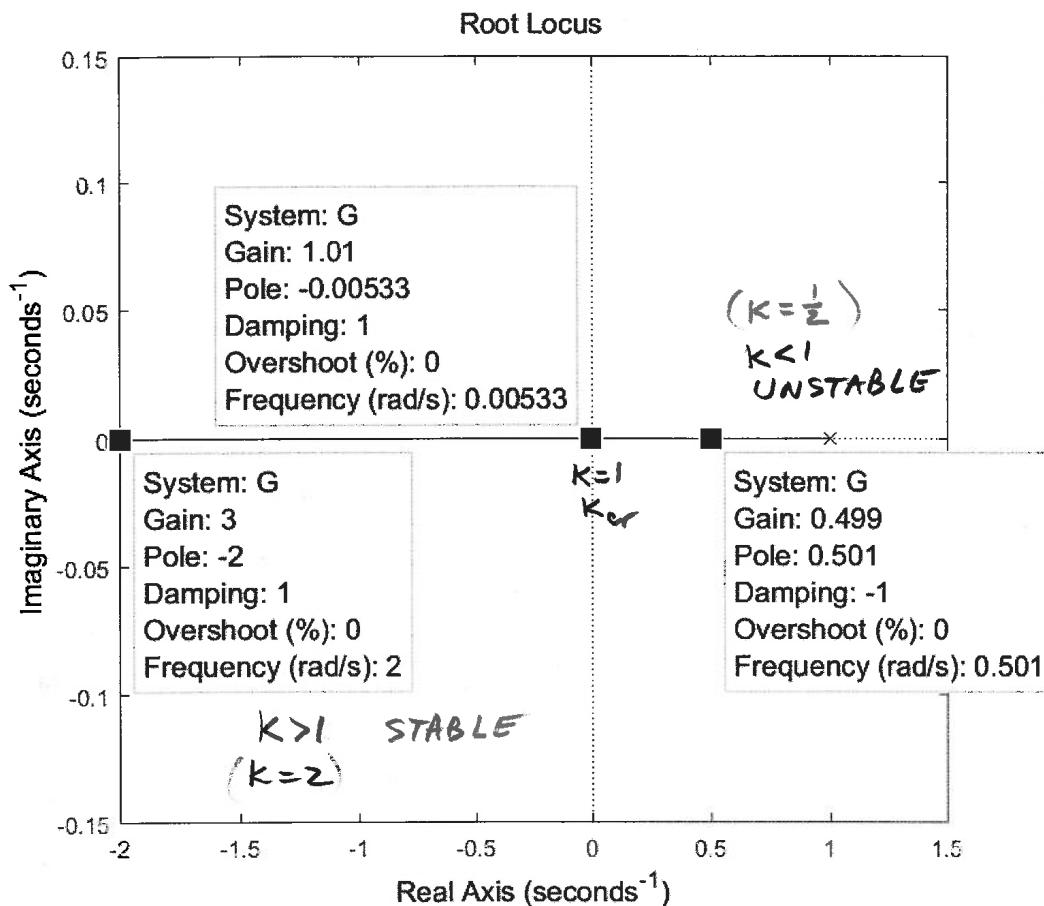
$$K = 0.4, \rho_{CL} = 0.6$$



$$\rho_{CL} = 1 - K_{cr} = 0$$

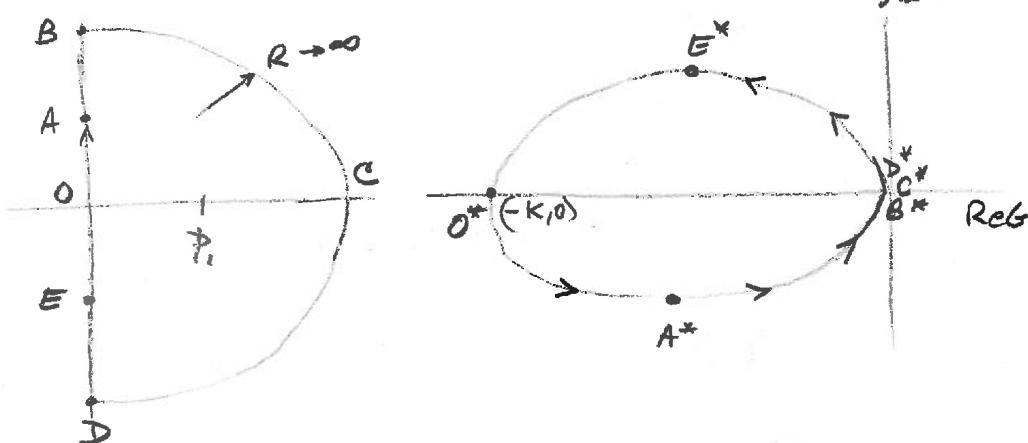
$$K_{cr} = 1$$



*All. 8*

All B.C. MANUAL PLOT

$$G = K \frac{1}{s-1} \quad \neq_1 = 1$$



$$0: \quad G(0) = K \frac{1}{0-1} = -K = K e^{i\pi}$$

$$A = i \quad G(i) = K \frac{1}{i-1} = -K \frac{1+i}{2} = K \frac{\sqrt{2}}{2} e^{i(\pi + \frac{\pi}{4})}$$

$$\angle G(i) = \frac{5\pi}{4}$$

$$B = iR \quad G(iR) = K \frac{1}{iR-1} = -K \frac{1+iR}{1+R^2} = \frac{K}{\sqrt{1+R^2}} e^{i(\pi + \varphi)}$$

$$|G(iR)| \xrightarrow[R \rightarrow \infty]{} 0 \quad \varphi = \tan^{-1} R \xrightarrow[R \rightarrow \infty]{} \frac{\pi}{2}$$

$$\angle G(iR) \xrightarrow[R \rightarrow \infty]{} \frac{3\pi}{2}$$

$$C = R \quad G(R) = K \frac{1}{R-1} \xrightarrow[R \rightarrow \infty]{} 0$$

$$D = -iR \quad G(-iR) = K \frac{1}{-iR-1} = -K \frac{1-iR}{1+R^2} = \frac{K}{\sqrt{1+R^2}} e^{i(\pi - \varphi)}$$

$$|G(-iR)| \xrightarrow[R \rightarrow \infty]{} 0 \quad \varphi = \tan^{-1} R \xrightarrow[R \rightarrow \infty]{} \frac{\pi}{2}$$

$$\angle G(-iR) \xrightarrow[R \rightarrow \infty]{} \pi - \frac{\pi}{2} = \frac{\pi}{2}$$

360/523

All 86

$$E = -i \quad G(\omega) = K \frac{1}{-i-1} = -K \frac{1-i}{2} = K \frac{\sqrt{2}}{2} e^{i(\pi - \frac{\pi}{4})}$$

$$\angle G(\omega) = \frac{3\pi}{4}$$

Nyquist criterion gives  $K < 1$  UNSTABLE

$K > 1$  STABLE

To verify, calculate the poles of  $G_{CL}$

$$G_{CL} = \frac{G}{1+G} = \frac{K}{s-1+K} = \frac{K}{s-(1-K)}$$

$$P_{CL} = 1-K$$

for  $K < 1$ ,  $P_{CL} > 0$ , in RHS, UNSTABLE

$K > 1$ ,  $P_{CL} < 0$ , in LHS, STABLE

QED

p634

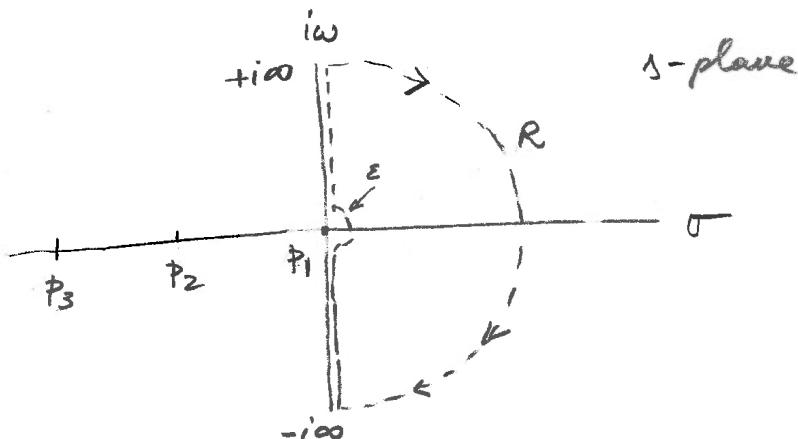
Example p634

$$G(s) = K \frac{1}{s(T_1 s + 1)(T_2 s + 1)} \quad T_2 > T_1$$

Poles :  $p_1 = 0$ 

$$p_2 = -\frac{1}{T_1}$$

$$p_3 = -\frac{1}{T_2}$$



Nyquist circuit will have to go around the origin on a small circle of radius  $\epsilon \rightarrow 0$ . We study four segments:

- positive  $i\omega$  axis  $s\epsilon(+i0, +i\infty)$
- big circle  $R \rightarrow \infty$
- negative  $i\omega$  axis  $s\epsilon(-i\infty, -i0)$
- small circle  $\epsilon \rightarrow 0$

P634  $\omega + i\omega \text{ axis.}$   
 $\Delta = (+i\epsilon, iR)$

$$G(+i\epsilon) = -K(T_1 + T_2) - i\frac{K}{\epsilon}$$

$$G(+i\epsilon) \xrightarrow{\epsilon \rightarrow 0} -K(T_1 + T_2) - i\infty$$

$$|G(i\epsilon)| \xrightarrow[\epsilon \rightarrow 0]{} \infty$$

$$\angle G(i\epsilon) \xrightarrow[\epsilon \rightarrow 0]{} -90^\circ$$

$$\epsilon \rightarrow 0 \\ R \rightarrow \infty$$

Proof

$$\begin{aligned} \frac{K}{i\epsilon(i\epsilon T_1 + 1)(i\epsilon T_2 + 1)} &= \\ \frac{K}{i\epsilon[i\epsilon^2 T_1 T_2 + i\epsilon(T_1 + T_2) + 1]} &= \\ \frac{K}{i\epsilon \frac{i\epsilon(T_1 + T_2) + 1}{i\epsilon^2(T_1 + T_2) + 1}} &= \\ \frac{K}{i\epsilon} \frac{-i\epsilon(T_1 + T_2) + 1}{i\epsilon^2(T_1 + T_2) + 1} &= \\ = -K(T_1 + T_2) - i\frac{K}{\epsilon} & \end{aligned}$$

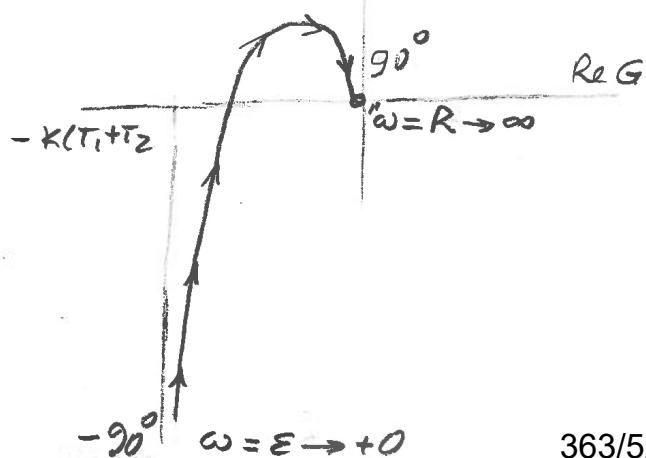
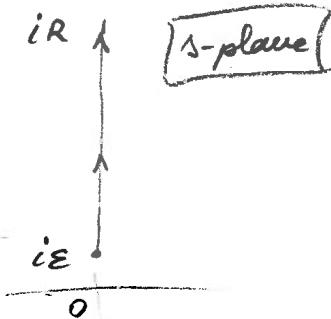
$$G(iR) = \frac{K}{iR(T_1 iR + 1)(T_2 iR + 1)} \approx \frac{K}{T_1 T_2 R^3 i^3} = \frac{K}{T_1 T_2 R^3} e^{i\frac{\pi}{2}}$$

$$\frac{1}{i^3} = \frac{i^4}{i^3} = i = e^{i\frac{\pi}{2}}$$

$$|G(iR)| = \frac{K}{T_1 T_2 R^3} \xrightarrow[R \rightarrow \infty]{} 0$$

$$\angle G(iR) = \frac{\pi}{2} = 90^\circ$$

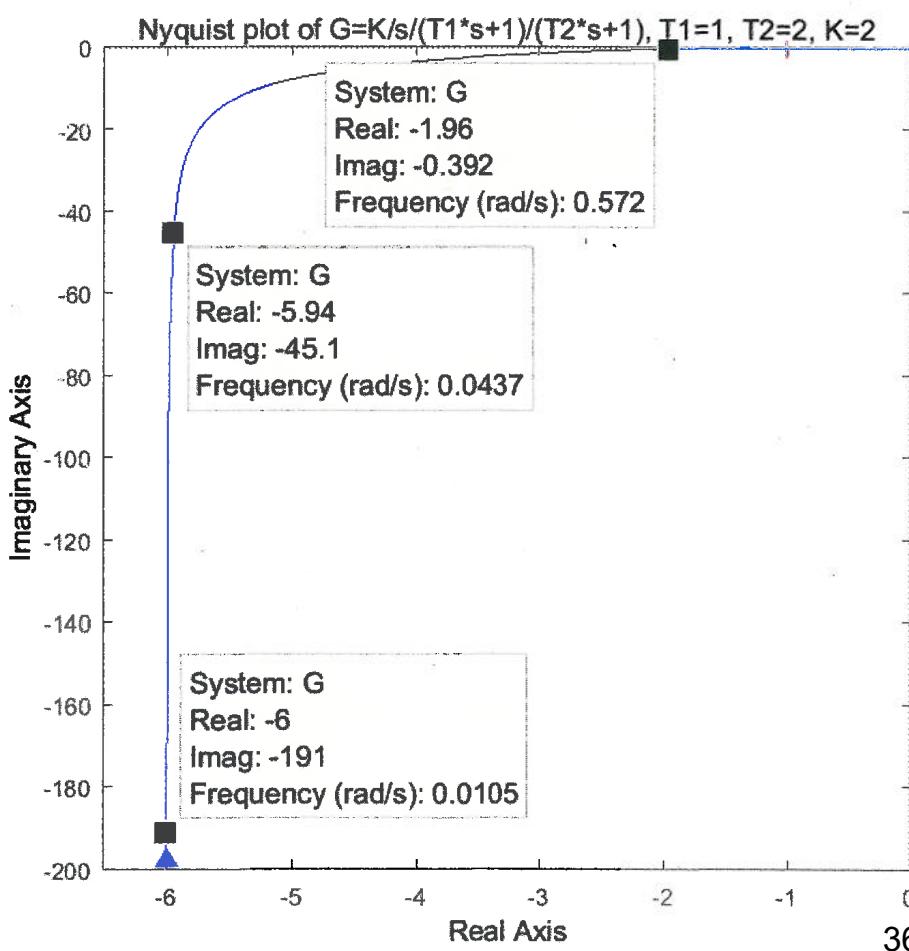
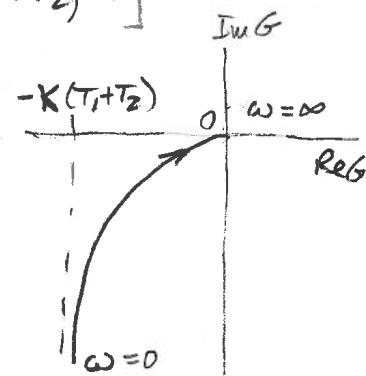
Inv G



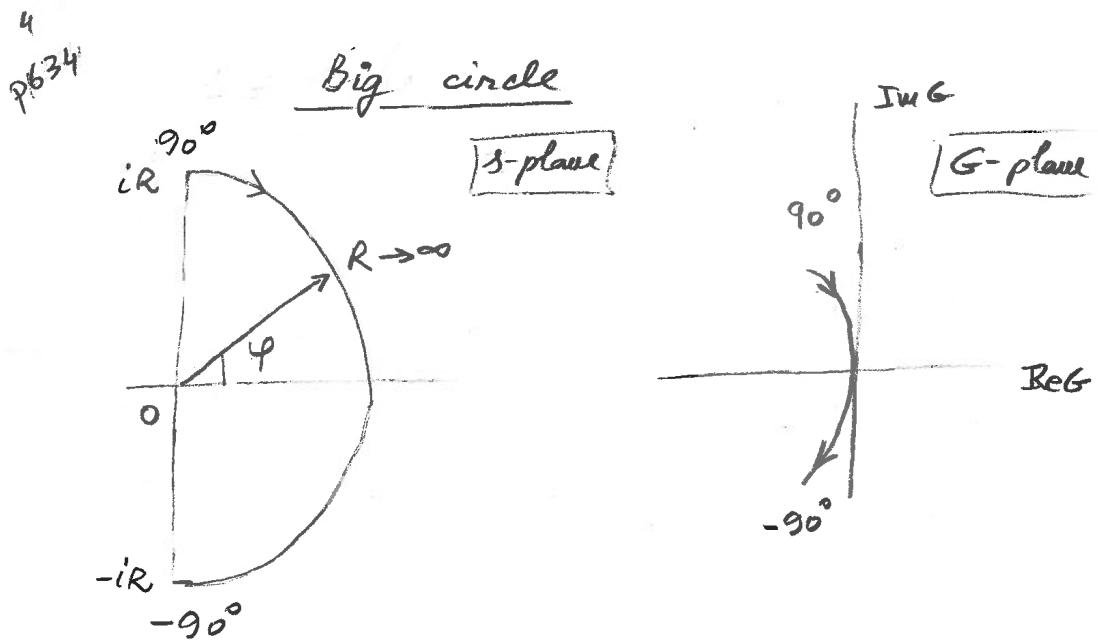
363/523

P634

$$\begin{aligned}
 G(i\epsilon) &= \frac{K}{i\epsilon(i\epsilon T_1 + 1)(i\omega T_2 + 1)} \\
 &= \frac{K}{i\epsilon(i^2\epsilon^2 T_1 T_2 + i\epsilon(T_1 + T_2) + 1)} = \frac{K}{\epsilon} \frac{1}{-\epsilon(T_1 + T_2) + i} \\
 &= \frac{K}{\epsilon} \frac{-\epsilon(T_1 + T_2) - i}{\epsilon^2(T_1 + T_2)^2 + 1} = -\frac{K}{\epsilon} [\epsilon(T_1 + T_2) + i] \\
 G(i\epsilon) &= -K(T_1 + T_2) - i\frac{K}{\epsilon} \\
 \xrightarrow{\epsilon \rightarrow 0} & -K(T_1 + T_2) - i\infty \\
 &\quad \text{Re } G = \text{const}
 \end{aligned}$$



364/523



$$G(s) = \frac{K}{Re^{i\varphi}(T_1 Re^{i\varphi} + 1)(T_2 Re^{i\varphi} + 1)}$$

$$G(s) \underset{R \gg 1}{\approx} \frac{K}{T_1 T_2 R^3 e^{3i\varphi}} = \frac{K}{T_1 T_2 R^3} e^{-3i\varphi}$$

$$|G(s)| = \frac{K}{T_1 T_2 R^3} \xrightarrow[R \rightarrow \infty]{} 0$$

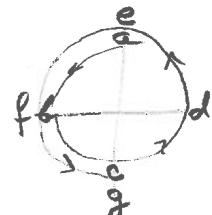
$$\angle G(s) = -3\varphi = (-270^\circ, +270^\circ)$$

$$(-270 + 360)^\circ = 90^\circ; 270 - 360 = -90^\circ$$

$$\angle G(s) = (90^\circ, -90^\circ)$$

$$\begin{array}{cccccccccc} \varphi = 90^\circ & 60^\circ & 30^\circ & 0^\circ & -30^\circ & -60^\circ & -90^\circ \\ \angle G -270 & -180 & -90 & 0 & 90 & 180 & 270 \\ 90 & 180 & -90 & 0 & 90 & 180 & -90 \\ a & b & c & d & e & f & g \end{array}$$

$G(s)$  goes round origin 1.5 times



365/523

p634

$$\underline{s} = \frac{-i\omega \text{ axis}}{(-iR - i\varepsilon)}$$

$$\varepsilon \rightarrow 0$$

$$R \rightarrow \infty$$

$$G(-i\varepsilon) \approx -K(T_1 + T_2) + i\frac{K}{\varepsilon}$$

$$G(-i\varepsilon) \xrightarrow{\varepsilon \rightarrow 0} -K(T_1 + T_2) + i\infty$$

$$|G(-i\varepsilon)| \xrightarrow{\varepsilon \rightarrow 0} \infty$$

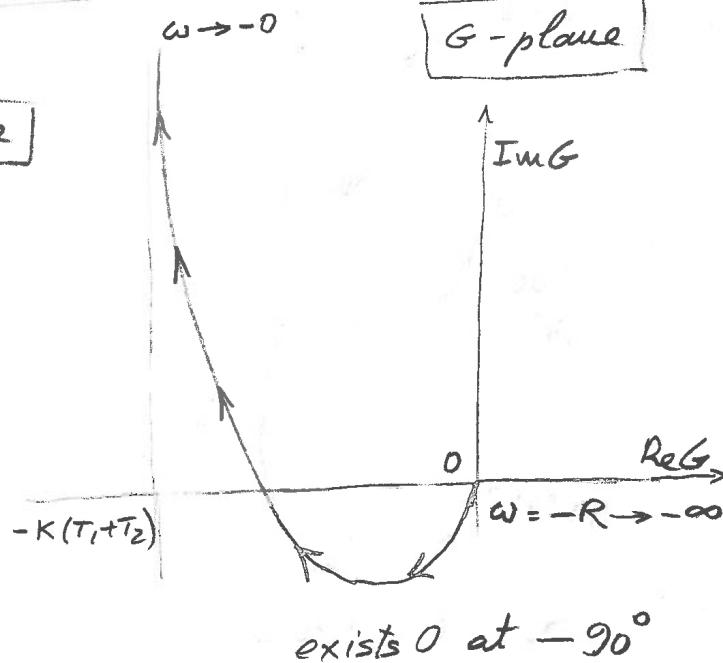
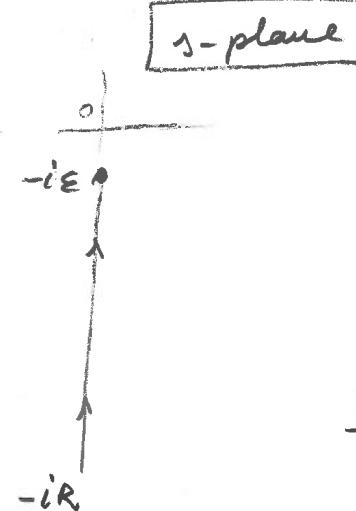
$$\angle G(-i\varepsilon) \xrightarrow{\varepsilon \rightarrow 0} 90^\circ$$

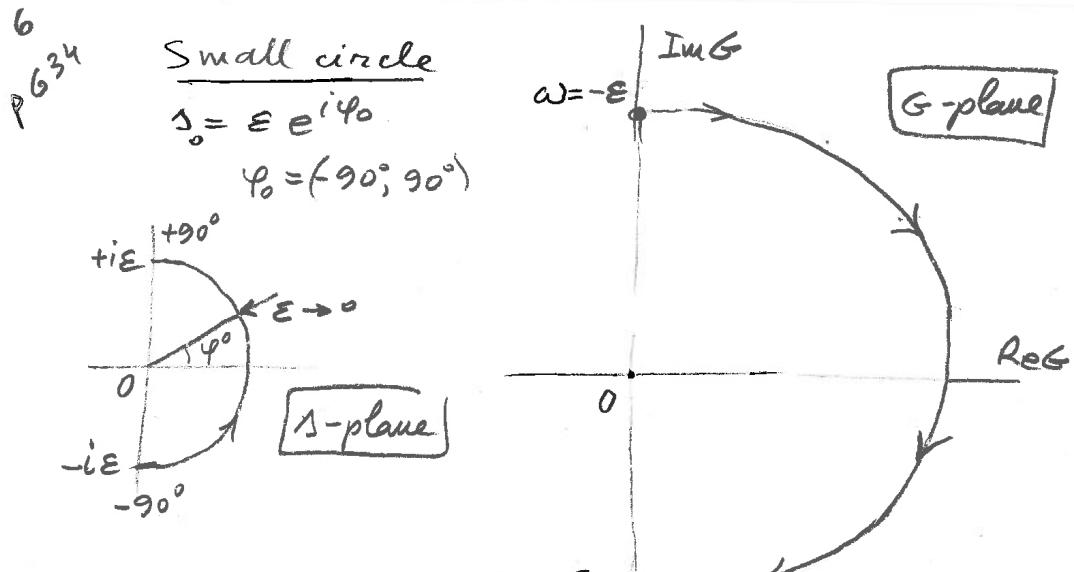
$$G(-iR) = \frac{K}{-iR(-T_1iR+1)(-T_2iR+1)} \approx \frac{K}{-T_1T_2R^3i^3} = \frac{K}{T_1T_2R^3} e^{-i\frac{\pi}{2}}$$

$$\frac{-1}{i^3} = \frac{i^2}{i^3} = \frac{1}{i} = -i$$

$$|G(-iR)| = \frac{K}{T_1T_2R^3} \xrightarrow{R \rightarrow \infty} 0$$

$$\angle G(iR) = -\frac{\pi}{2} = -90^\circ$$





$$G(s_0) = \frac{K}{\varepsilon e^{i\varphi_0} (T_1 \varepsilon e^{i\varphi_0} + 1)(T_2 \varepsilon e^{i\varphi_0} + 1)}$$

$$G(s_0) \approx \frac{K}{\varepsilon e^{i\varphi_0}} = \frac{K}{\varepsilon} e^{-i\varphi_0}$$

$$|G(s_0)| = \frac{K}{\varepsilon} \xrightarrow{\varepsilon \rightarrow 0} \infty$$

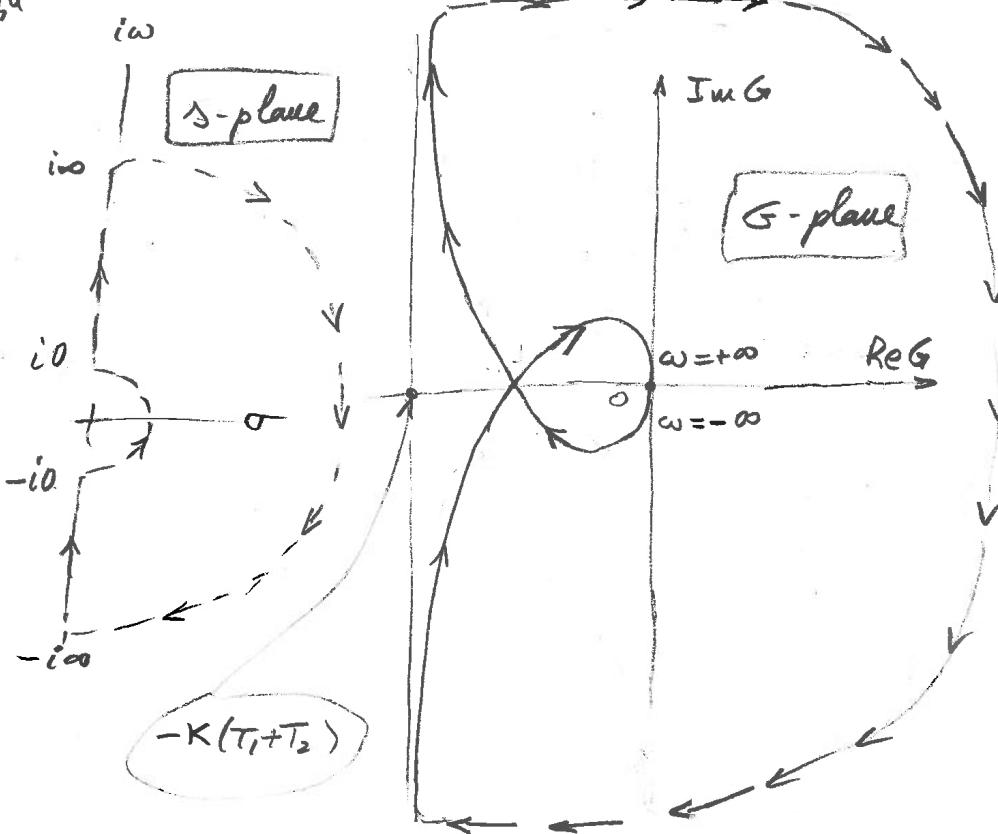
$$\angle G(s_0) = -\varphi_0 = (90^\circ, -90^\circ)$$

$$\varphi_0 = -90^\circ, \frac{90^\circ}{-i\varepsilon, i\varepsilon}$$

$G(s_0)$  travels a big semicircle from  $90^\circ$  to  $-90^\circ$ .

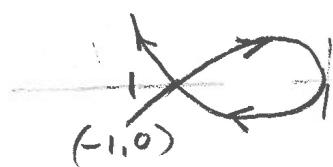
8  
P<sub>364</sub>  
7  
P<sub>634</sub>

Finally, assemble the four segments:



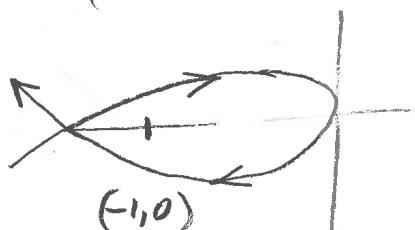
Note that  $G(s)$  goes around two times!

Depending on  $K$ , it may or may not enclose the point  $(-1, 0)$ .



STABLE

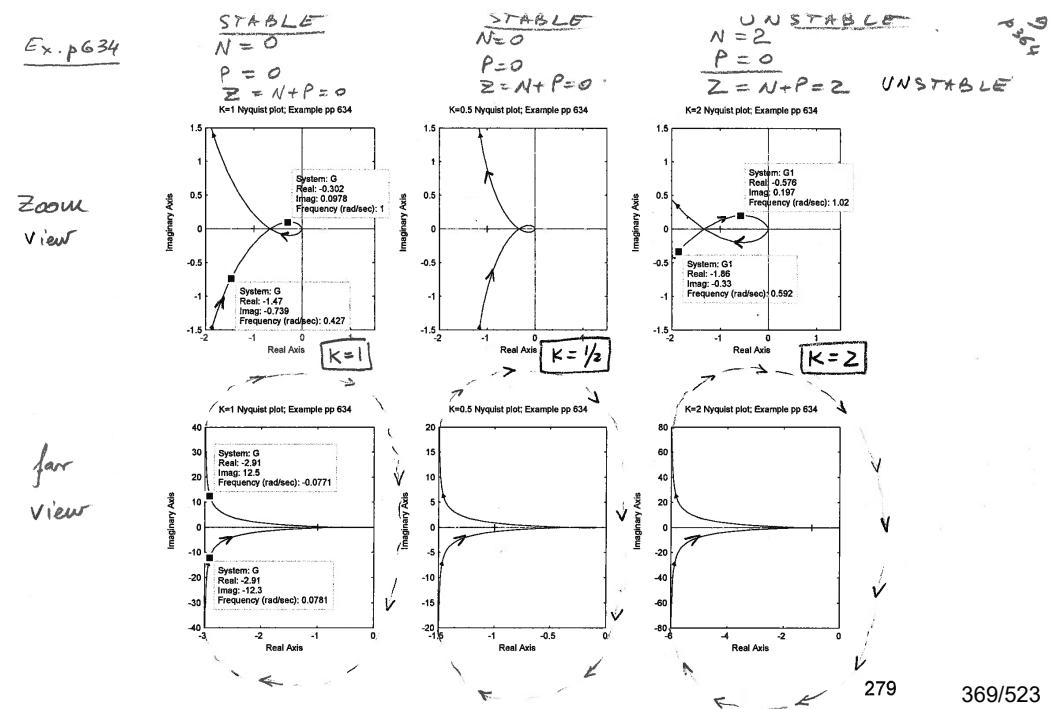
$$\frac{N=0}{\begin{matrix} P=0 \\ Z=0 \end{matrix}}$$



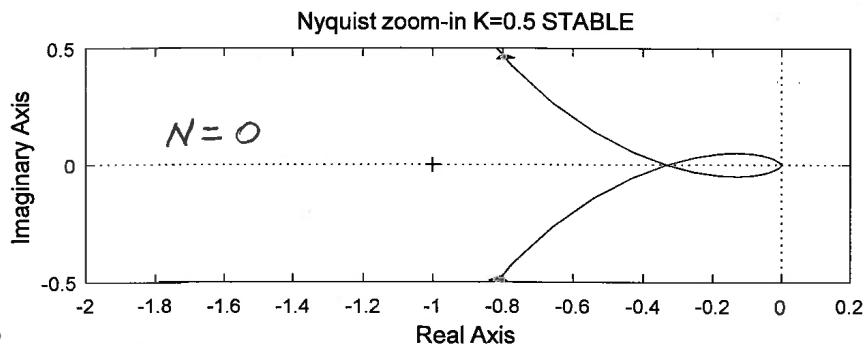
UNSTABLE

$$\frac{N=2}{\begin{matrix} P=0 \\ Z=2 \end{matrix}}$$

(see MATLAB plot next page) 368/523



10  
PG34



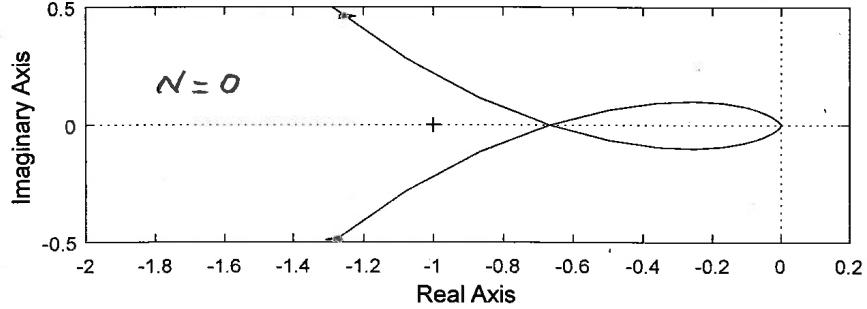
$N = 0$

$P = 0$

$Z = 0$

stable

Nyquist zoom-in  $K=1$  STABLE



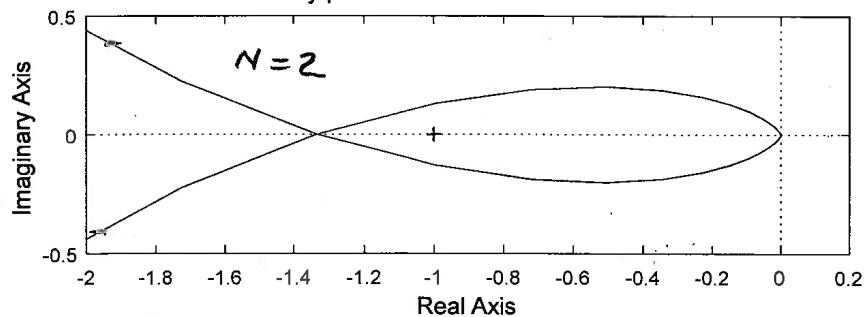
$N = 2$

$P = 0$

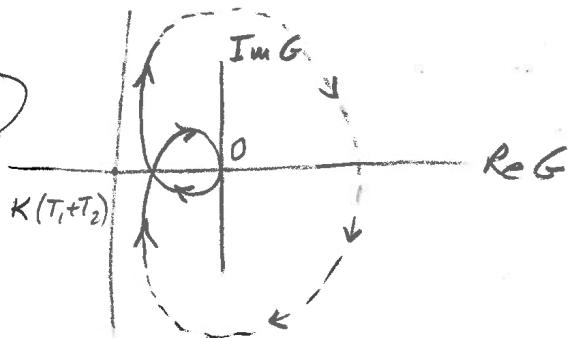
$Z = 2$

unstable

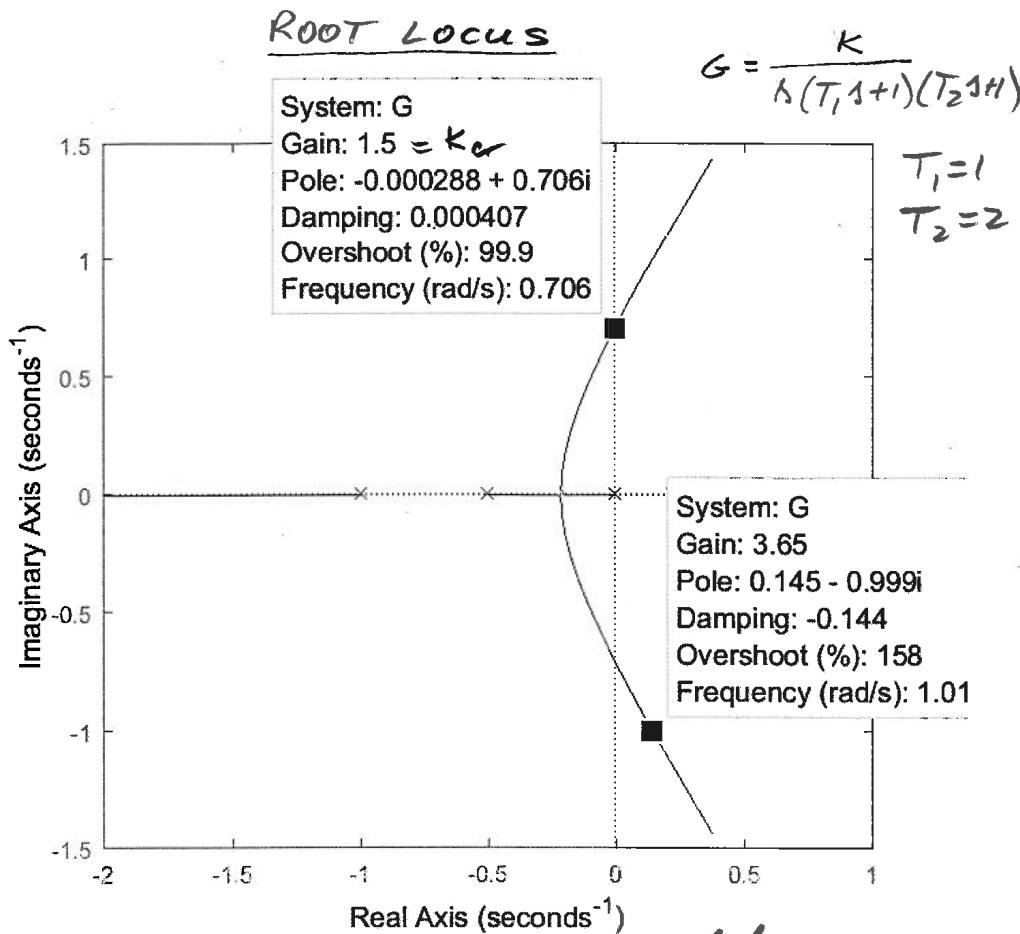
Nyquist zoom-in  $K=2$  UNSTABLE



two times  
around



370/523



$K_{cr} = 1.5$  according to rlocus plot

Verify using  $G_{CL} = \frac{G}{1+G} = \frac{K}{s(T_1s+1)(T_2s+1)+K}$

$$s(T_1s+1)(T_2s+1)+K \Big|_{\substack{T_1=1 \\ T_2=2}} = s(s+1)(2s+1)+K$$

$$s(2s^2+3s+1)+K = 2s^3+3s^2+s+K$$

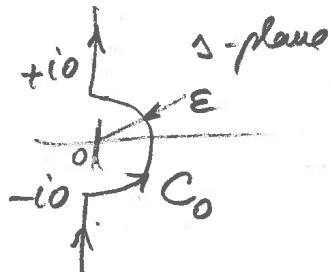
For sign change:  
 $3-2K=0$

$$K_{cr} = \frac{3}{2} = 1.5$$

Routh criterion	
$s^3$	2 1
$s^2$	3 $K$
$s^1$	$\frac{3-2K}{3}$
$s^0$	$K$

QED.

N3d

Tips on Nyquist Plot

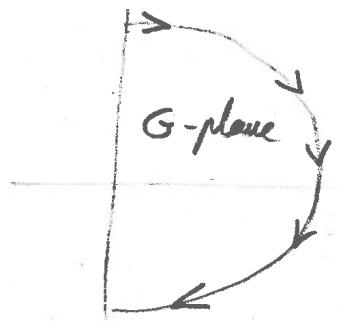
$$s_0 = \epsilon e^{i\phi}$$

$$-90^\circ < \phi < 90^\circ$$

$$G = \frac{1}{s}, \quad G(s_0) = \frac{1}{\epsilon e^{i\phi}} = \frac{1}{\epsilon} e^{-i\phi}$$

(Type 1 system)

$$+90^\circ > \angle G(s_0) > -90^\circ$$

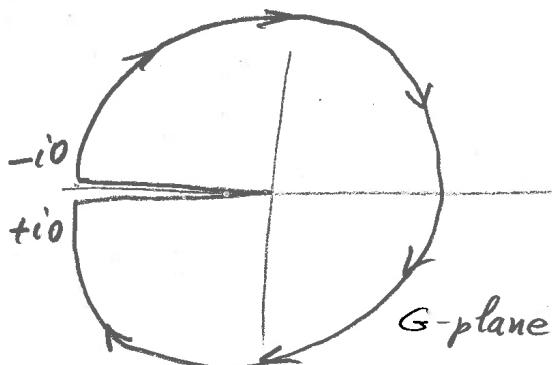


Half-circle, clockwise.

$$G = \frac{1}{s^2}, \quad G(s_0) = \frac{1}{(\epsilon e^{i\phi})^2} = \frac{1}{\epsilon^2} e^{-i2\phi}$$

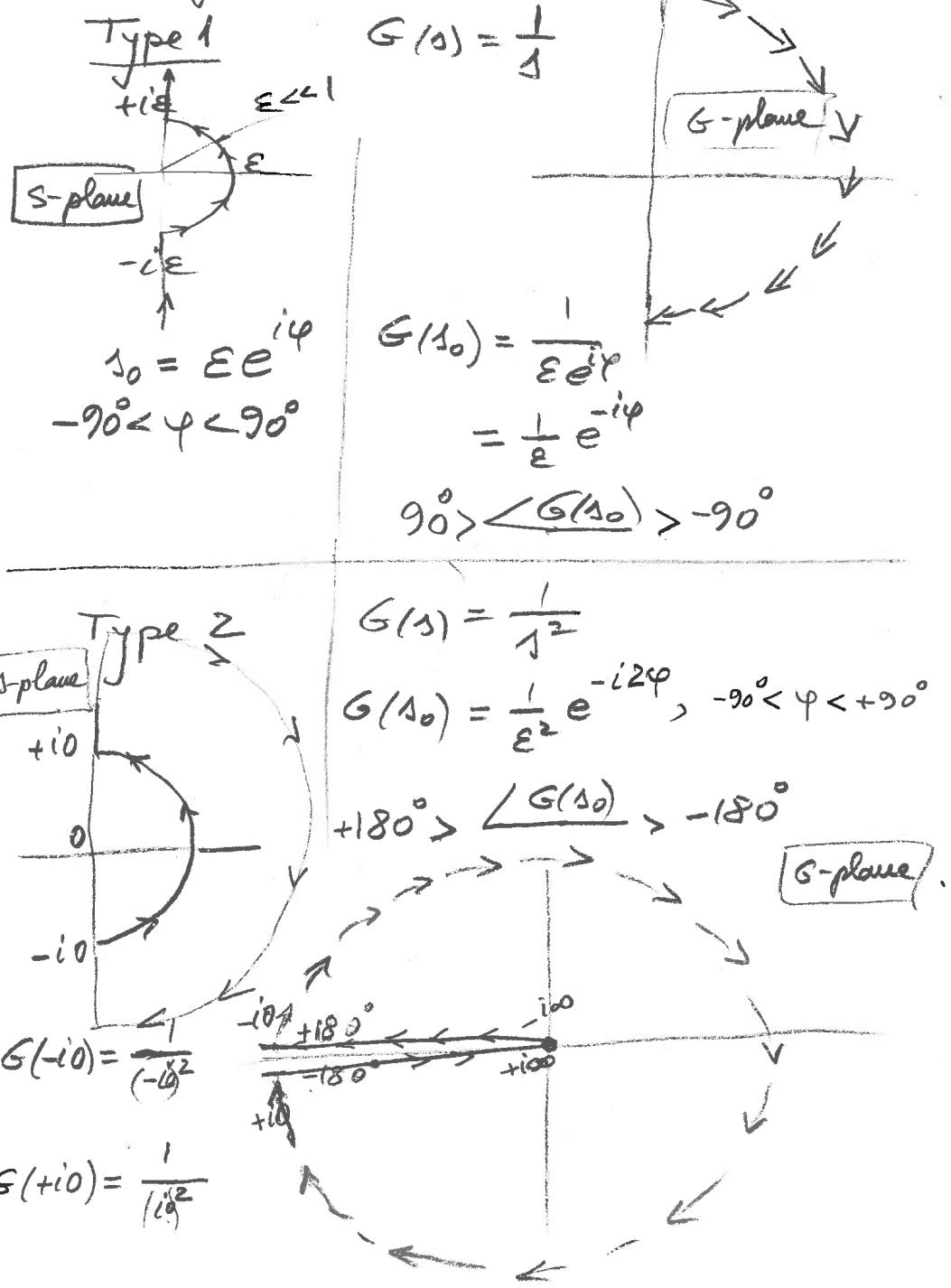
(Type 2 system)

$$+180^\circ > \angle G(s_0) > -180^\circ$$



Full circle,  
clockwise.

Details on behavior of common functions on the Nyquist circuit



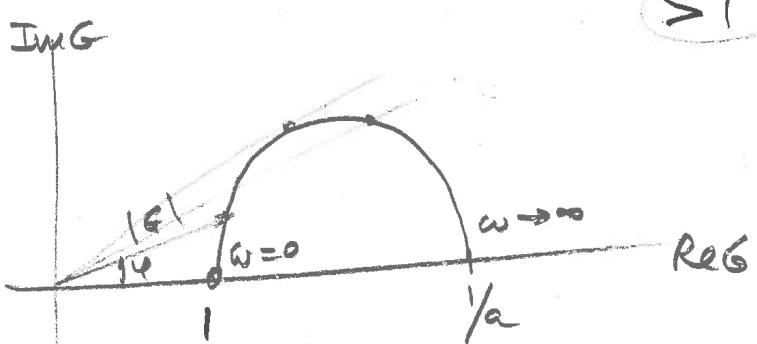
$N/a$ 

$$G(s) = \frac{Ts+1}{aTs+1} \quad 0 < a < 1$$

$$G(i\omega) = \frac{T(i\omega)+1}{a \cdot T(i\omega)+1}$$

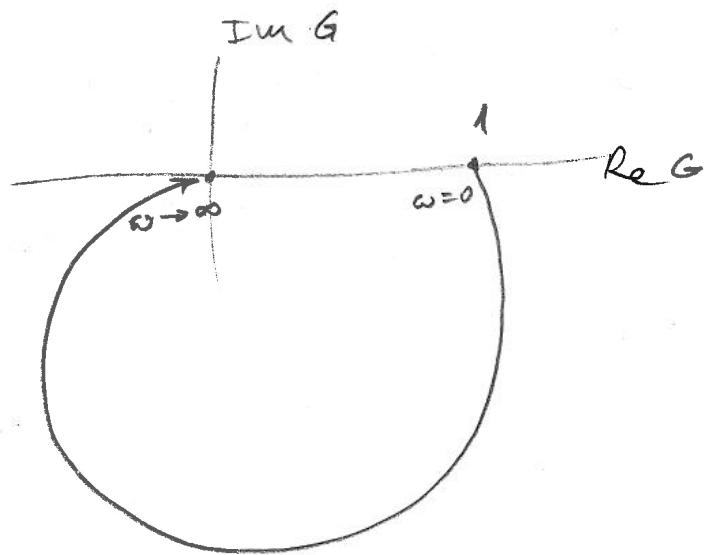
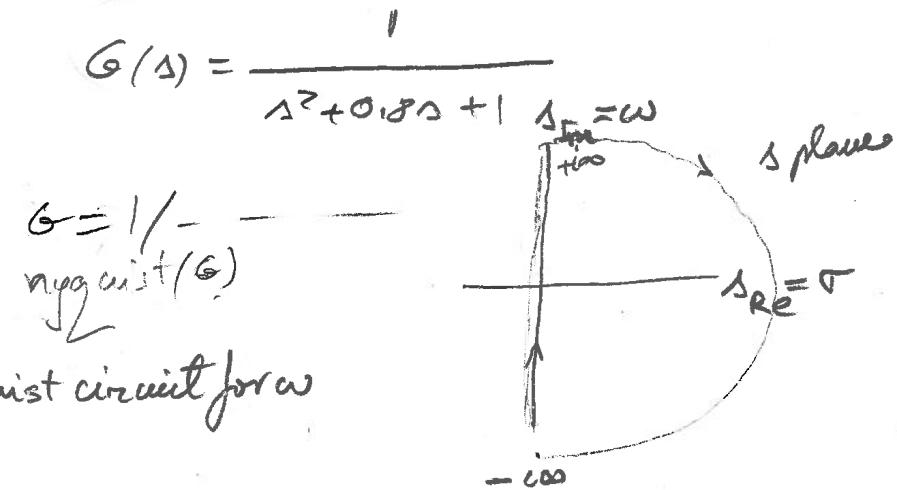
$$\omega \rightarrow 0, \quad G(i0) = 1$$

$$\omega \rightarrow \infty, \quad G(i\infty) = \lim_{\omega \rightarrow \infty} \frac{T(i\omega)+1}{a \cdot T(i\omega)+1} = \frac{T}{aT} = \frac{1}{a} > 1$$



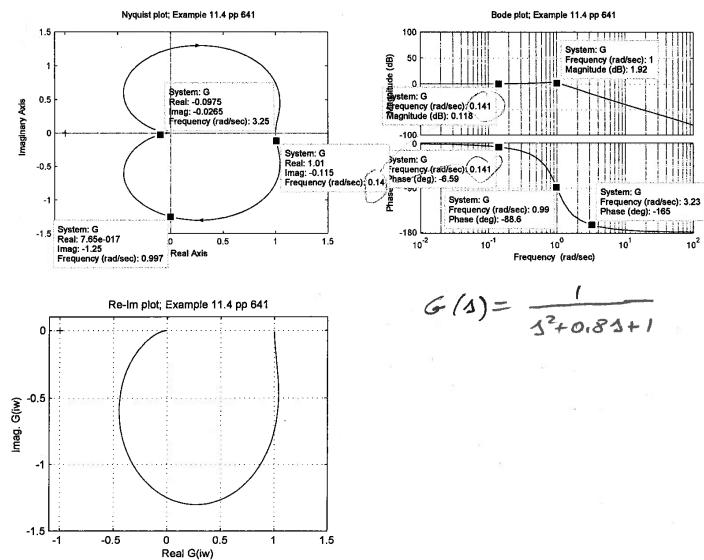
N16-

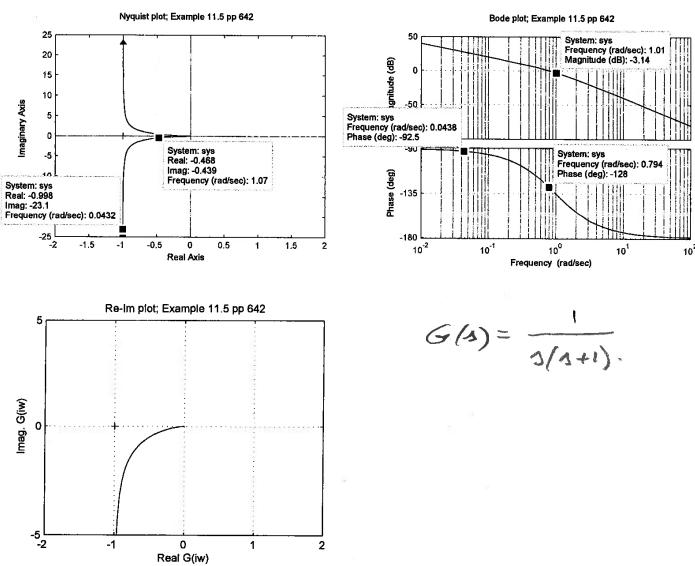
Ex 11.4 , p 641



Ex. 11.5 p 642

$$G(s) = \frac{1}{s(1+s)}$$





$$G(s) = \frac{1}{s(1+s)}$$

247

377/523

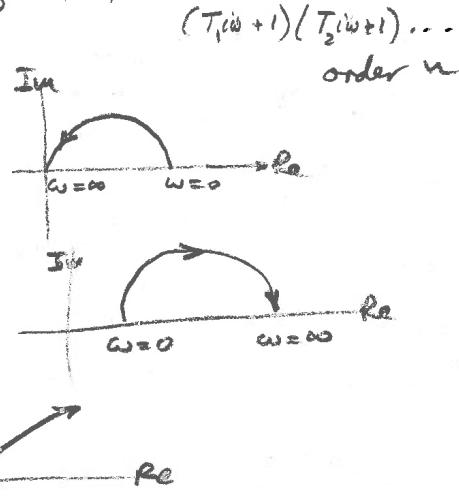
N2 Nyquist plot trends

$$G(s) = \frac{K}{s^N} \frac{(T_1 s + 1)(T_2 s + 1) \dots}{(T_1 s + 1)(T_2 s + 1) \dots}$$

Type 0 systems ( $N=0$ ) ,  $G(i\omega) = K \frac{(T_1 i\omega + 1)(T_2 i\omega + 1) \dots}{(T_1 i\omega + 1)(T_2 i\omega + 1) \dots}$

$$\omega = 0 \quad G(i0) = K$$

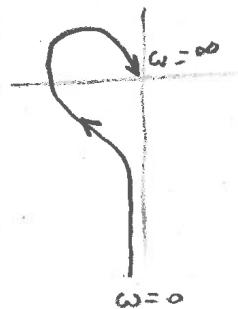
$$\omega \rightarrow \infty \quad G(i\infty) = \begin{cases} 0 & \text{for } n > m \\ \text{const.}, n = m \\ \infty, & n < m \end{cases}$$

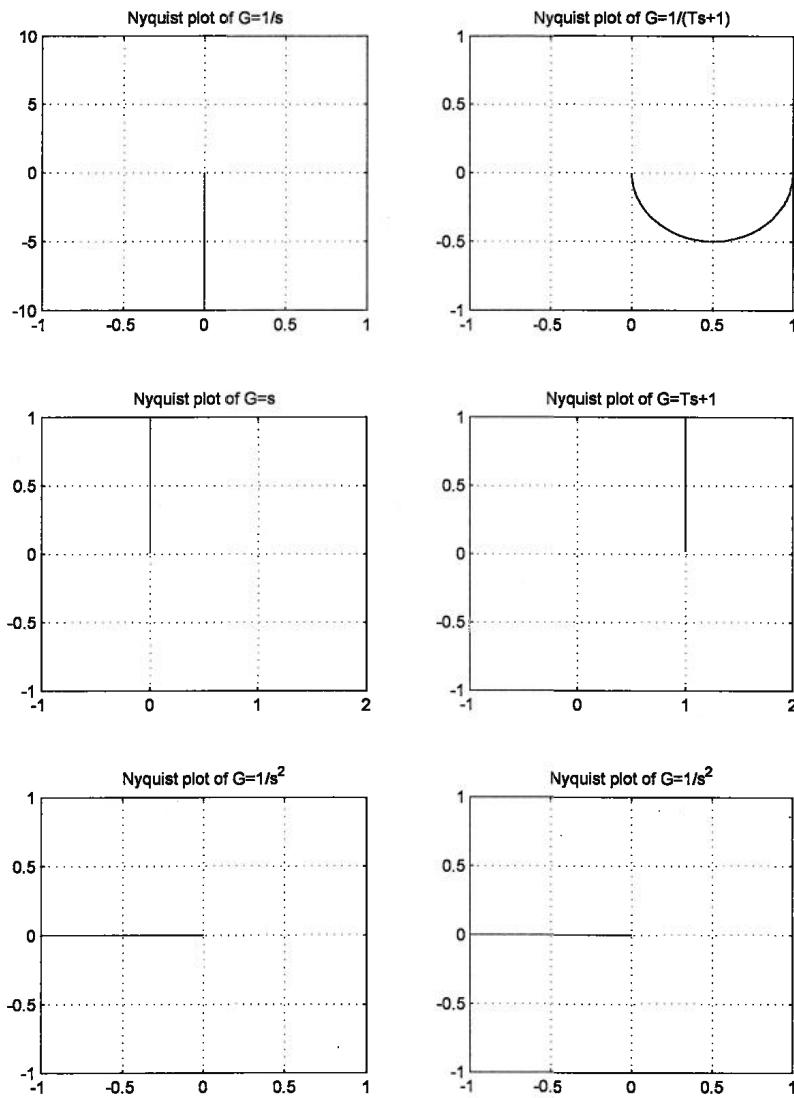


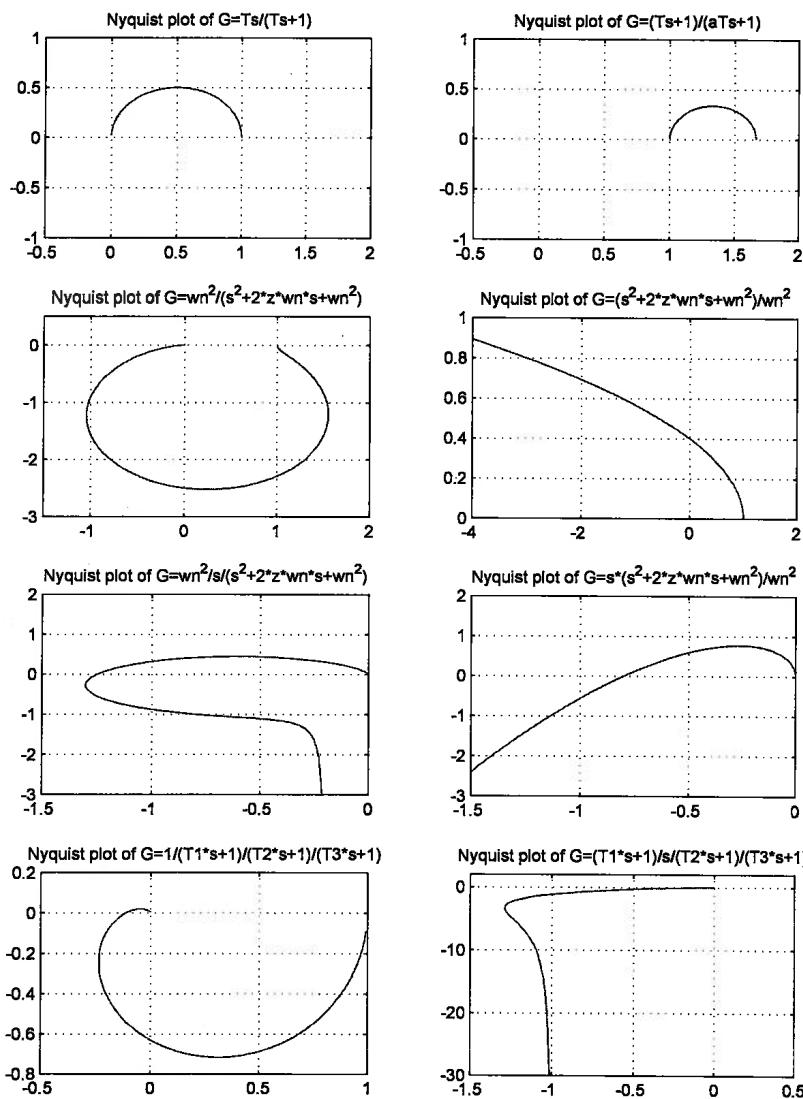
Type 1 systems ( $N=1$ ) ,  $G(i\omega) = \frac{K}{i\omega} \cdot \frac{(\ )^m}{(\ )( )^n}$

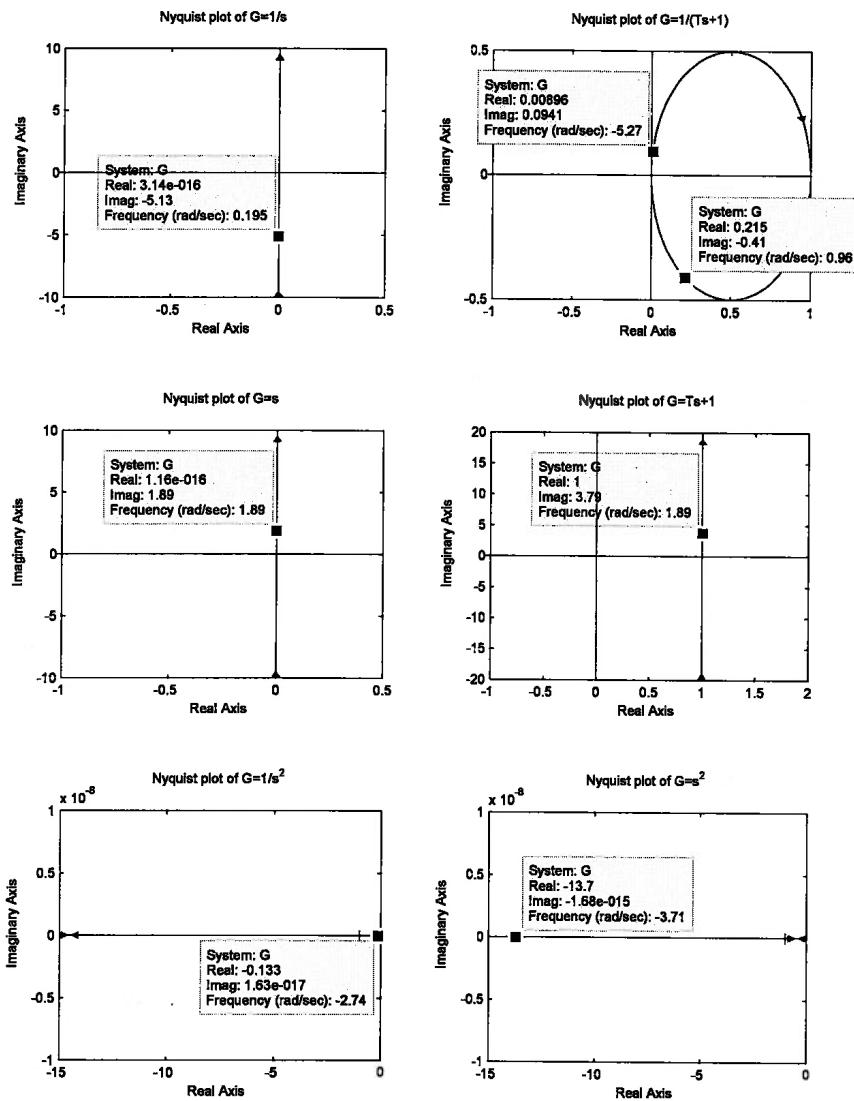
for  $n=m$  ,  $G(i\infty) = 0$

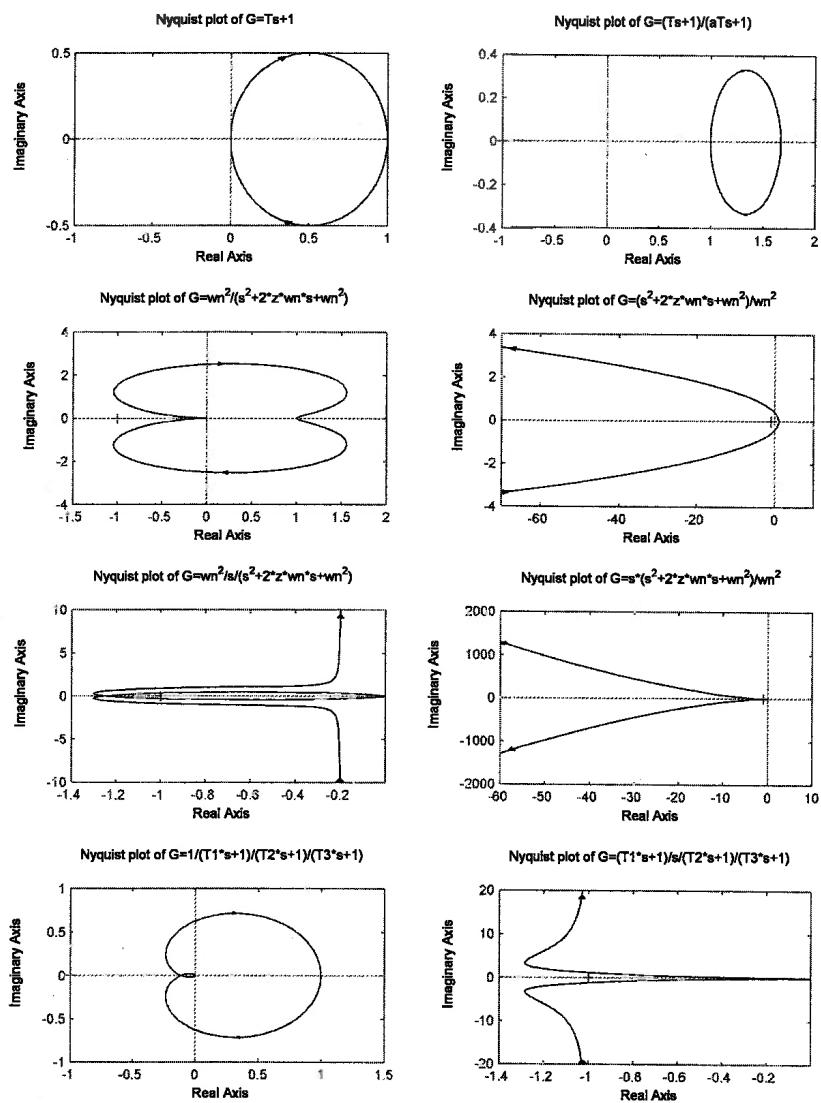
$$G(i0) = \frac{1}{i\infty} , \quad \angle G(i0) = -90^\circ$$

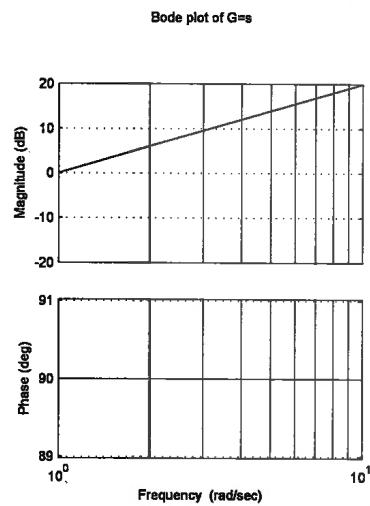
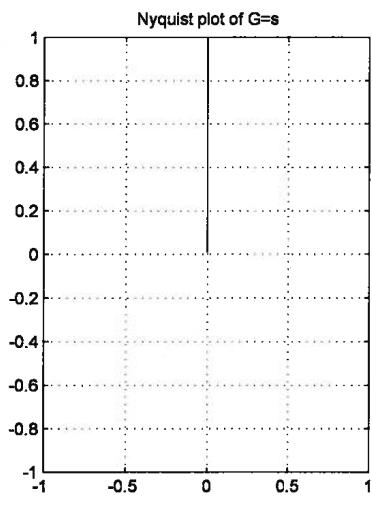
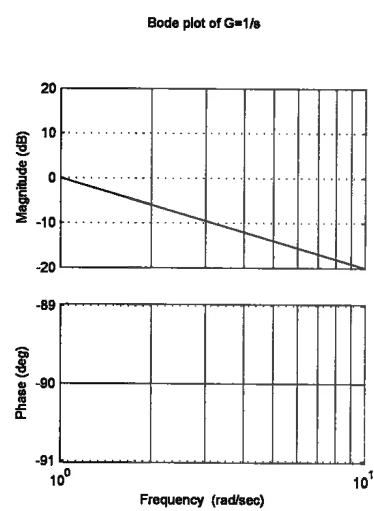
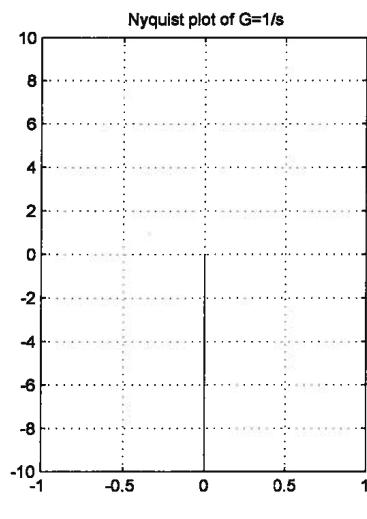


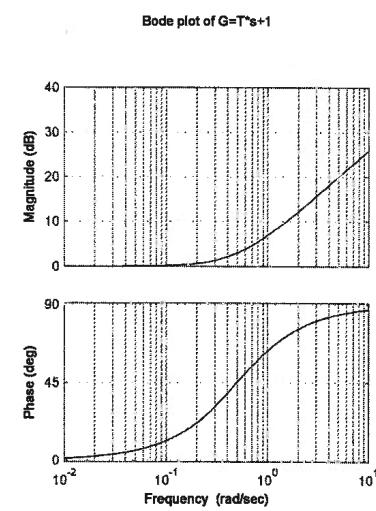
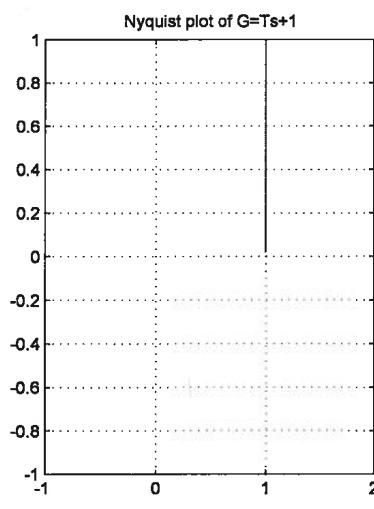
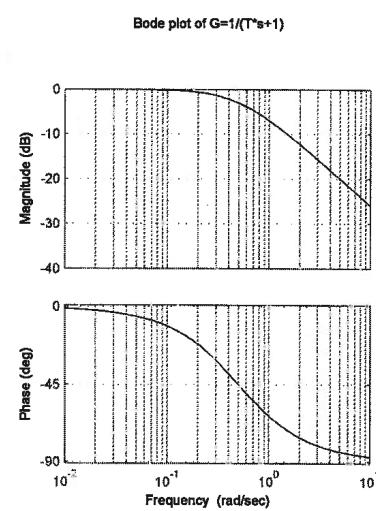
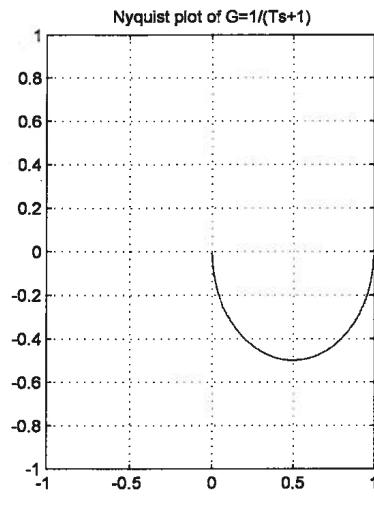


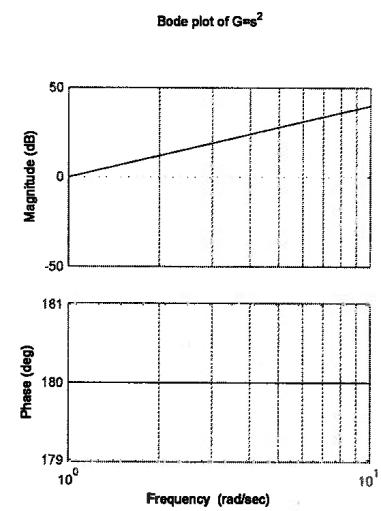
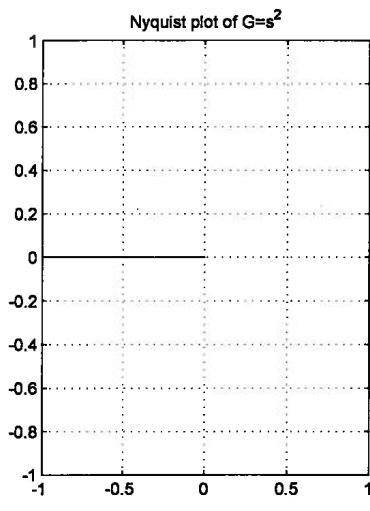
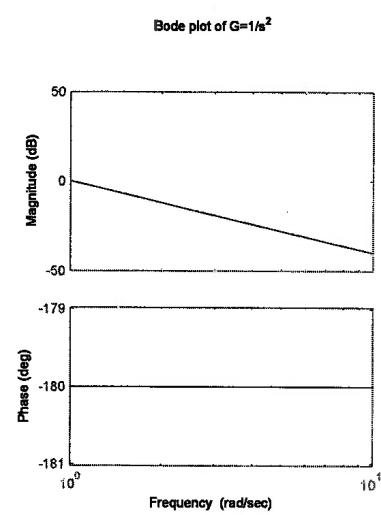
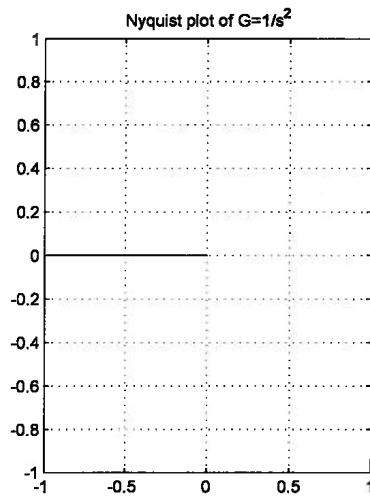


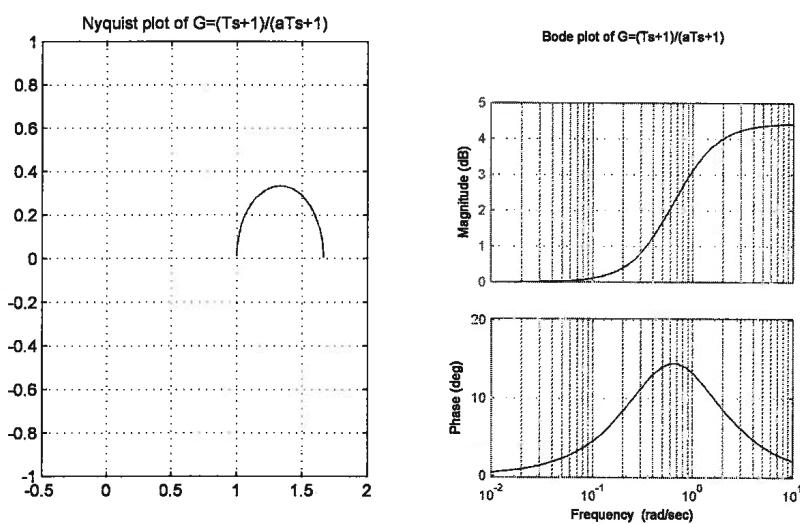
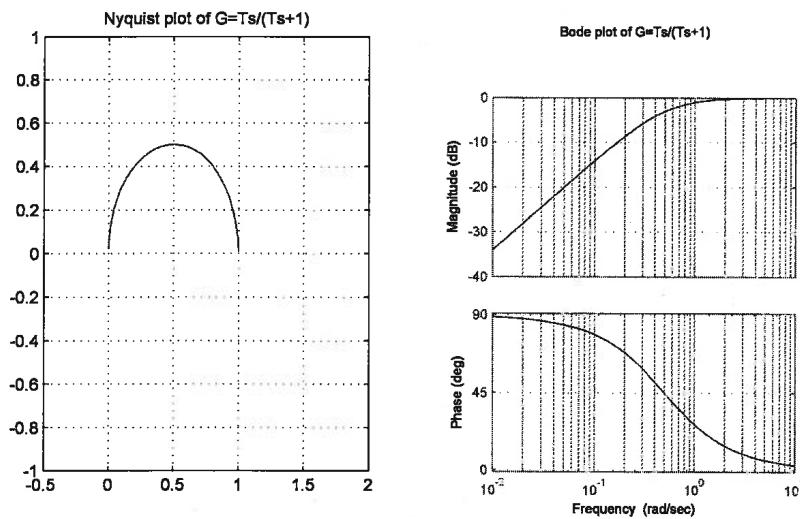


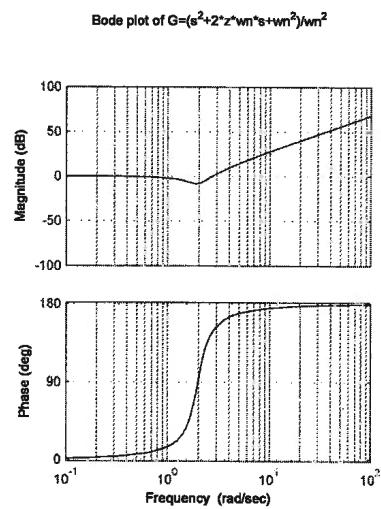
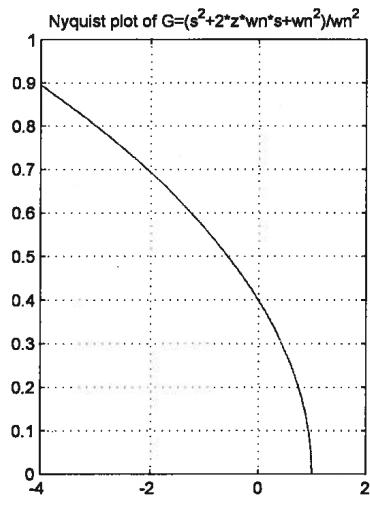
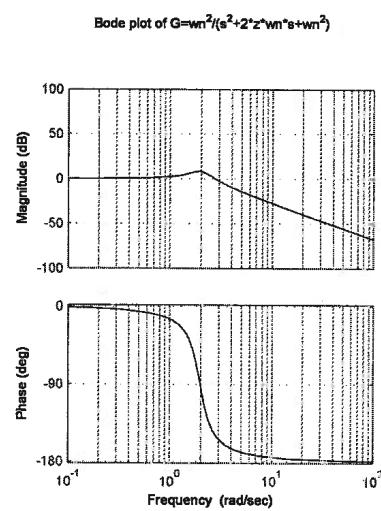
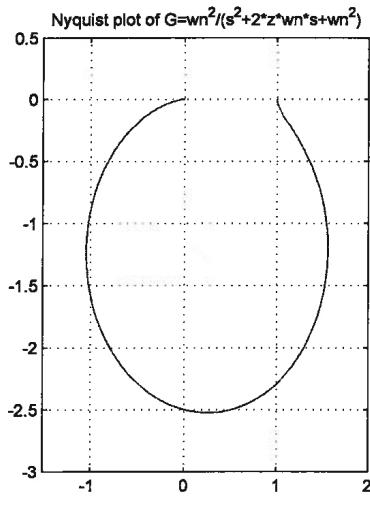


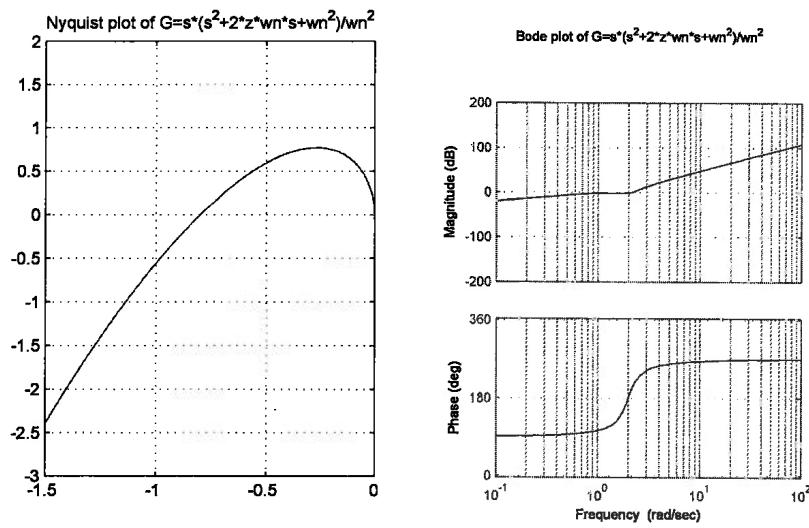
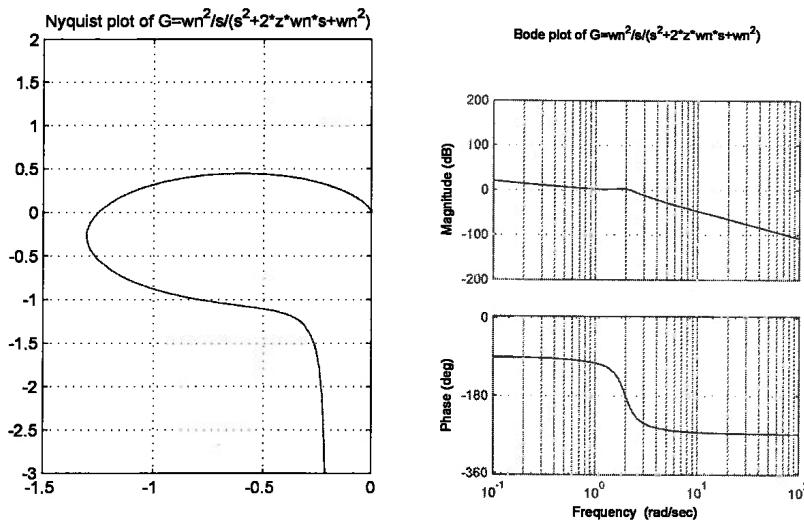


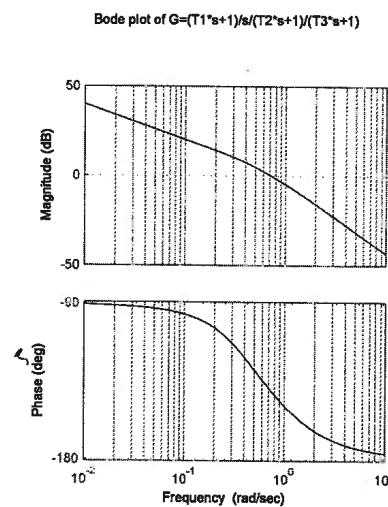
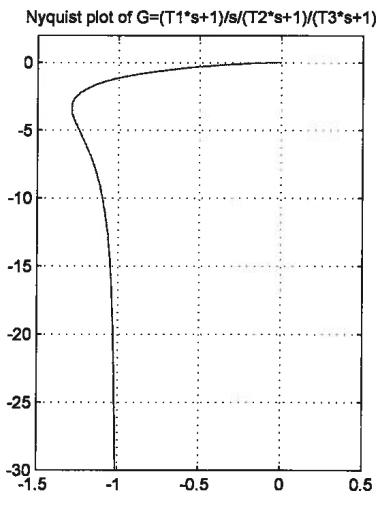
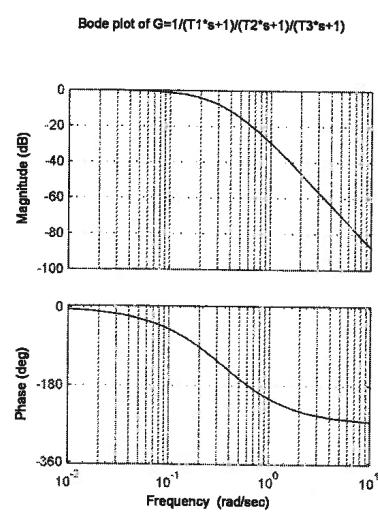
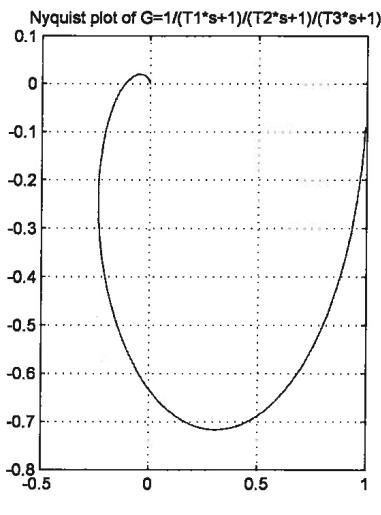












## 8.6 Stability Margins

1  
M

MIL-DTL-9490E defines margin requirements for aircraft flight control systems (FCS) with feedback:

$GM$  = gain margin

$PM$  = phase margin

TABLE III. Gain and phase margin requirements (dB, degrees).

Air speed Mode Frequency Hz	Below $V_{cMIN}$	$V_{cMIN}$ to $V_{cMAX}$	At Limit Airspeed ( $V_L$ )	At $1.15 V_L$
$f_M < 0.06$		$GM = \pm 4.5$ $PM = \pm 30$	$GM = \pm 3.0$ $PM = \pm 20$	
$0.06 \leq f_M <$ First Aero- elastic Mode	$GM = 6$ dB (No Phase Requirement Below $V_{cMIN}$ )	$GM = \pm 6.0$ $PM = \pm 45$	$GM = \pm 4.5$ $PM = \pm 30$	$GM=0$ $PM=0$ (Stable at Nominal Phase and Gain)
$f_M >$ First Aero- Elastic Mode		$GM = \pm 8.0$ $PM = \pm 60$	$GM = \pm 6.0$ $PM = \pm 45$	

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#### MIL-DTL-9490E

Where:

$V_L$	= Limit airspeed (MIL-A-8860).
$V_{oMIN}$	= Minimum operational airspeed (MIL-F-8785)
$V_{oMAX}$	= Maximum operational airspeed (MIL-F-8785)
Mode	= A characteristics aeroelastic response of the aircraft as described by the an aeroelastic characteristic root of the coupled aircraft/FCS dynamic equation of motion.
GM (Gain Margin)	= The minimum chain in loop gain at normal phase, which results in instability beyond that allowed as a residual oscillation.
PM (Phase Margin)	= The minimum change in phase at normal loop gain which results in instability.
$f_M$	= Mode frequency in Hz
Nominal Phase and Gain	= The contractor's best estimate or measurement of FCS and aircraft phase and gain characteristics available at the time of requirement verification.

3.1.3.6.2 Sensitivity analysis. Tolerances on feedback gain and phase shall be established at the system level based on the anticipated range of gain and phase errors which will exist between nominal test values or predictions and in-service operation due to such factors as poorly defined nonlinear and higher order dynamics, anticipated manufacturing tolerances, aging, wear, maintenance and noncritical material failures. Gain and phase margins shall be defined, based on these tolerances, which will assure satisfactory operation in fleet usage. These gain and phase tolerances shall be established based on variations in system characteristics either anticipated or allowed by component or subsystem specification. The contractor shall establish, with the approval of the procuring agency, the range of variation to be considered based on a selected probability of exceedance for each type of variation. The contractor shall select the exceedance probability based on the criticality of the flight control function being provided. The stability requirements established through this sensitivity analysis shall not be less than 50 percent of the magnitude and phase requirements of 3.1.3.6.1.

3.1.3.7 Operation in turbulence. In Operational State I, while flying in the following applicable random and discrete turbulence environment, the FCS shall provide a safe level of operation and maintain mission accomplishment capabilities. For essential and flight phase

Downloaded from <http://www.everspec.com>

## MIL-DTL-9490E

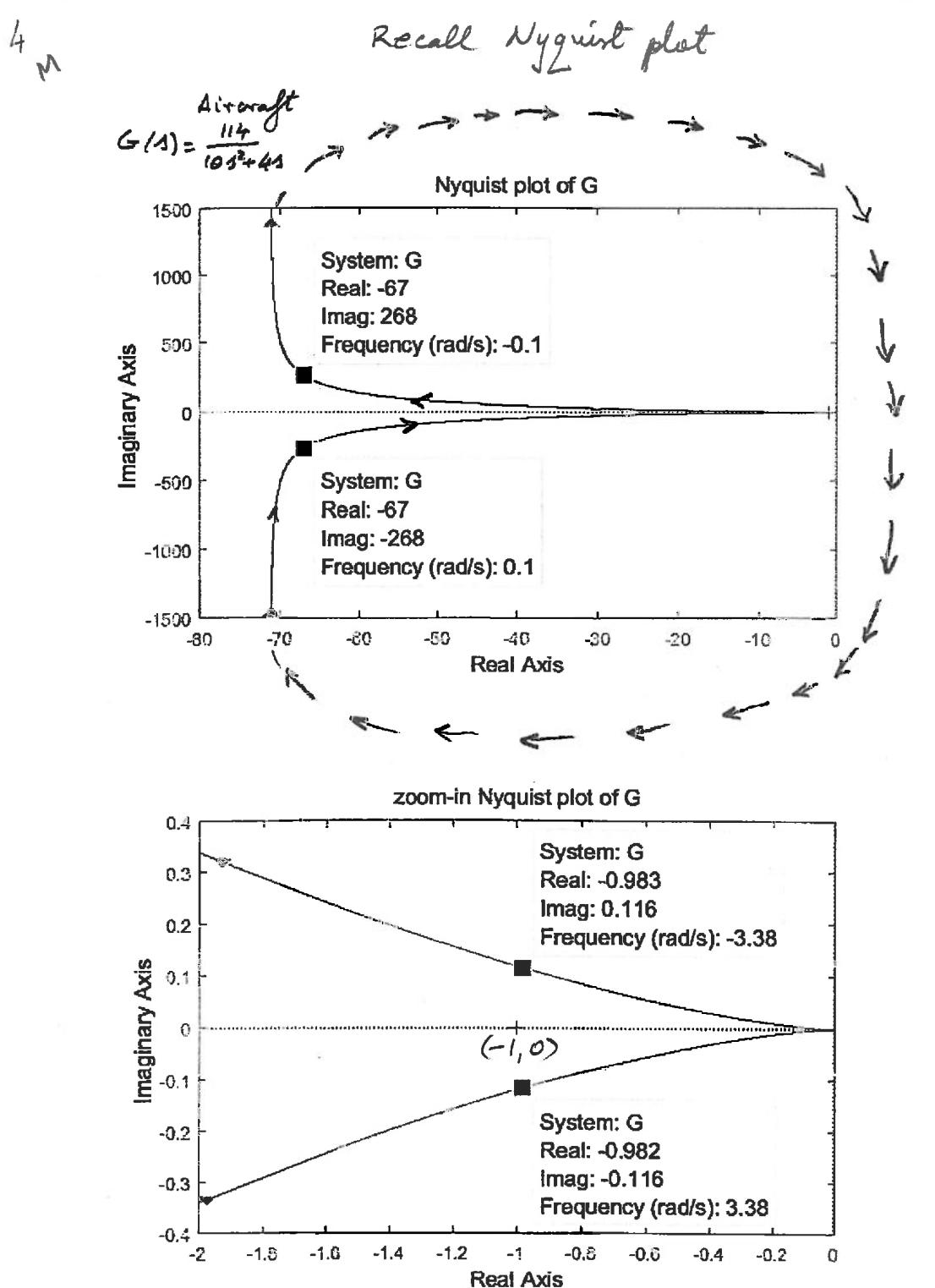
commands by two aircrew members from causing any operation in opposing directions at the same time.

**3.1.3.6 Stability.** For FCS using feedback systems, the stability as specified in 3.1.3.6.1 shall be provided. Alternatively, when approved by the procuring activity, the stability defined by the contractor through the sensitivity analyses of 3.1.3.6.2 shall be provided. Where analysis is used to demonstrate compliance with these stability requirements, the effects of major system nonlinearities shall be included.

**3.1.3.6.1 Stability margins.** Required gain and phase margins about nominal are specified in table III for all aerodynamically closed loop FCS. With these gain or phase variations included, no oscillatory instabilities shall exist with amplitudes greater than those allowed for residual oscillations in 3.1.3.8, and any non oscillatory divergence of the aircraft shall remain within the applicable limits of MIL-F-8785 or MIL-F-83300. AFCS loops shall be stable with these gain or phase variations included for any amplitudes greater than those allowed for residual oscillations in 3.1.3.8. In multiple loop systems, variations shall be made with all gain and phase values in the feedback paths held at nominal values except for the path under investigation. A path is defined to include those elements connecting a sensor to a force or moment producer. For both aerodynamic and nonaerodynamic closed loops, at least 6 dB gain margin shall exist at zero airspeed. At the end of system wear tests, at least 4.5 dB gain margin shall exist for all loops at zero airspeed. The margins specified by table III shall be maintained under flight conditions of most adverse center-of-gravity, mass distribution, and external store configuration throughout the operational envelope and during ground operations.

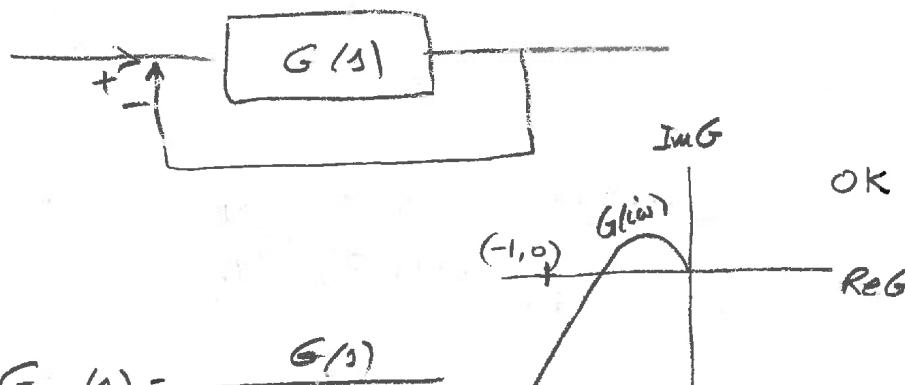
TABLE III. Gain and phase margin requirements (dB, degrees).

Air speed Mode Frequency Hz	Below $V_{oMIN}$	$V_{oMIN}$ to $V_{oMAX}$	At Limit Airspeed ( $V_L$ )	At $1.15 V_L$
$f_M < 0.06$	$GM = 6 \text{ dB}$ (No Phase Requirement Below $V_{oMIN}$ )	$GM = \pm 4.5$ $PM = \pm 30$	$GM = \pm 3.0$ $PM = \pm 20$	$GM=0$ $PM=0$ (Stable at Nominal Phase and Gain)
$0.06 \leq f_M < \text{FirstAero-elasticMode}$		$GM = \pm 6.0$ $PM = \pm 45$	$GM = \pm 4.5$ $PM = \pm 30$	
$f_M > \text{First Aero-Elastic Mode}$		$GM = \pm 8.0$ $PM = \pm 60$	$GM = \pm 6.0$ $PM = \pm 45$	



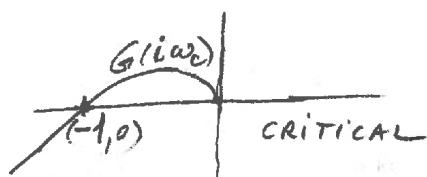
5  
MMargins analysis

- Gain margin
- Phase margin



$$G_{CL}(s) = \frac{G(s)}{1 + G(s)}$$

$$G_{CL}(i\omega) \xrightarrow{\text{if } 1 + G(i\omega) \rightarrow 0} \infty$$

Critical condition  $\omega_c$ 

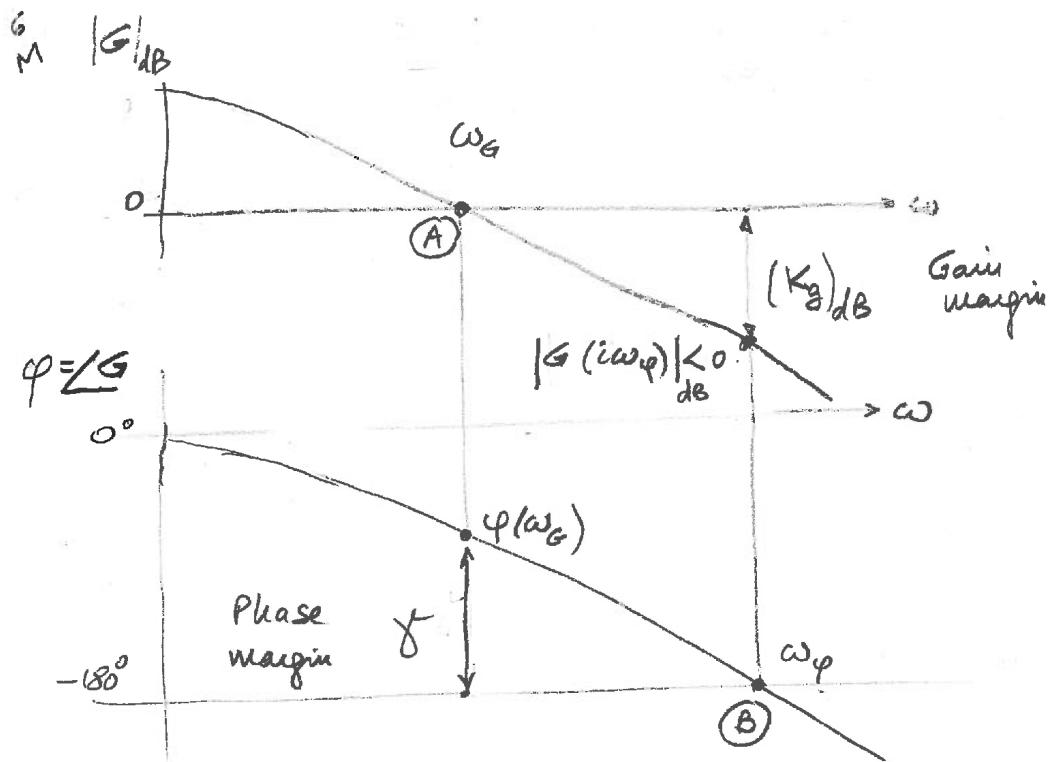
$$\boxed{G(i\omega_c) = -1} \quad \left| \begin{array}{l} |G(i\omega_c)| = 0 \text{ dB} \\ \angle G(i\omega_c) = -180^\circ \end{array} \right.$$

Objective : stay away from  $(-1, 0)$  point!Method

(1) Plot Bode diagram

(2) Determine margins:

- Gain margin : distance from 0dB line
- Phase margin : distance from  $-180^\circ$  line

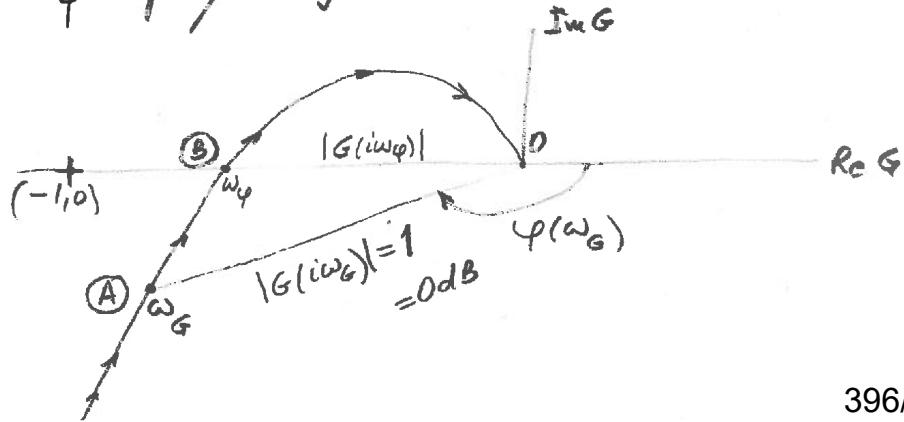


$$\text{Gain margin } (K_g)_{dB} = 0 - |G(i\omega_\varphi)| = -|G(i\omega_\varphi)| > 0$$

$$\text{Phase margin } \gamma = \angle G(i\omega_G) - (-180^\circ) = 180^\circ + \angle G(i\omega_G) > 0$$

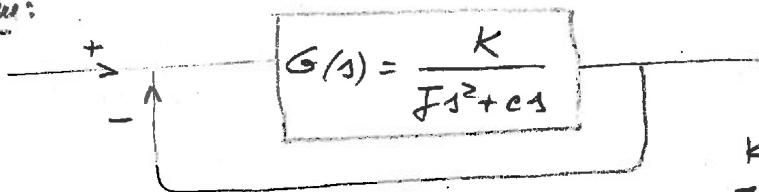
$\omega_G$  = frequency when  $|G(i\omega_G)| = 0$  dB

$\omega_\varphi$  = frequency when  $\varphi = -180^\circ$



19  
21

Example : aircraft roll model

Given:

$$K = 114$$

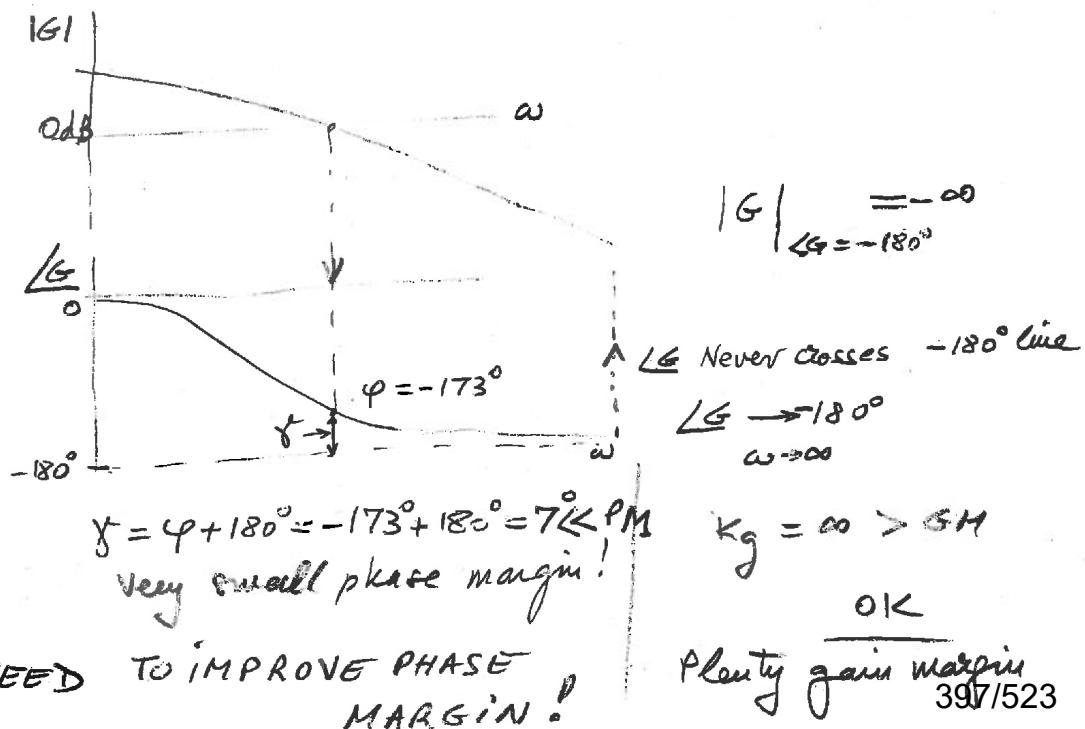
$$J = 10$$

$$c = 4$$

Find:

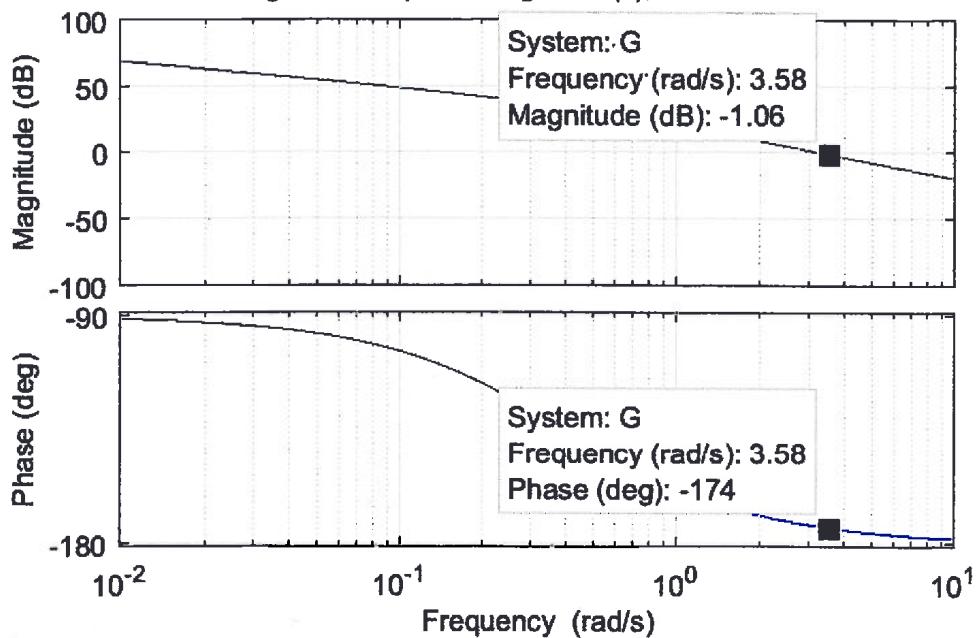
- gain margin,  $(K_g)$  dB

- phase margin  $\gamma$  deg

Solution : Nichols plot (see MATLAB plot) (Fig. 1)

Aircraft roll model  $G(s) = \frac{114}{10s^2 + 4s}$

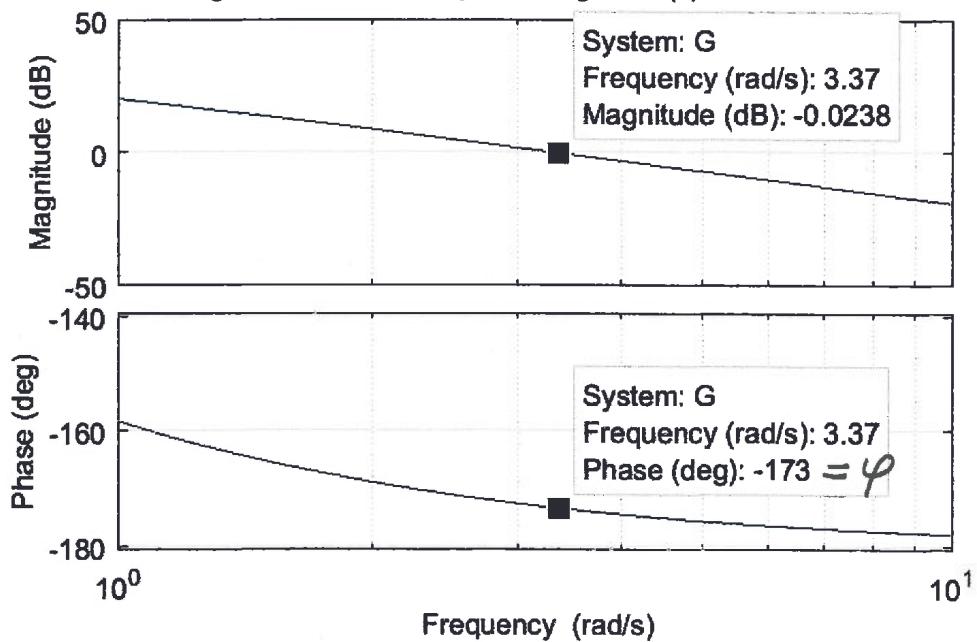
Fig.1a Bode plot of original G(s); aircraft model



$$\text{Phase margin } \gamma^\circ = 180^\circ - 173^\circ = 7^\circ \quad \left. \begin{array}{l} \text{Gain margin (K}_g\text{)} \\ \text{dB} = \infty \end{array} \right\}$$

$$= 180^\circ - \varphi$$

Fig.1b zoom-in Bode plot of original G(s); aircraft model



8  
21

Phase margin: aircraft model  
input data

K | J | c =

114 10 4

G =

$$\frac{114}{10 s^2 + 4 s}$$

Continuous-time transfer function.

---

GM, dB | PM, deg =

10 60

---

phi = phase at |G|=0 dB point, deg =

-173

---

gamma = phase margin, deg =

7

---

## margin

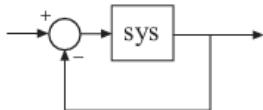
Gain margin, phase margin, and crossover frequencies

### Syntax

```
[Gm,Pm,Wgm,Wpm] = margin(sys)
[Gm,Pm,Wgm,Wpm] = margin(mag,phase,w)
margin(sys)
```

### Description

`margin` calculates the minimum gain margin,  $G_m$ , phase margin,  $P_m$ , and associated frequencies  $W_{gm}$  and  $W_{pm}$  of SISO open-loop models. The gain and phase margin of a system `sys` indicates the relative stability of the closed-loop system formed by applying unit negative feedback to `sys`, as in the following illustration.



The gain margin is the amount of gain increase or decrease required to make the loop gain unity at the frequency  $W_{gm}$  where the phase angle is  $-180^\circ$  (modulo  $360^\circ$ ). In other words, the gain margin is  $1/g$  if  $g$  is the gain at the  $-180^\circ$  phase frequency. Similarly, the phase margin is the difference between the phase of the response and  $-180^\circ$  when the loop gain is 1.0. The frequency  $W_{pm}$  at which the magnitude is 1.0 is called the *unity-gain frequency* or *gain crossover frequency*. It is generally found that gain margins of three or more combined with phase margins between 30 and 60 degrees result in reasonable trade-offs between bandwidth and stability.

`[Gm,Pm,Wgm,Wpm] = margin(sys)` computes the gain margin  $G_m$ , the phase margin  $P_m$ , and the corresponding frequencies  $W_{gm}$  and  $W_{pm}$ , given the SISO open-loop dynamic system model `sys`.  $W_{gm}$  is the frequency where the gain margin is measured, which is a  $-180^\circ$  degree phase crossing frequency.  $W_{pm}$  is the frequency where the phase margin is measured, which is a 0dB gain crossing frequency. These frequencies are expressed in radians/TimeUnit, where `TimeUnit` is the unit specified in the `TimeUnit` property of `sys`. When `sys` has several crossovers, `margin` returns the smallest gain and phase margins and corresponding frequencies.

The phase margin  $P_m$  is in degrees. The gain margin  $G_m$  is an absolute magnitude. You can compute the gain margin in dB by

$$Gm\_dB = 20 * \log10(Gm)$$

`[Gm,Pm,Wgm,Wpm] = margin(mag,phase,w)` derives the gain and phase margins from Bode frequency response data (magnitude, phase, and frequency vector). `margin` interpolates between the frequency points to estimate the margin values. Provide the gain data `mag` in absolute units, and phase data `phase` in degrees. You can provide the frequency vector `w` in any units; `margin` returns  $W_{gm}$  and  $W_{pm}$  in the same units.

#### Note

When you use `margin(mag,phase,w)`, `margin` relies on interpolation to approximate the margins, which generally produces less accurate results. For example, if there is no 0 dB crossing within the `w` range, `margin` returns a phase margin of `Inf`. Therefore, if you have an analytical model `sys`, using `[Gm,Pm,Wgm,Wpm] = margin(sys)` is the most robust way to obtain the margins.

`margin(sys)`, without output arguments, plots the Bode response of `sys` on the screen and indicates the gain and

400/523

phase margins on the plot. By default, gain margins are expressed in dB on the plot.

## Examples

### Gain and Phase Margins of Open-Loop Transfer Function

Create an open-loop discrete-time transfer function.

```
hd = tf([0.04798 0.0464],[1 -1.81 0.9048],0.1)
```

```
hd =
```

```
0.04798 z + 0.0464  
-----  
z^2 - 1.81 z + 0.9048
```

Sample time: 0.1 seconds

Discrete-time transfer function.

Compute the gain and phase margins.

```
[Gm,Pm,Wgm,Wpm] = margin(hd)
```

```
Gm =
```

```
2.0517
```

```
Pm =
```

```
13.5711
```

```
Wgm =
```

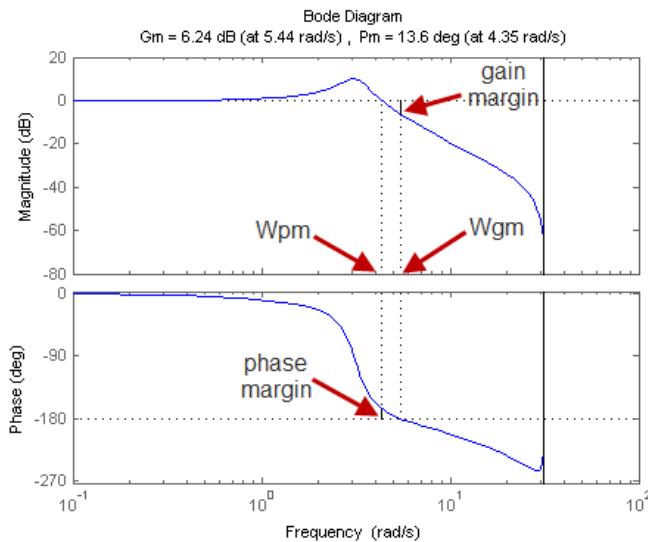
```
5.4374
```

```
Wpm =
```

```
4.3544
```

Display the gain and phase margins graphically.

```
margin(hd)
```



Solid vertical lines mark the gain margin and phase margin. The dashed vertical lines indicate the locations of  $W_{pm}$ , the frequency where the phase margin is measured, and  $W_{gm}$ , the frequency where the gain margin is measured.

## Algorithms

The phase margin is computed using  $H_\infty$  theory, and the gain margin by solving  $H(j\omega) = \overline{H(j\omega)}$  for the frequency  $\omega$ .

## See Also

[Linear System Analyzer](#) | [bode](#)

---

Introduced before R2006a

---

```

margins_aircraft.m × +
1 %% initialization
2 clc %clear command window
3 clear %removes all variables from workspace; release memory
4 format compact
5 close all %closes all figures
6 s=tf('s');
7 %% original aircraft model
8 display('Phase margin: aircraft model')
9 K=114; % gain
10 J=10; % inertia
11 c=4; % damping
12 display('input data')
13 display([K J c], ' K | J | c')
14 G=K/(J*s^2+c*s) % G(s)
15 figure(1)
16 subplot(2,1,1)
17 bode(G)
18 grid
19 title('Fig.1a Bode plot of aircraft model')
20 subplot(2,1,2)
21 d1=0;d2=1;N=1e3; w=logspace(d1,d2,N);
22 bode(G,w)
23 grid
24 title('Fig.1b zoom-in Bode plot of aircraft model')
25 % READ ON PLOT: phase at |G|=0 dB point, deg
26 % phi=-173;
27 phi=input('Input phase read on Bode plot in deg, phi=');
28 gamma=phi-(-180); % gamma = phase margin, deg
29 display([gamma],'gamma = phase margin, deg')
30 figure(2)
31 margin(G)

```

```

Phase margin: aircraft model
input data
    K   |   J   |   c =
    114     10      4

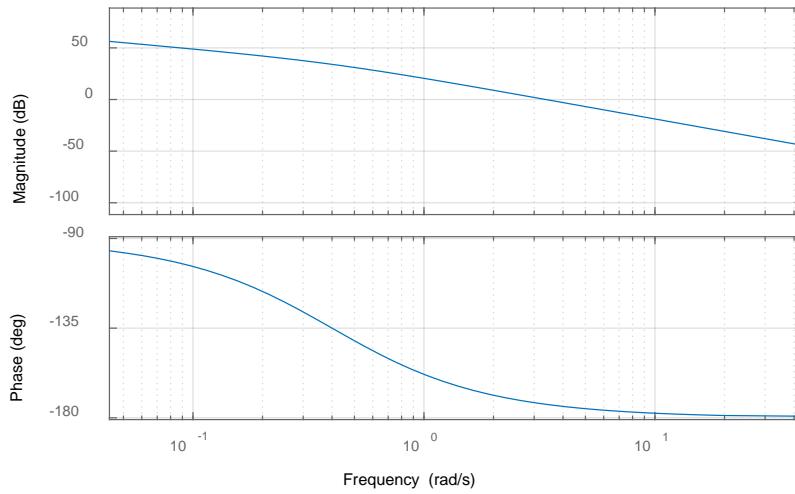
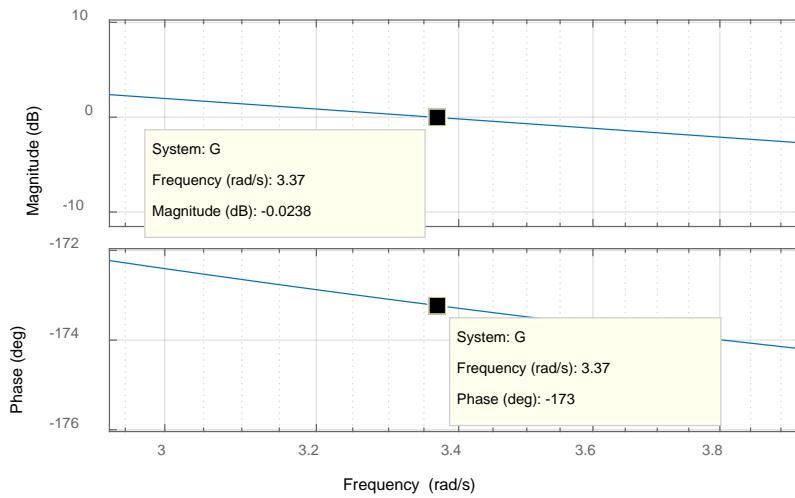
G =

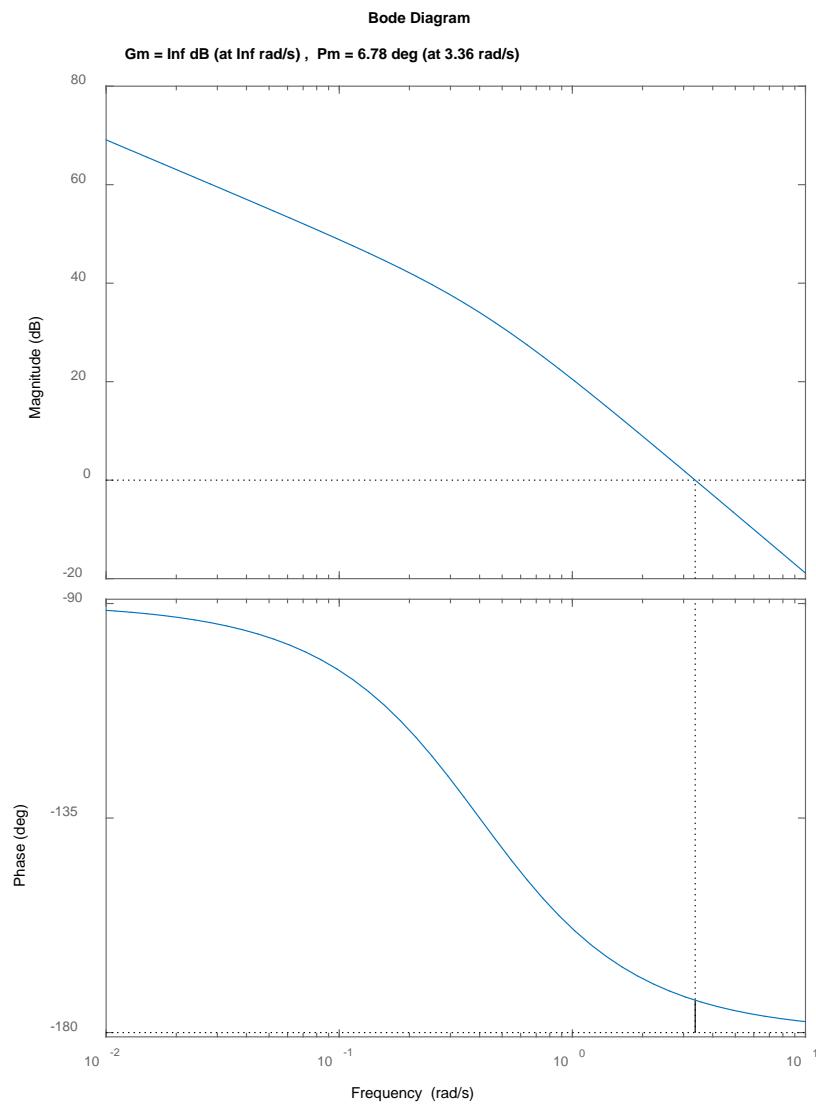
    114
    -----
    10 s^2 + 4 s

Continuous-time transfer function.

Input phase read on Bode plot in deg, phi=-173
gamma = phase margin, deg =
    7

```

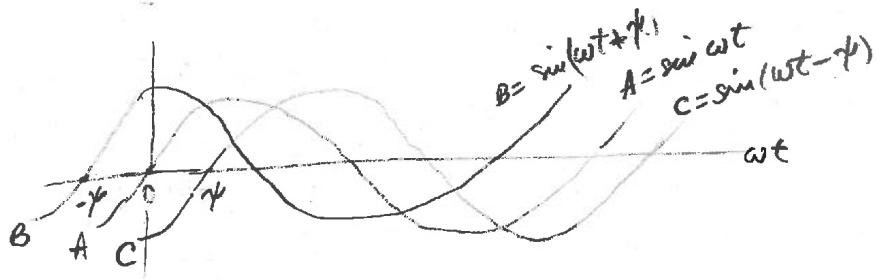
**Fig.1a** Bode plot of aircraft model**Fig.1b** zoom-in Bode plot of aircraft model



## 8.7 Phase Compensators

## C1 PHASE COMPENSATORS

Recall signal phase definition:



$$A: \sin(\omega t) = 0 \quad @ t=0$$

B:  $\sin(\omega t + \phi) = 0 \quad @ \omega t = -\phi$  "ahead of time"  
it leads

C:  $\sin(\omega t - \phi) = 0 \quad @ \omega t = \phi$  "delayed"  
it lags

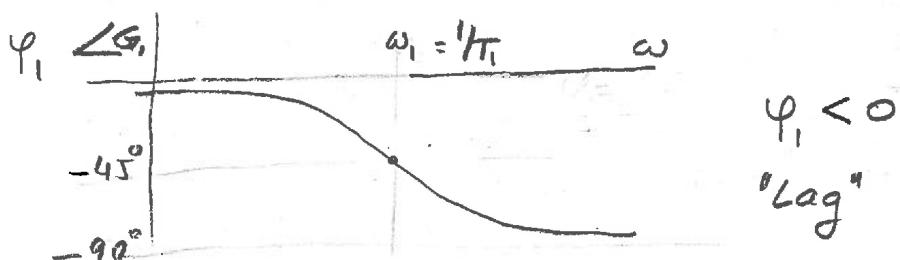
### Phase compensator

$$G_c(i\omega) = |G_c(i\omega)| e^{i\varphi_c}, \quad \varphi_c = \angle G_c(i\omega)$$

if  $\varphi_c > 0$ , then "lead compensator"

$\varphi_c < 0$  "lag compensator"

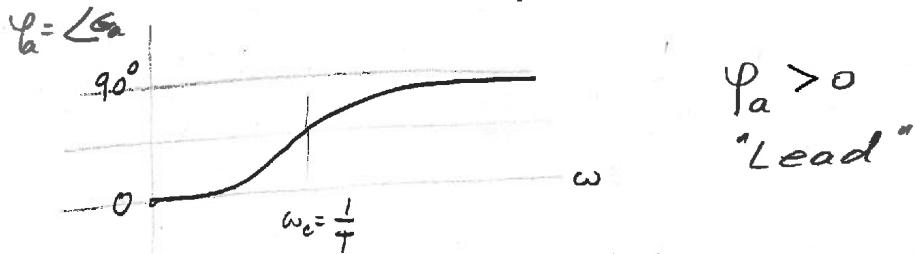
Example 1: 1<sup>st</sup> order system  $G_1(s) = \frac{1}{sT_1 + 1}$



C2

$$\underline{\text{Example 2}} : \quad G_a = sT_a + 1$$

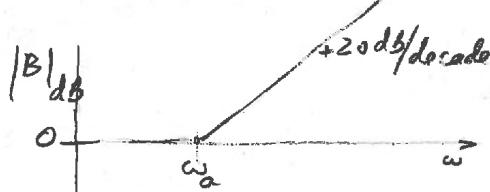
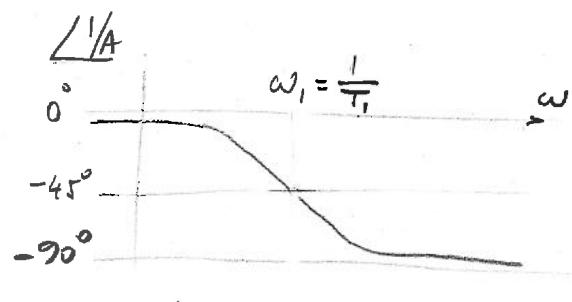
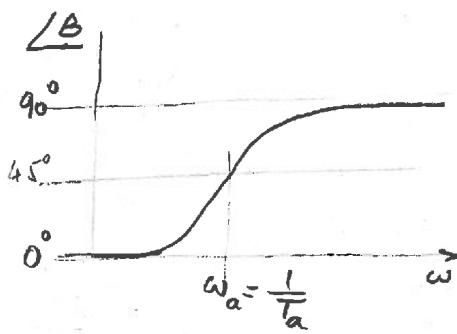
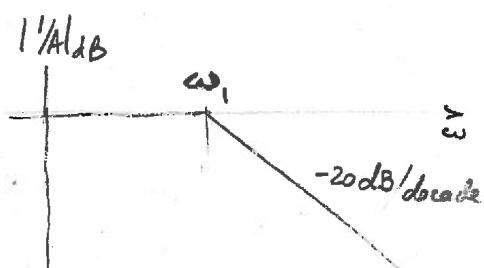
$$G_a(i\omega) = i\omega T_a + 1; \quad \varphi_a = \angle G_a = \tan^{-1} \omega T_a > 0$$



$$\underline{\text{Example 3}} \quad G_c = \frac{sT_a + 1}{sT_1 + 1} = \frac{B(s)}{A(s)}$$

$$G_c(i\omega) = \frac{i\omega T_a + 1}{i\omega T_1 + 1} = \frac{B(i\omega)}{A(i\omega)}$$

$$|G_c|_{dB} = |B|_{dB} - |A|_{dB} \quad ; \quad \angle G_c = \angle B - \angle A$$

B(iω): NUMERATORA(iω): DENOMINATOR

'Lag'

C2a

LEAD COMPENSATOR  
FOR PHASE MARGIN IMPROVEMENT

409/523

C3

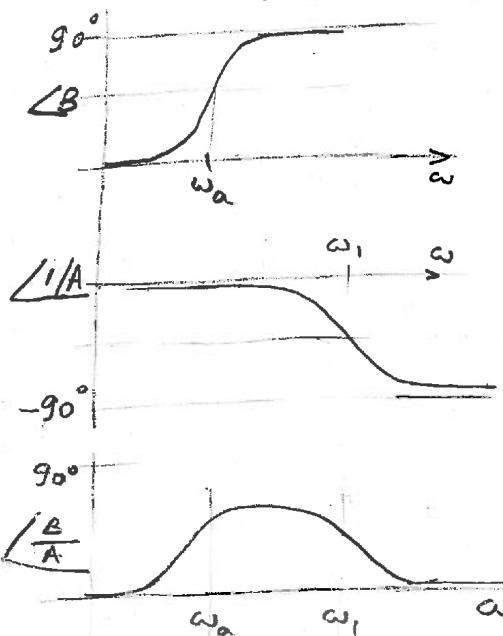
LEAD COMPENSATOR

$$\omega_a < \omega_1 \Rightarrow T_1 < T_a$$

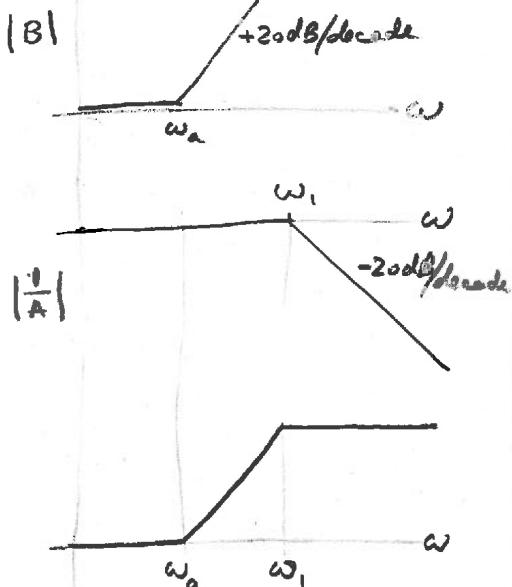
$$G_c = \frac{sT_a + 1}{sT_1 + 1}$$

$$\omega_a = \frac{1}{T_a} \Rightarrow \omega_1 = \frac{1}{T_1}$$

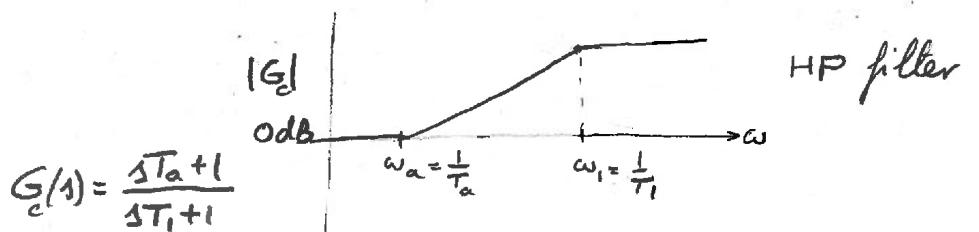
-Phase-



—Gain—

"phase lead" between  $\omega_a$  &  $\omega_1$ 

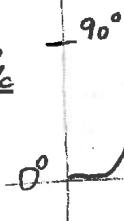
High pass filter  
(higher frequencies are amplified)



$$G(1) = \frac{sT_a + 1}{sT_1 + 1}$$

$$T_1 < T_a$$

$$\angle G_c$$



maximum phase lead

C3a

Lead Compensator Design

Use lead compensator to move phase plot upward  
and improve phase margin

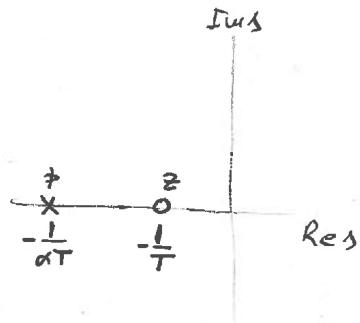
$$G_c = \frac{sT_a + 1}{sT_1 + 1} = \frac{sT + 1}{s\alpha T + 1} \quad \text{lead compensator (1)}$$

$\omega_a < \omega_1, \quad 0 < \alpha < 1$   
 $T_a > T_1$

Poles & zeros :

$$sT + 1 = 0 \rightarrow z = -\frac{1}{T}$$

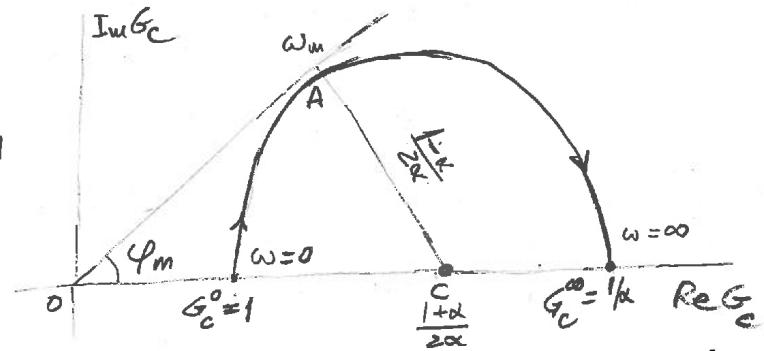
$$s\alpha T + 1 = 0 \rightarrow p = -\frac{1}{\alpha T}$$

Nyquist plot

$$G_c(i\omega) = \frac{i\omega T + 1}{i\omega\alpha T + 1}$$

$$G_c^0(i0) = 1$$

$$G_c^\infty(i\infty) = \frac{1}{\alpha}$$



$$\sin \varphi_m = \frac{1-\alpha}{1+\alpha}$$

maximum phase lead

$$\varphi_m = \tan^{-1} \frac{1-\alpha}{1+\alpha}$$

(2)

$$\text{Proof : } OC = \frac{1}{2} \left( 1 + \frac{1}{\alpha} \right) = \frac{1+\alpha}{2\alpha}$$

$$\text{Radius} = \frac{1+\alpha}{2\alpha} - 1 = \frac{1-\alpha}{2\alpha} = AC$$

$$\sin \varphi_m = \frac{AC}{OC} = \frac{1-\alpha}{1+\alpha} \quad \text{QED}$$

411/523

C36 Calculation of  $\alpha$   
 $\text{Eq. (2)}$  can be used to determine the value of  $\alpha$  in Eq. (1), i.e.,

$$\sin \varphi_m = \frac{1-\alpha}{1+\alpha}$$

$$\frac{1+\sin \varphi_m}{1-\sin \varphi_m} = \frac{1-\alpha+1+\alpha}{1+\alpha-(1-\alpha)} = \frac{2}{2\alpha} = \frac{1}{\alpha}$$

$$\alpha = \frac{1-\sin \varphi_m}{1+\sin \varphi_m} \quad \text{--- (3)}$$

### Calculation of $T$

- $\omega_m$  is the geometric mean of the corner frequencies

$$\omega_m = \sqrt{\omega_a \omega_1} = \sqrt{\frac{1}{T_a} \cdot \frac{1}{T_1}} \quad \begin{cases} T_a = T \\ T_1 = \alpha T \end{cases} = \sqrt{\frac{1}{T} \cdot \frac{1}{\alpha T}} = \frac{1}{T} \frac{1}{\sqrt{\alpha}} \quad \text{--- (4)}$$

$$\begin{aligned} |G_c(\omega_m)| &= \left| \frac{i\omega_m T + 1}{i\omega_m \times T + 1} \right| = \left| \frac{i \frac{1}{\sqrt{\alpha}} T + 1}{i \alpha \frac{1}{\sqrt{\alpha}} T + 1} \right| = \left| \frac{i + \sqrt{\alpha}}{i\sqrt{\alpha} + 1} \right| \frac{1}{\sqrt{\alpha}} \\ &= \frac{\sqrt{1+\alpha}}{\sqrt{1+\alpha}} \frac{1}{\sqrt{\alpha}} = \frac{1}{\sqrt{\alpha}} \end{aligned}$$

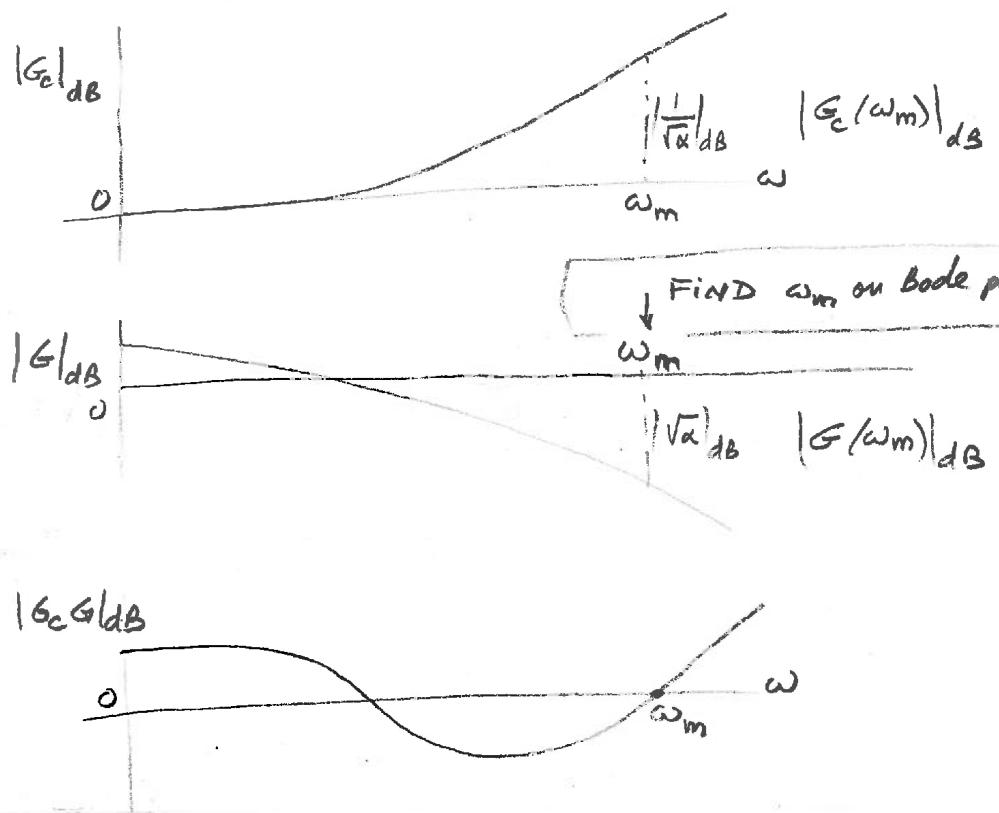
$$|G_c(\omega_n)| = \frac{1}{\sqrt{\alpha}}, \quad 0 < \alpha < 1 \quad \text{--- (5)}$$

The phase compensator will produce a gain  $\frac{1}{\sqrt{\alpha}}$ . To find  $\omega_m$ , examine the Bode diagram of the original system  $G$ , and identify the freq. at which  $|G(\omega_m)| = \sqrt{\alpha}$ ; note that

412/523

C3c

$\omega_m$  is selected at a value at which the gain drop  $|G(\omega_n)|$  of the original system balances the gain addition  $|G_c(\omega_m)|$  due to the compensator (balance condition)



- After determining  $\omega_m$  graphically, calculate  $T$  with Eq.(4) i.e.,

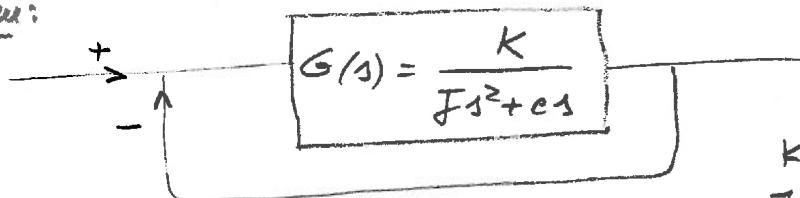
$$T = \frac{1}{\omega_n} \cdot \frac{1}{\sqrt{\alpha}} \quad (6)$$

Thus:

$$G_c = \frac{sT+1}{s\alpha T+1} \quad (7)$$

E

Example : aircraft roll model with lead compns.

Given:

$$K = 114$$

$$J = 10$$

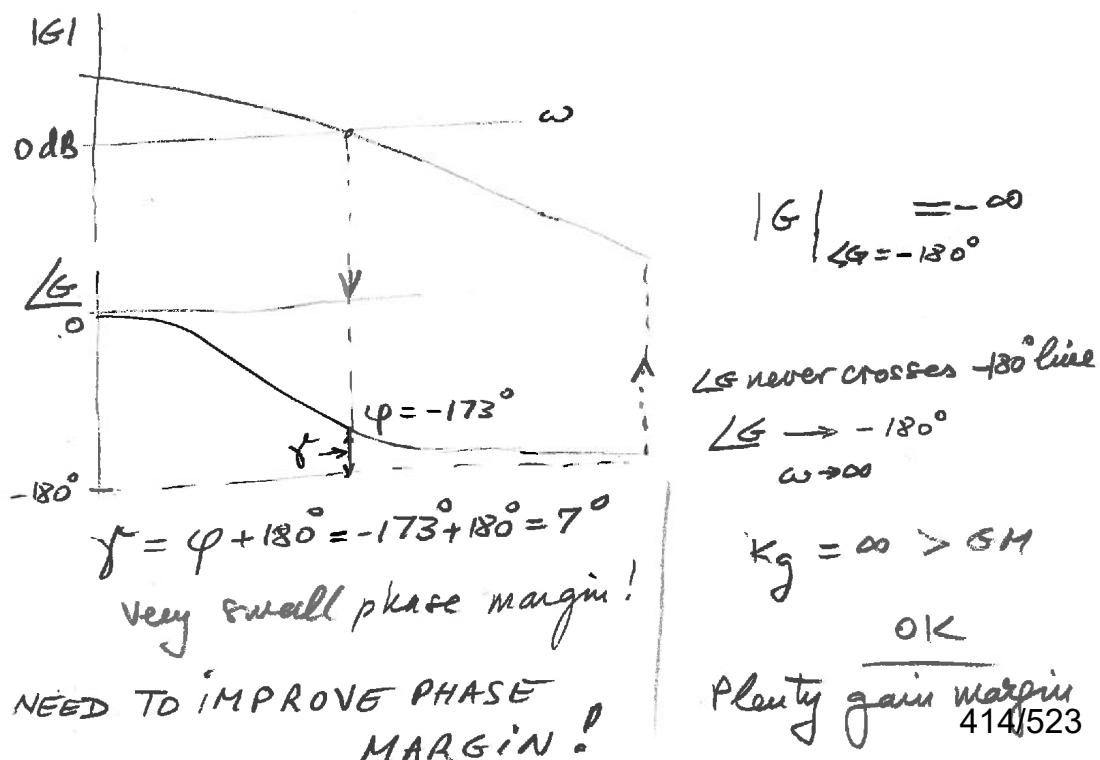
$$c = 4$$

Find(a) - gain margin, ( $K_g$ ) dB- phase margin  $\gamma$  deg

(b) design compensator to achieve

$$GM = 10 \text{ dB} \quad (1)$$

(c) plot time response

Solution (a) Bode plot (see MATLAB plot, Fig. 1)

2/11

Phase margin: aircraft model  
input data

K |  $\omega$  | c =

114 10 4

G =

$$\frac{114}{10 s^2 + 4 s}$$

Continuous-time transfer function.

GM, dB | PM, deg =

10 60

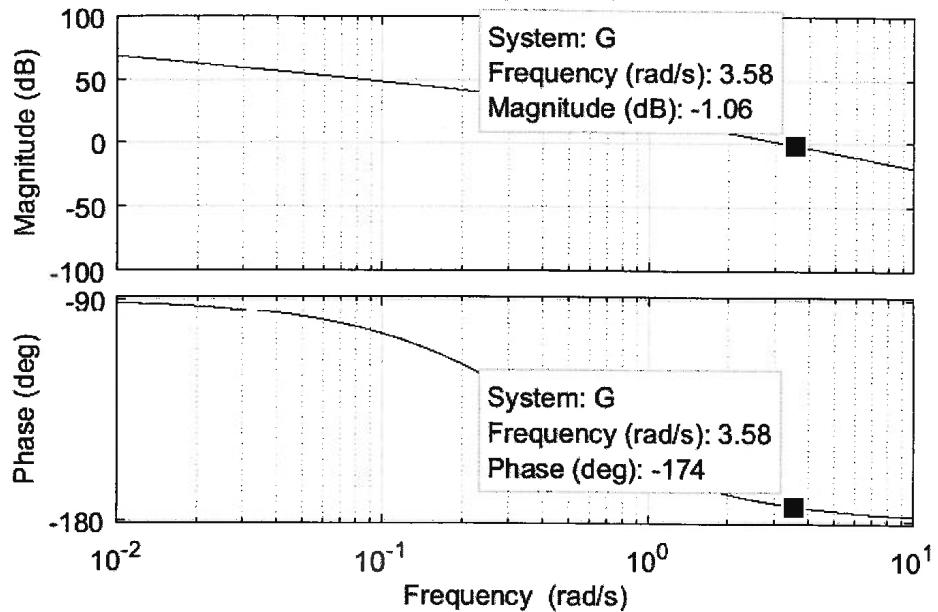
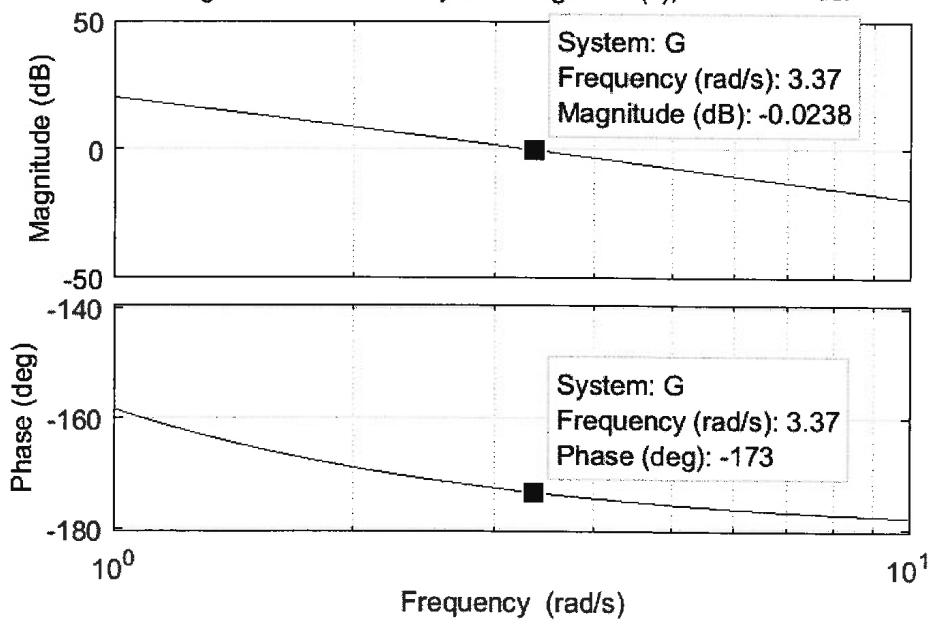
phi = phase at |G|=0 dB point, deg =

-173

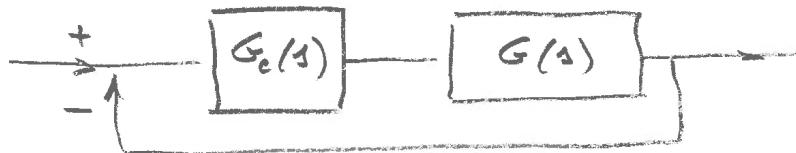
gamma = phase margin, deg =

7

MW

Fig.1a Bode plot of original  $G(s)$ ; aircraft modelFig.1b zoom-in Bode plot of original  $G(s)$ ; aircraft model

<sup>4</sup>E (b) Design phase compensator to improve phase margin



$$G_c(s) = \frac{T_s + 1}{\alpha T_s + 1} \quad \text{lead compensator (2)}$$

(b1) We need to improve phase margin from  $\gamma = 7^\circ$  to  $PM = 60^\circ$ , i.e., the phase compensator must add  $\varphi_m = 53^\circ$ .

- To calculate  $\alpha$ , recall

$$\alpha = \frac{1 - \sin \varphi_m}{1 + \sin \varphi_m} = 0.1120 \quad (3)$$

$\varphi_m = 53^\circ$

- To calculate  $T$ , recall  $T = \frac{1}{\omega_m \sqrt{\alpha}}$

We need  $\omega_n$ . We find  $\omega_m$  from the balance condition, i.e.,

$$|G_c(\omega_m)| = \frac{1}{\sqrt{\alpha}} = 2.9887 = 9.5096 \text{ dB} \quad (4)$$

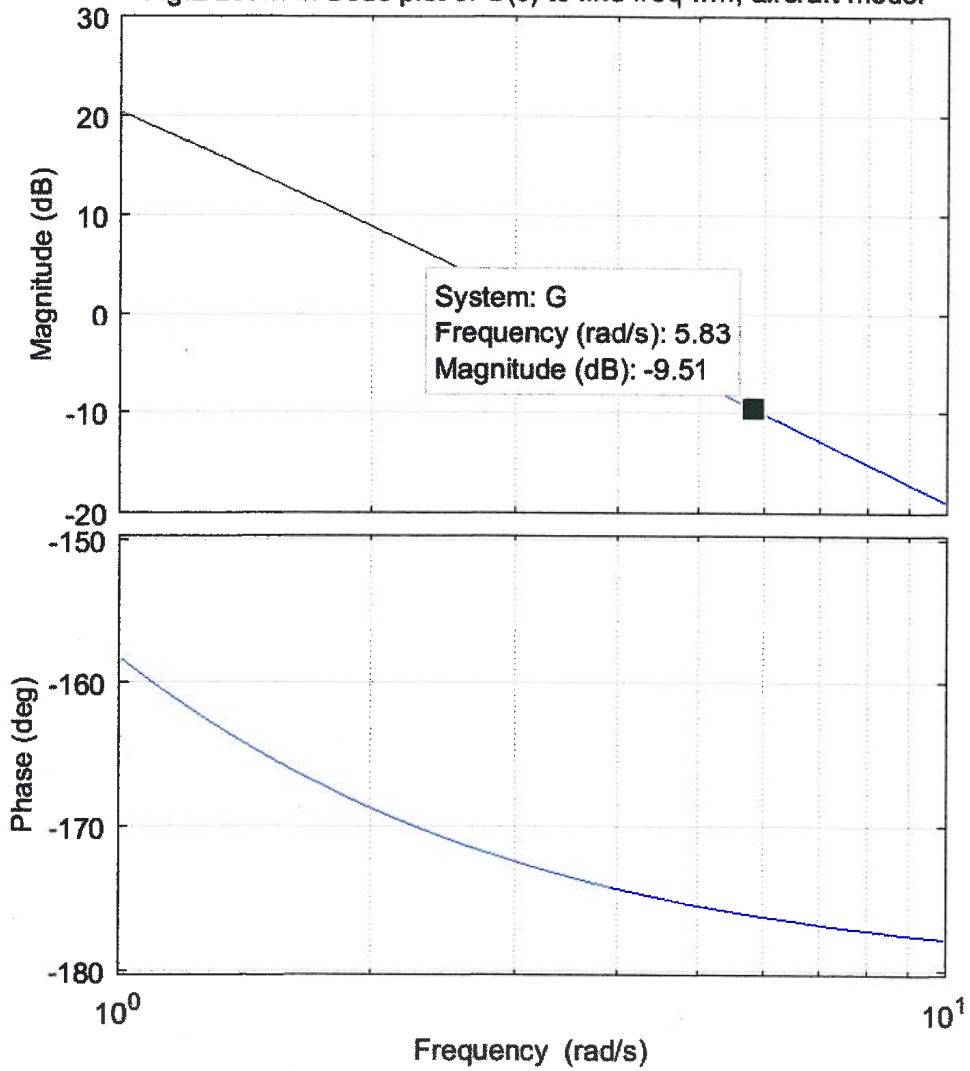
$$|G(\omega_m)| = -|G_c(\omega_m)| = -9.5096 \text{ dB} \quad (5)$$

From Bode plot Fig. 2, we find  $\omega_n = 5.83 \text{ rad/sec}$ .

Hence  $T = 0.5126 \text{ sec}$ .

A<sub>6</sub><sup>17/523</sup>

↙

Fig.2 zoom-in Bode plot of  $G(s)$  to find freq  $\omega_m$ ; aircraft model

$$\omega_m = 5.83 \text{ rad/s}$$

$$|G(\omega_m)|_{dB} = -9.51 \text{ dB.}$$

6 (b2): Test compensated system  $G_c G$

Fig. 3 shows the performance of the compensated system  $G_c G$  in comparison to the original system  $G$ . The compensated system  $G_c G$  has

$\varphi_e = -123^\circ \quad \omega_c = 5.83 \text{ rad/s}$  where  $|G_c G| = 0 \text{ dB}$   
The phase margin of the compensated system is

$$\gamma_c = \varphi_c + 180^\circ = -123^\circ + 180^\circ = 57^\circ < 60^\circ \text{ PM}$$

The system has improved, but the phase margin is still less than PM.

(b3). Need to adjust the compensation to add a little more phase shift,

$$\Delta\varphi = \text{PM} - \gamma_c = 60^\circ - 57^\circ = 3^\circ \quad (7)$$

The new  $\varphi_m$  is

$$\varphi_{m_1} = \varphi_m + \Delta\varphi = 53^\circ + 3^\circ = 56^\circ \quad (8)$$

The new  $\alpha$  is

$$\alpha_1 = \frac{1 - \sin \varphi_{m_1}}{1 + \sin \varphi_{m_1}} = 0.0935 \quad (9)$$

We keep same time constant,  $T_1 = T$  419/523

7  
E

(6)

Phase compensator design: aircraft model

phi\_m =

53

alpha =

0.1120

1/sqrt(alpha) =

2.9887

Gc\_wm, dB =

9.5096

wm =

5.9300

T =

0.5126

Gc =

0.5126 s + 1

-----

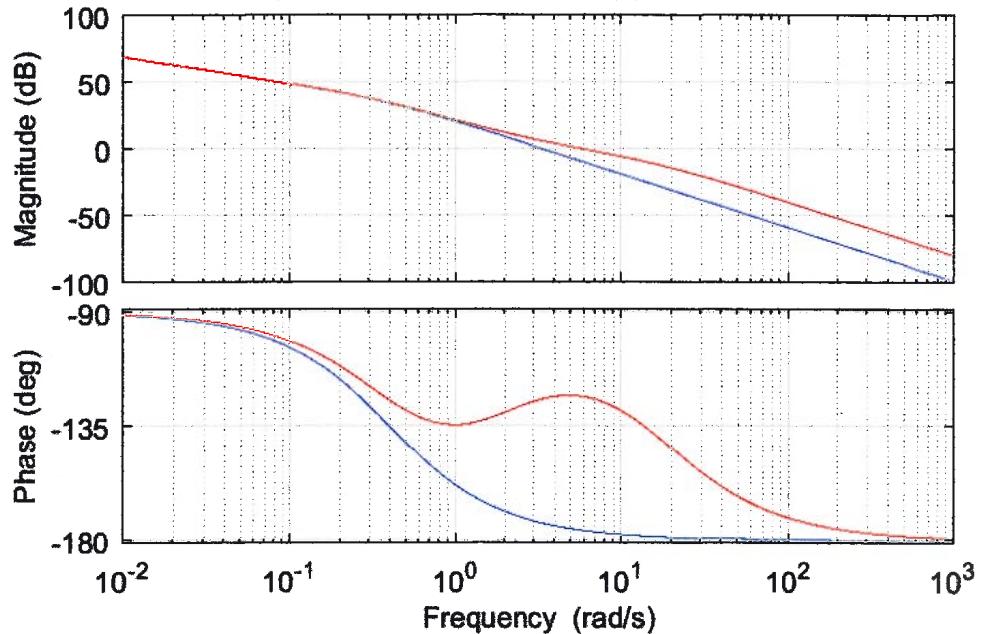
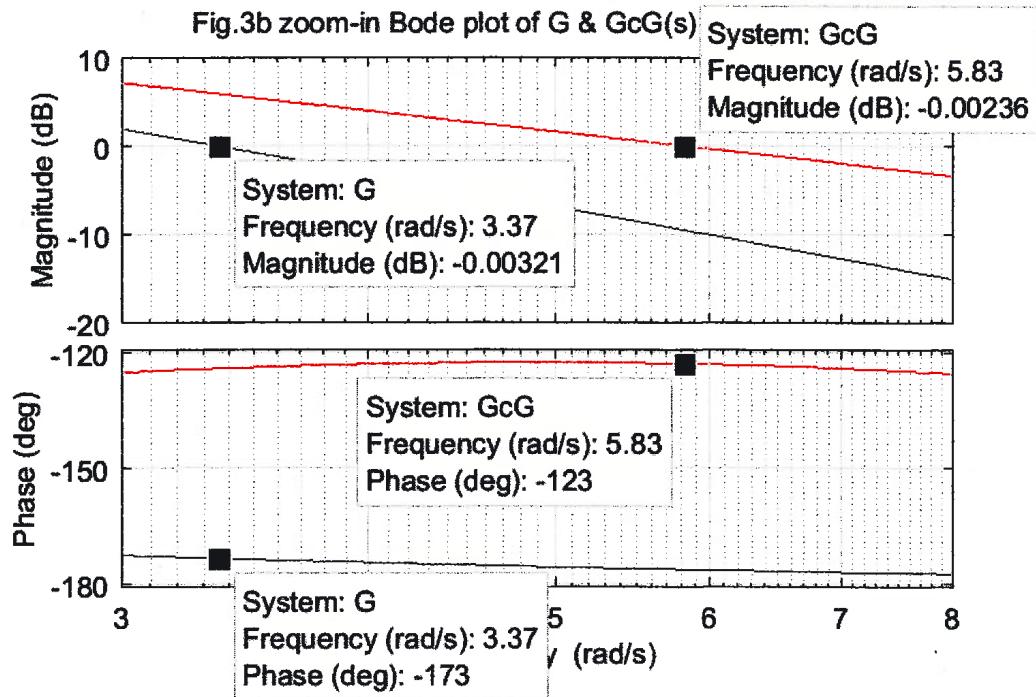
0.05739 s + 1

Continuous-time transfer function.

phi\_c = phase at  $|G_c G| = 0$  dB point, deg =  
-123

gamma\_c = phase margin of  $G_c G$ , deg =  
57

8  
E  
(b2)

Fig.3a Bode plot of  $G$  &  $GcG(s)$ ; aircraft modelFig.3b zoom-in Bode plot of  $G$  &  $GcG(s)$ 

9  
E L

(63)

Adjustment of phase compensator: aircraft model

```
dphi =  
      3  
phi_m1 =  
      56  
alpha1 =  
      0.0935  
T1 =  
      0.5126
```

```
Gc1 =  
  
      0.5126 s + 1  
-----  
      0.04792 s + 1
```

Continuous-time transfer function.

<sup>10</sup>E (64) Test adjusted compensator

Fig. 4 shows the performance of the system with adjusted compensator  $G_c, G$  compared to the original system  $G$ .

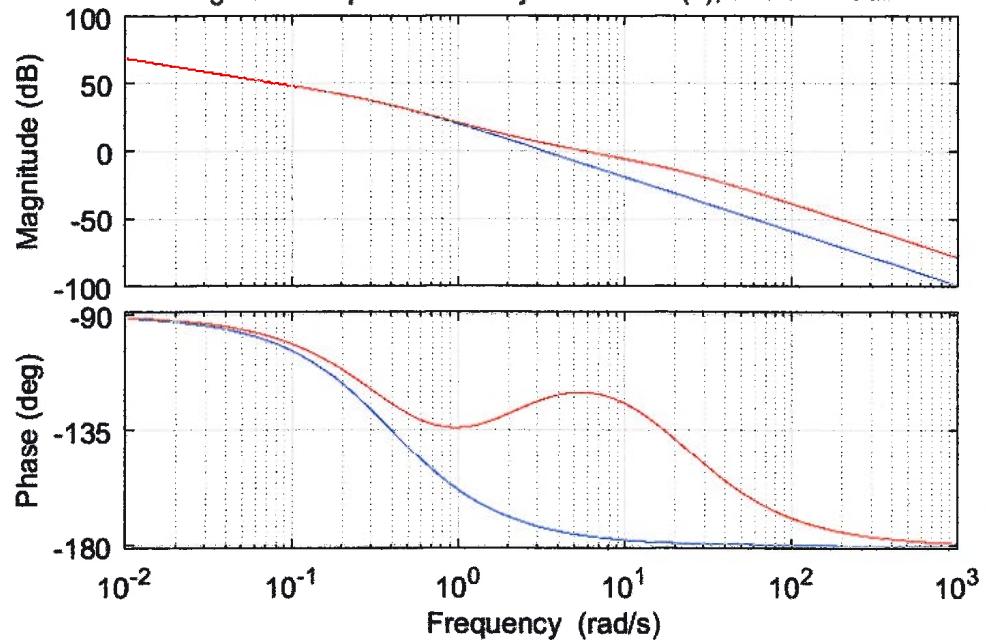
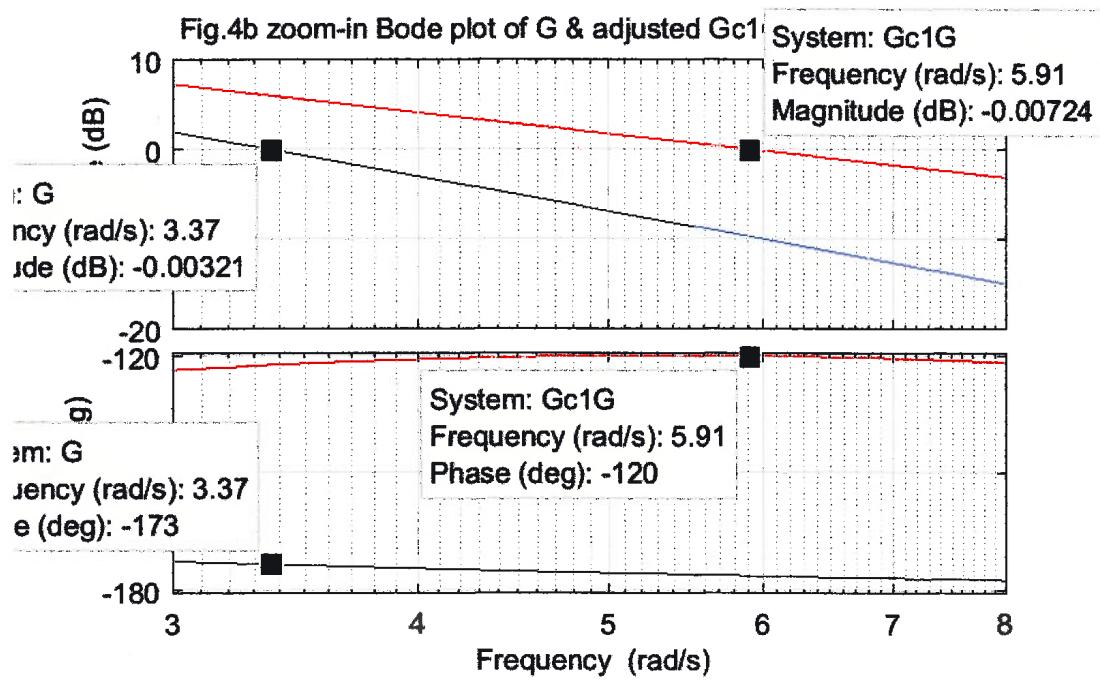
The new phase value at  $(G_c, G) = 0 \text{ dB}$  is:

$$\varphi_{c_1} = -120^\circ \text{ at } 5.91 \text{ rad/s where } |G_c, G| = 0 \text{ dB.}$$

The new phase margin is

$$\gamma_{c_1} = \varphi + 180^\circ = -120^\circ + 180^\circ = 60^\circ = \text{PM (II).}$$

The adjusted system meets the phase margin specification  $\text{PM} = 60^\circ$ .

11  
eFig.4a Bode plot of G & adjusted  $Gc_1G(s)$ ; aircraft modelFig.4b zoom-in Bode plot of G & adjusted  $Gc_1G(s)$ 

<sup>12</sup>  
6

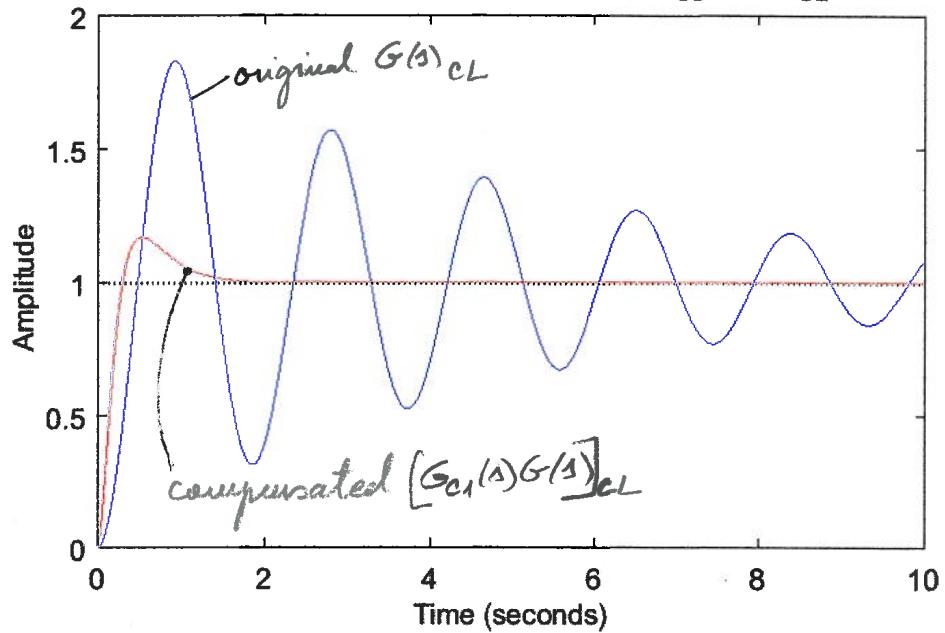
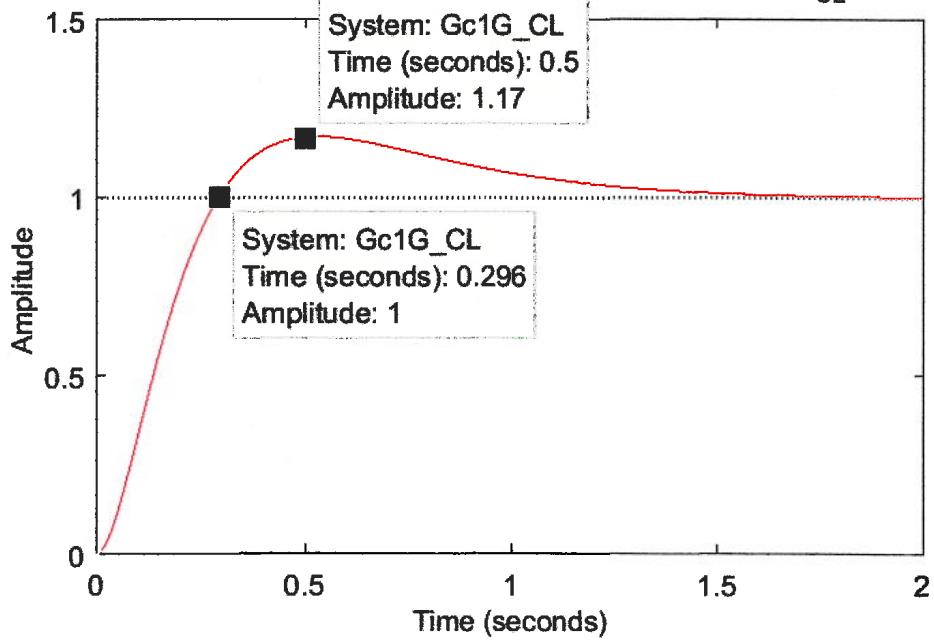
(c) Time response behavior of the compensated system

Fig 5 displays the time response of the compensated system  $(G_c G)_{CL}$  compared with the response of the original system  $G_{CL}$ . It can be observed that the compensated system has a fast rise time and a small overshoot

$$t_r = 0.296 \text{ sec} \quad (12)$$

$$M_p = 17\%$$

The original system had a longer rise time, a much higher overshoot, and took a long time to settle.

13  
6Fig.5a Step response aircraft model  $G_{CL}$ ,  $Gc_1G_{CL}$ Fig.5b zoom-in Step response aircraft model  $Gc_1G_{CL}$ 

```

1 % Phase compensator design: aircraft model
2 %% initialization
3 - clc %clear command window
4 - clear %removes all variables from workspace; release memory
5 - format compact
6 - close all %closes all figures
7 - s=tf('s');
8 %% original aircraft model
9 - display('Phase margin: aircraft model')
10 - K=114; % gain
11 - J=10; % inertia
12 - c=4; % damping
13 - display('input data')
14 - display([K J c],' K | J | c')
15 - G=K/(J*s^2+c*s) % G(s)
16 - display('=====')
17 %% specs: MIL-DTL-9490E margin requirements
18 - GM=10; % gain margin spec, dB
19 - PM=60; % phase margin spec, deg
20 - display([GM PM], 'GM, dB | PM, deg')
21 - figure(1)
22 - subplot(2,1,1)
23 - bode(G)
24 - grid
25 - title('Bode plot of original G(s); aircraft model')
26 - subplot(2,1,2)
27 - d1=0;d2=1;N=1e3; w=logspace(d1,d2,N);
28 - bode(G,w)
29 - grid
30 - title('zoom-in Bode plot of original G(s); aircraft model')
31 - display('=====')
32 - % READ ON PLOT: phase in deg for |G|=0 dB
33 - % phi=-173;
34 - phi=input('Input phase in deg when |G|=0 dB, phi=');
35 - gamma=phi-(-180); % gamma = phase margin, deg
36 - % display([phi], 'phi = phase at |G|=0 dB point, deg')
37 - display([gamma], 'gamma = phase margin, deg')
38 - display('=====')
39 %% add compensator Gc(s)
40 - display('Phase compensator design: aircraft model')
41 - phi_m=PM-gamma % maximum compensator phase that need to be obtained
42 - alpha=(1-sind(phi_m))/(1+sind(phi_m)) % compensator attenuation factor alpha
43 - display(1/sqrt(alpha), '1/sqrt(alpha)') % expected gain rise from compensator
44 - Gc_wm=mag2db(1/sqrt(alpha)); % expected gain rise from compensator, dB
45 - G_wm=-Gc_wm;
46 - display(G_wm, 'G_wm, dB')
47 - figure(2) % Bode plot to find the freq wm
48 - d1=log10(3);d2=log10(12);N=1e3; w=logspace(d1,d2,N);
49 - bode(G,w)
50 - grid
51 - title('zoom-in Bode plot of |G| to find freq wm; aircraft model')
52 - % READ ON PLOT: frequency for which |G|=G_wm
53 - % wm=5.83
54 - wm=input('Input frequency in rad/s when |G|=G_wm, wm=');
55 - T=1/wm/sqrt(alpha) % time constant of the compensator

```

```

56 %% Test compensated system
57 figure(3)
58 subplot(2,1,1)
59 Gc=(s*T+1)/(s*alpha*T+1)
60 GcG=Gc*G;
61 bode(G,GcG)
62 grid
63 title('Bode plot of G & GcG(s); aircraft model')
64 subplot(2,1,2)
65 d1=log10(3);d2=log10(8);N=1e3; w=logspace(d1,d2,N);
66 bode(G,GcG,w)
67 grid
68 title('zoom-in Bode plot of G & GcG(s); aircraft model')
69 % READ ON PLOT: phase in deg for |GcG|=0 dB
70 % phi_c=-123;
71 phi_c=input('Input phase in deg when |GcG|=0 dB, phi_c=');
72 gamma_c=phi_c-(-180); % gamma_c = phase margin of GcG, deg
73 display([phi_c],'phi_c = phase at |GcG|=0 dB point, deg')
74 display([gamma_c],'gamma_c = phase margin of GcG, deg')
75 display('====')
76 %% Adjust compensator to reach PM requirements
77 display('Adjustment of phase compensator: aircraft model')
78 dphi=PM-gamma_c % additional phase shift needed
79 phi_m=phi_m+dphi % adjusted phase shift
80 alpha1=(1-sind(phi_m))/(1+sind(phi_m)) % adjusted alpha
81 Gc1_wm=mag2db(1/sqrt(alpha1)); % expected gain rise from compensator, dB
82 G_wm1=-Gc1_wm;
83 display(G_wm1,'G_wm1, dB')
84 figure(4) % Bode plot to find the freq w_m for |G_wm|dB=-|Gc_wm|dB
85 d1=0;d2=1;N=1e3; w=logspace(d1,d2,N);
86 bode(G,w)
87 grid
88 title('zoom-in Bode plot of G(s) to find new freq w_m1; aircraft model')
89 % READ ON PLOT: frequency for which |G|=G_wm1, dB
90 % w_m1=6.1;
91 w_m1=input('Input frequency rad/s when |G|=G_wm1, w_m1=');
92 T1=w_m1/sqrt(alpha1) % updated time constant of the compensator
93 %% Final compensator design
94 Gc1=(s*T1+1)/(s*alpha1*T1+1)
95 Gc1G=Gc1*G;
96 figure(5)
97 subplot(2,1,1)
98 bode(G,Gc1G)
99 grid
100 title('Bode plot of G & adjusted Gc1G(s); aircraft model')
101 subplot(2,1,2)
102 d1=log10(3);d2=log10(8);N=1e3; w=logspace(d1,d2,N);
103 bode(G,Gc1G,w)
104 grid
105 title('zoom-in Bode plot of G & adjusted Gc1G(s); aircraft model')
106 phi_c1=input('Input phase in deg for |Gc1G|=0 dB, phi_c1=');
107 gamma_c1=phi_c1-(-180); % gamma_c1 = phase margin of Gc1G, deg
108 display([phi_c1],'phi_c1 = phase at |Gc1G|=0 dB point, deg')
109 display([gamma_c1],'gamma_c1 = phase margin of Gc1G, deg')
110 %% Plot step response response of the original and compensated systems
111 figure(6)
112 Tf=10; dt=0.01; t=0:dt:Tf;
113 G_CL=feedback(G,1);
114 Gc1G_CL=feedback(Gc1G,1);
115 subplot(2,1,1)
116 step(G_CL,Gc1G_CL,t)
117 title('step response aircraft model G_C_L, Gc1G_C_L')
118 subplot(2,1,2)
119 Tzoom=2; Nt=1e3; dt=Tzoom/Nt; tz=0:dt:Tzoom;
120 step(Gc1G_CL,tz,'r')
121 title('zoom-in step response aircraft model Gc1G_C_L')
122 ylim([0 1.5])

```

Phase margin: aircraft model

input data

$$\begin{array}{|c|c|c|} \hline K & | & J \\ \hline 114 & & 10 \\ \end{array} \quad \begin{array}{|c|c|} \hline c = & 4 \\ \hline \end{array}$$

$G =$

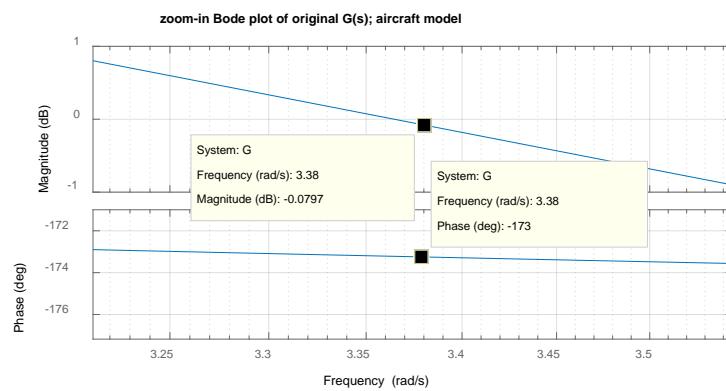
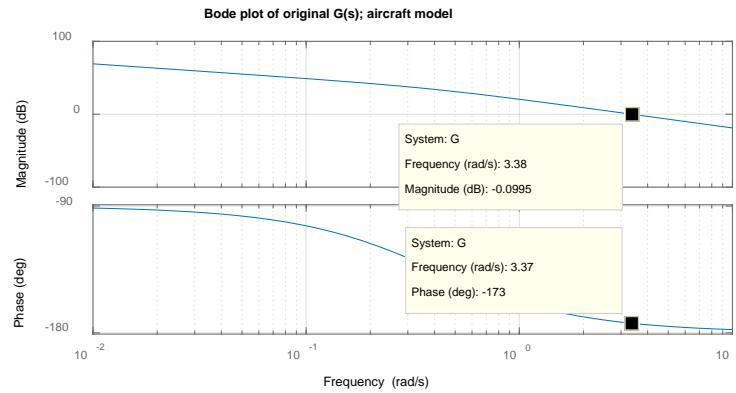
$$\frac{114}{10 s^2 + 4 s}$$

Continuous-time transfer function.

=====

$$\begin{array}{|c|c|} \hline GM, dB & PM, deg \\ \hline 10 & 60 \\ \hline \end{array}$$

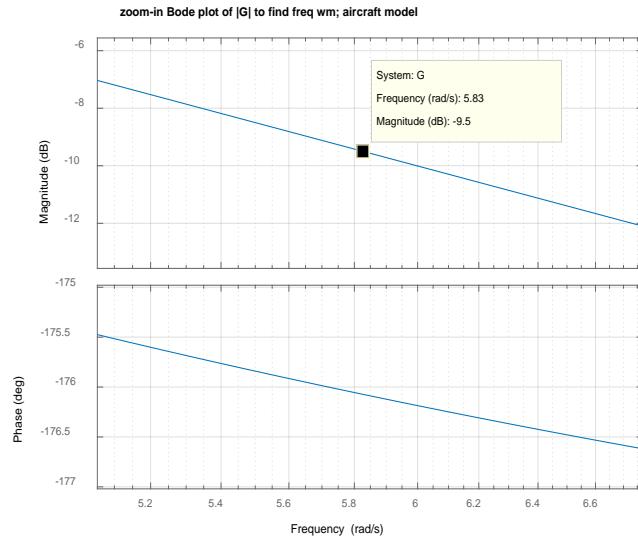
=====



Input phase in deg when  $|G|=0$  dB,  $\phi=-173$   
 $\gamma = \text{phase margin, deg} =$

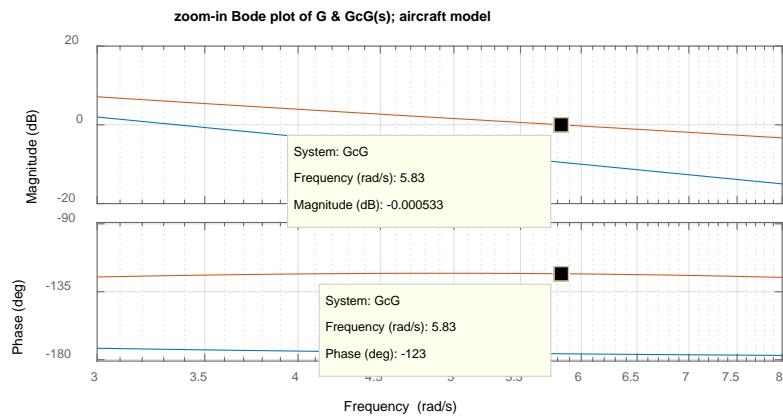
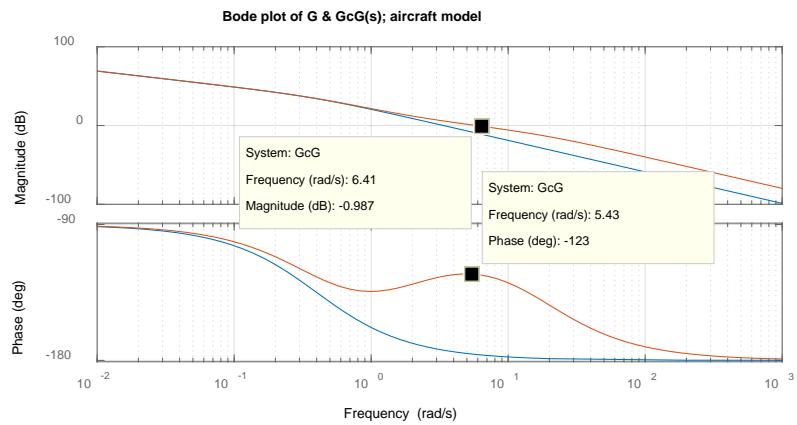
7

```
Phase compensator design: aircraft model
phi_m =
    53
alpha =
    0.1120
1/sqrt(alpha) =
    2.9887
G_wm, dB =
    -9.5096
```



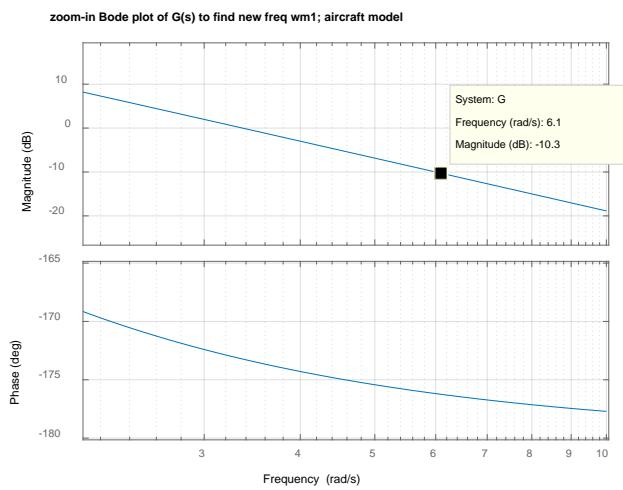
```
Input frequency in rad/s when  $|G|=G_{wm}$ ,  $w_m=5.83$ 
T =
    0.5126
```

```
Gc =
0.5126 s + 1
-----
0.05739 s + 1
```



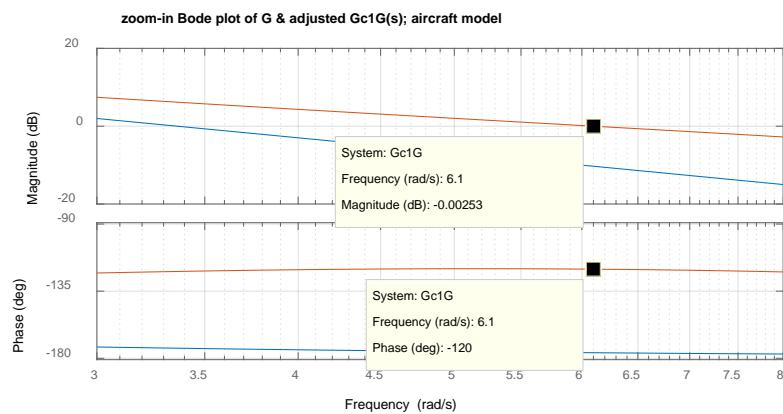
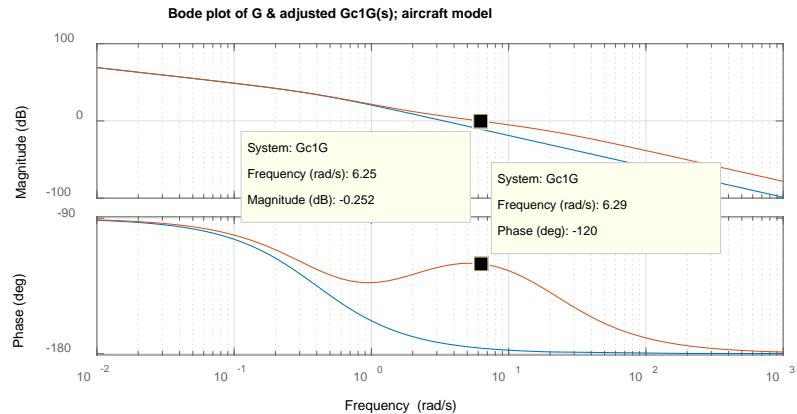
```
Input phase in deg when |GcG|=0 dB, phi_c=-123
phi_c = phase at |GcG|=0 dB point, deg =
-123
gamma_c = phase margin of GcG, deg =
57
```

```
Adjustment of phase compensator: aircraft model
dphi =
    3
phi_m1 =
    56
alpha1 =
    0.0935
G_wm1, dB =
-10.2932
```

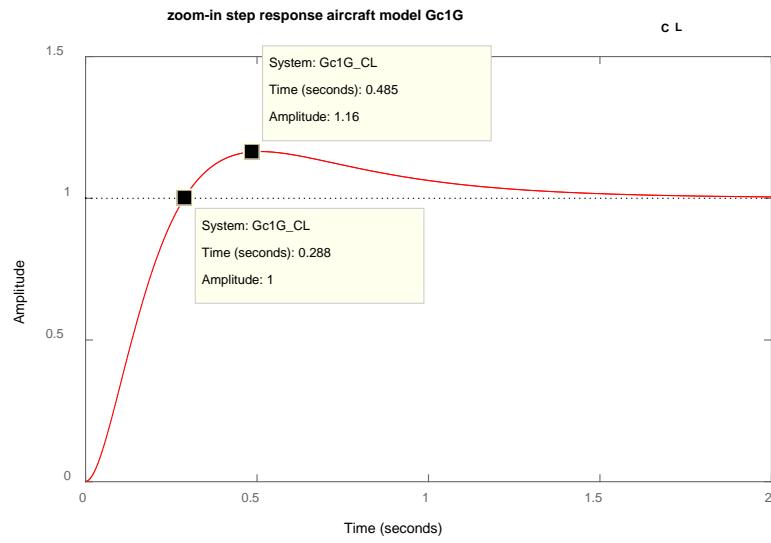
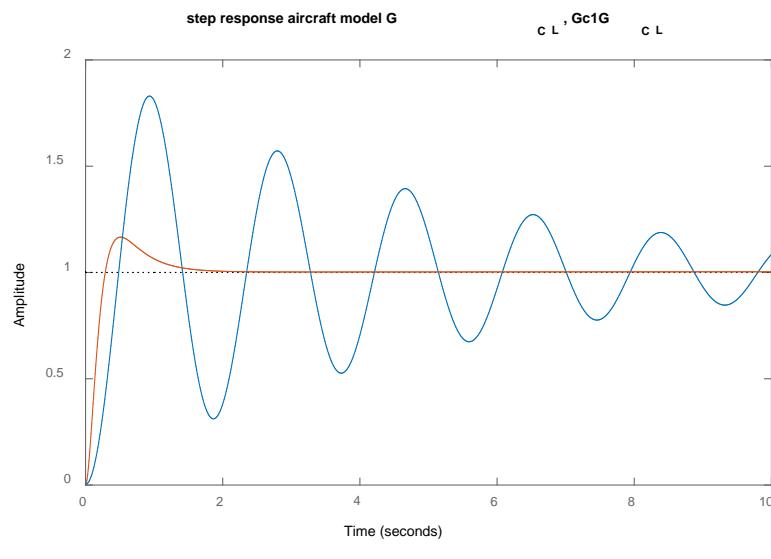


```
Input frequency rad/s when |G|=G_wm1, wm1=6.1
T1 =
    0.5362
```

```
Gc1 =
0.5362 s + 1
-----
0.05012 s + 1
```



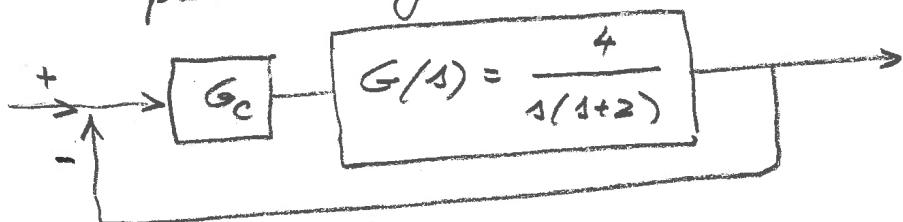
```
Input phase in deg for |Gc1G|=0 dB, phi_c1=-120
phi_c1 = phase at |Gc1G|=0 dB point, deg =
-120
gamma_c = phase margin of GcG, deg =
60
```



C7

Ex. 11.6

Design a lead compensator to reduce servomotor ramp error and meet phase and gain margin requirements.



Given :  $G(s) = \frac{4}{s(s+2)}$

- Find :
- (a) static velocity error const.  $K_v$
  - phase margin,  $\gamma$
  - gain margin,  $(K_g)_{dB}$

(b) design compensator to achieve:

$$K_v \geq 20 \text{ /sec}$$

$$\gamma \geq 50^\circ \quad (\text{PM} = 50^\circ)$$

$$(K_g)_{dB} \geq 10 \text{ dB} \quad (\text{GM} = 10 \text{ dB})$$

Solution

(a) characterize current system

$$\bullet K_v = \lim_{s \rightarrow 0} s G(s) = \lim_{s \rightarrow 0} s \frac{4}{s(s+2)} = \frac{4}{2} = 2 \text{ /sec.}$$

$$\bullet \angle G(s=0 \text{ dB}) = -128^\circ; \gamma = -128^\circ + 180^\circ = 52^\circ$$

$$\bullet (K_g)_{dB} = \infty \text{ because } \angle G \text{ never crosses } -180^\circ \text{ line}$$

C7a

Fig.1a Bode plot of original G(s); Example 11.6

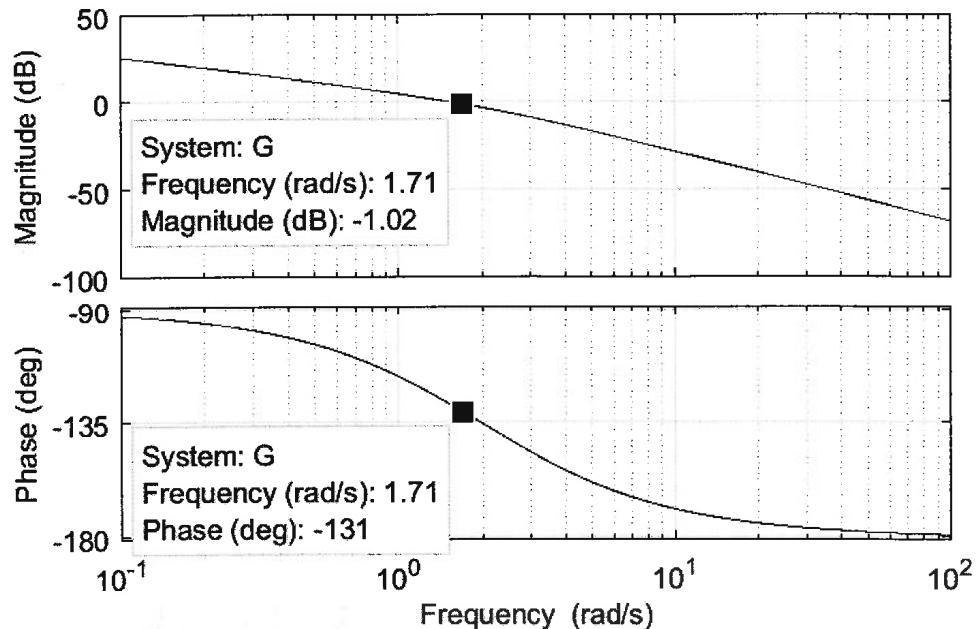
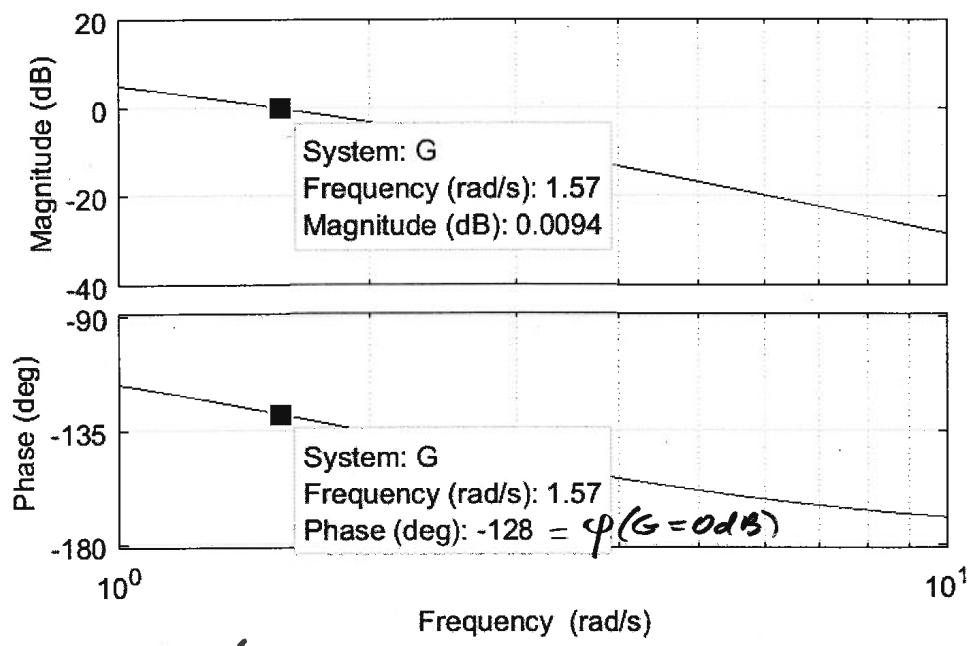


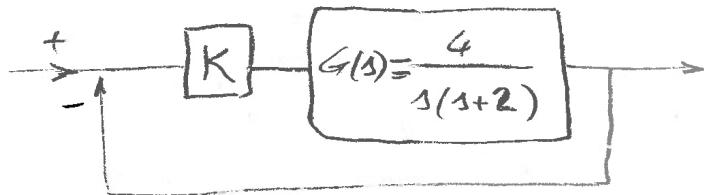
Fig.1b zoom-in Bode plot of original G(s); Example 11.6



$$\gamma = -128 + 180 = 52^\circ$$

436/523

C8) (b) Design P-controller to improve  $K_v$



Roadmap

- (1) add  $K$  to improve  $K_v$  to  $K_v^* = 20/\text{sec}$
- (2) Notice that adding  $K$  shifts magnitude plot upward and spoils the phase margin.
- (3) add compensator  $G_c$  to improve phase and bring phase back to  $\gamma^* \leq 50^\circ$

Design

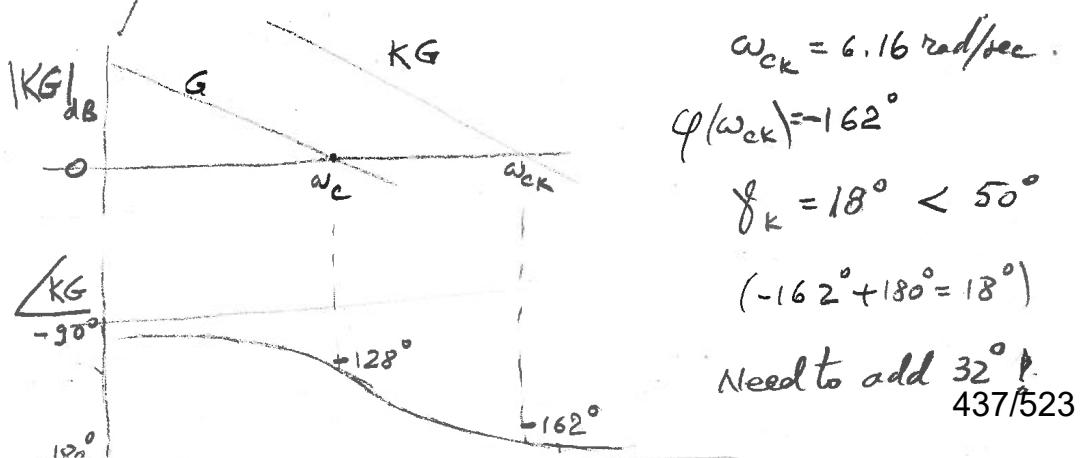
D1. Calculate  $K$  needed to bring  $K_v$  to value  $K_v^* = 20/\text{sec}$

$$K_v^* = \lim_{s \rightarrow 0} s G(s) = \lim_{s \rightarrow 0} K \cdot \frac{4}{s(s+2)} = 2K = 20$$

$$\rightarrow K = 10$$

$$K_{dB} = 20 \text{ dB}$$

D2. Draw Bode diagram for new system. We find that gain crossover freq  $\omega_c$  has moved to the right and the phase margin has decreased



c8a

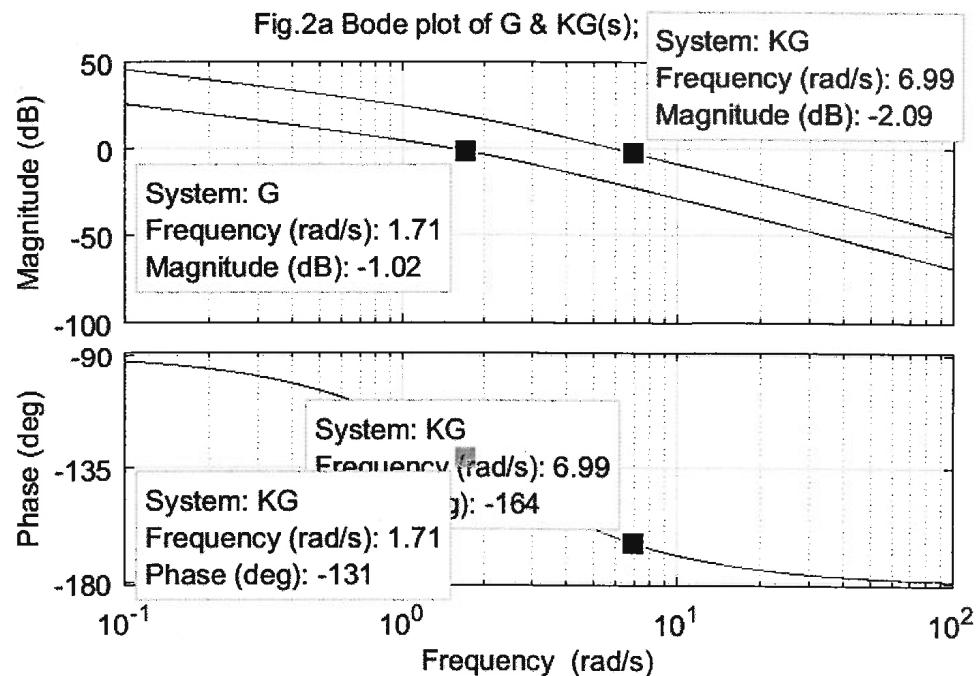
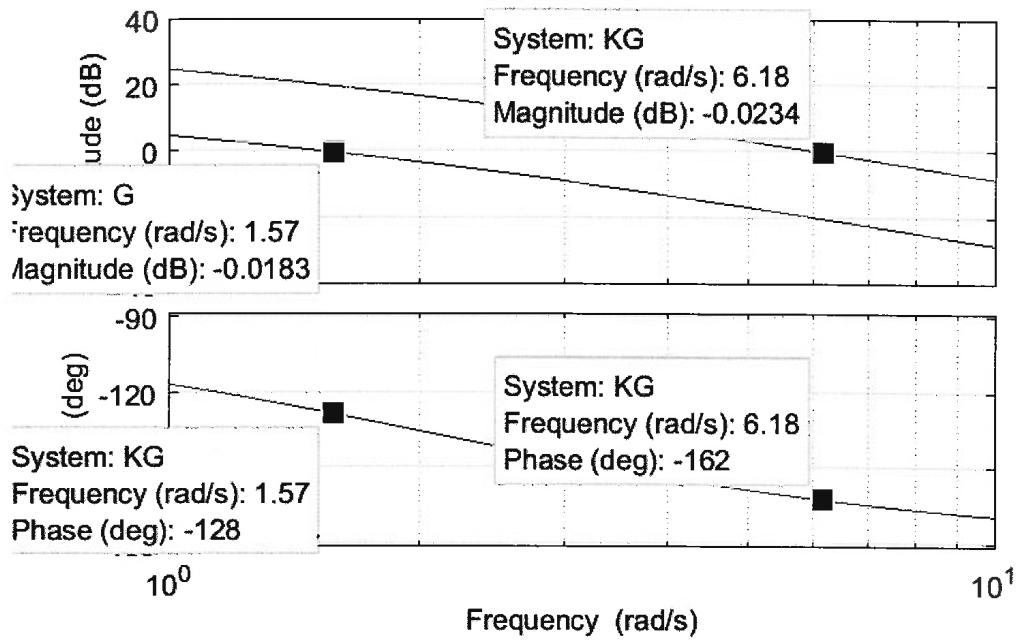
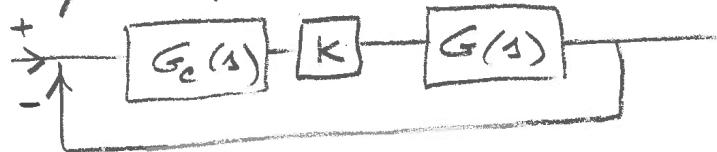


Fig.2b zoom-in Bode plot of G &amp; KG(s); Example 11.6



C9 Design compensator to improve  $\gamma$ .



We need lead compensator

$$G_c(s) = \frac{Ts + 1}{\alpha Ts + 1}$$

We need to improve the phase margin from  $\gamma = 18^\circ$  to  $\gamma_M = 50^\circ$ , i.e., we need the phase compensator to add  $\varphi_m = 32^\circ$

$$\text{Recall } \alpha = \left. \frac{1 - \sin \varphi_m}{1 + \sin \varphi_m} \right|_{\varphi_m = 32^\circ} = 0.3073$$

To calculate  $\omega_m$ , recall

$$|G_c(\omega_m)| = \frac{1}{\sqrt{\alpha}} = 1.8040 = 5.125 \text{ dB}$$

We identify  $\omega_n$  as the point on the Bode plot of  $KG(\omega)$  where

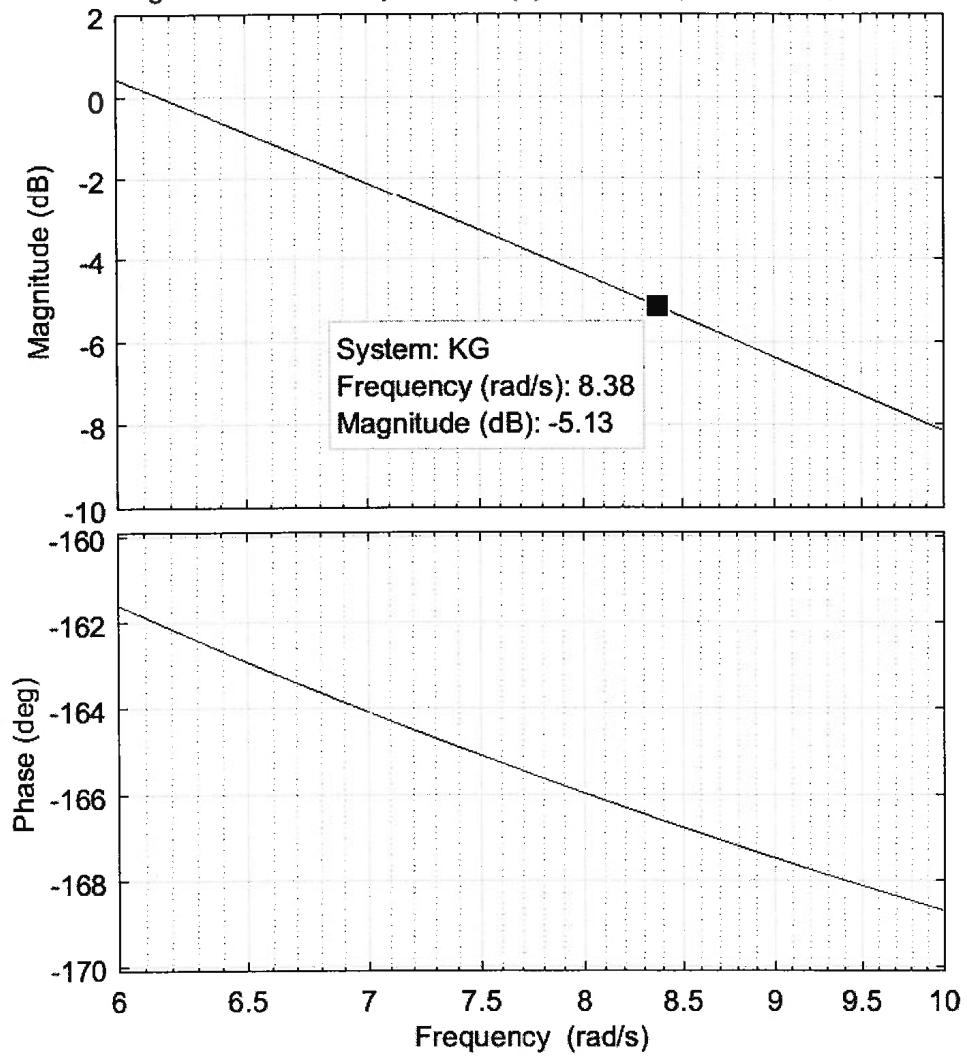
$$|KG(\omega_m)|_{\text{dB}} = -|G_c(\omega_n)|_{\text{dB}} = -5.125 \text{ dB.}$$

From Bode plot, we read

$$\omega_m = 8.38 \text{ rad/sec for } |KG|_{\text{dB}} = -5.13 \text{ dB}$$

$$\text{Hence } T = \frac{1}{\omega_m \sqrt{\alpha}} = 0.2153 \text{ sec.}$$

C9a

Fig.3 zoom-in Bode plot of KG(s) to find freq  $\omega_m$ ; Example 11.6

$$\omega_m = 8.38 \text{ rad/sec} \text{ for } |KG|_{\text{dB}} = -5.13 \text{ dB}$$

C96

Phase compensator design: Example 11.6

phi =

-128

gamma =

52

alpha =

0.3073

i/sqrt(alpha) =

1.8040

Gc\_wm, dB =

5.1250

T =

0.2153

Gc =

$$\frac{0.2153 s + 1}{0.06615 s + 1}$$

C9c

Fig.4a Bode plot of KG &amp; GcKG(s); Example 11.6

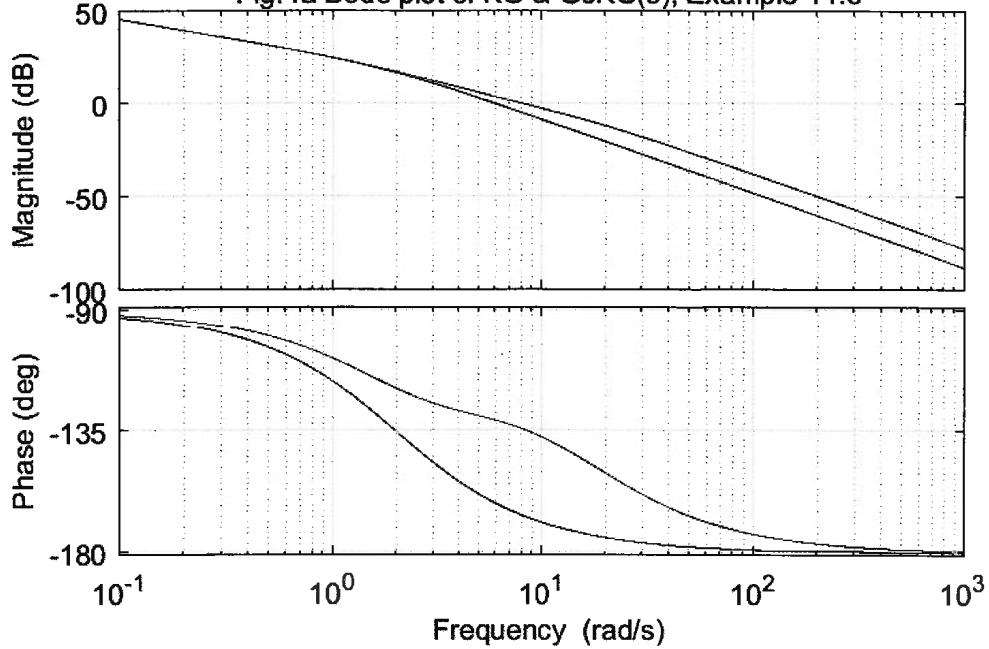
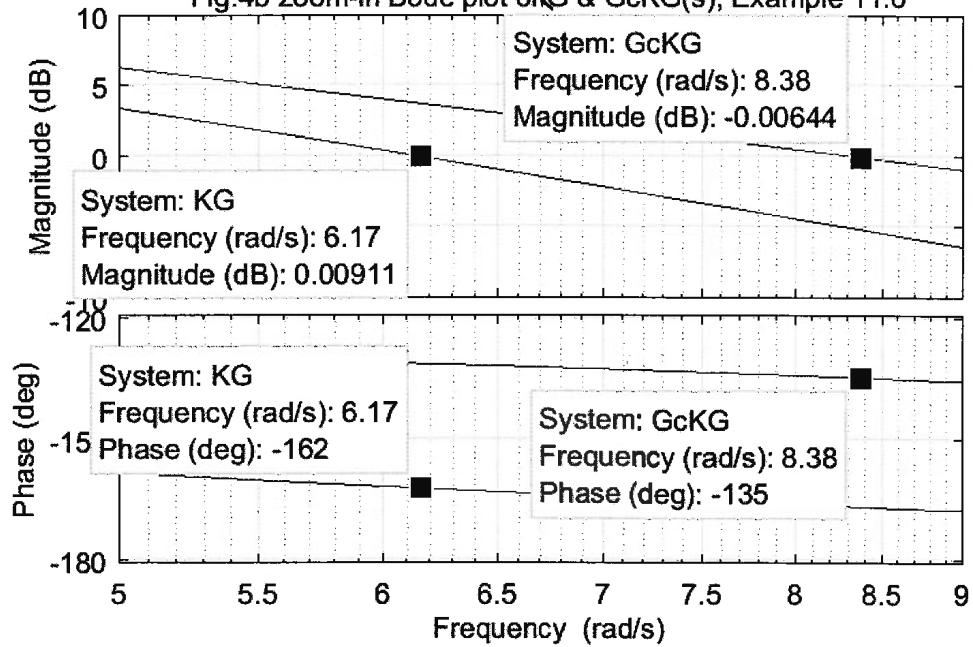


Fig.4b zoom-in Bode plot of KG &amp; GcKG(s); Example 11.6



C10 Test the compensator

Fig 4 shows that the compensator is not sufficient because the compensated phase margin is

$$\gamma_c = -135 + 180 = 45^\circ$$

We need  $50^\circ$ ; thus, we need to adjust the compensator to get to  $50^\circ$ . We need an additional  $5^\circ$  of compensation, i.e.,  $\Delta\varphi = 5^\circ$ . The new  $\varphi_m$  is

$$\varphi_{m_1} = \varphi_m + \Delta\varphi = 32 + 5 = 37^\circ$$

$$\text{The new } \alpha \text{ is } \alpha_1 = \frac{1 - \sin \varphi_{m_1}}{1 + \sin \varphi_{m_1}} = 0.2486$$

We keep  $T_1 = T$  and calculate new compensator

$$G_{C_1} = \frac{T_1 s + 1}{\alpha_1 T_1 s + 1}$$

With this new compensator, the Bode plot is as shown in Fig. 5. The phase at  $G_{C_1} KG = 0 \text{ dB}$  is  $-130^\circ$

$$\text{The margin is } \gamma_{c_1} = -130^\circ + 180^\circ = 50^\circ = \gamma_M$$

We have met the specification!!

*C10a*

```
dphi =
5

phi_m1 =
37

alpha1 =
0.2436

T1 =
0.2153

Gc1 =
0.2153 s + 1
-----
0.05352 s + 1
```

C16b

Fig.5a Bode plot of KG &amp; adjusted Gc1KG(s); Example 11.6

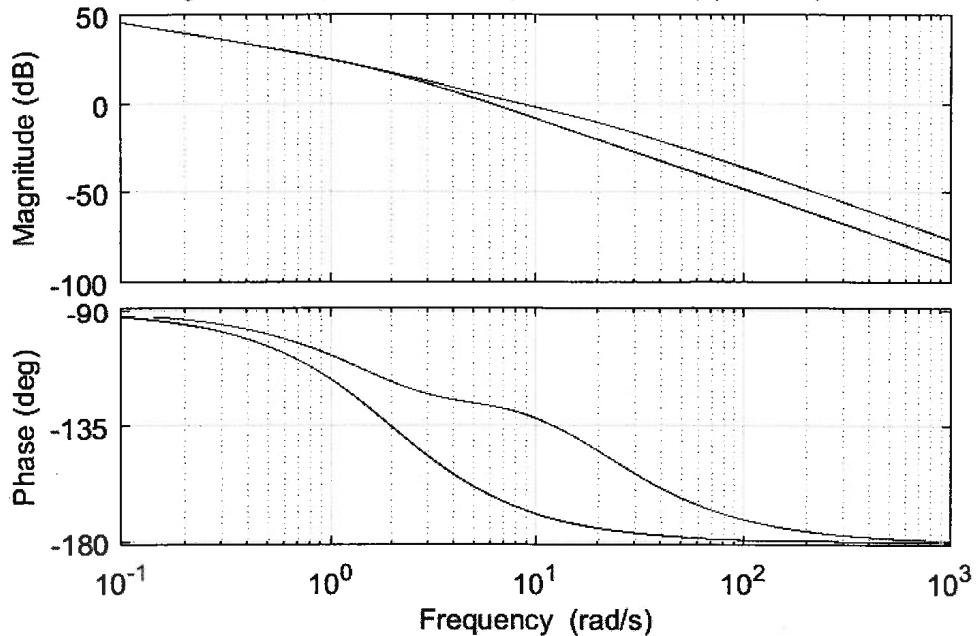
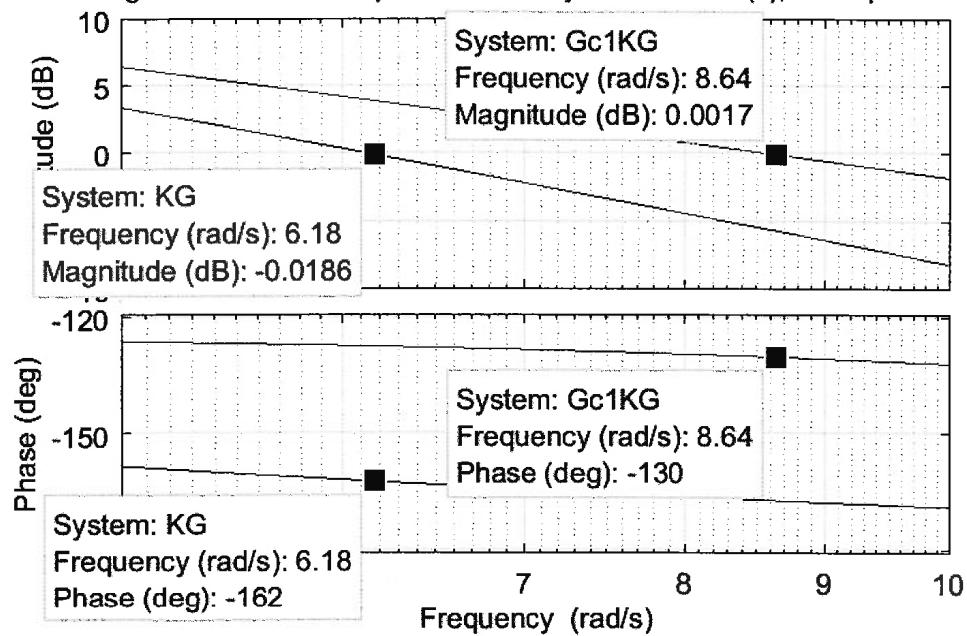
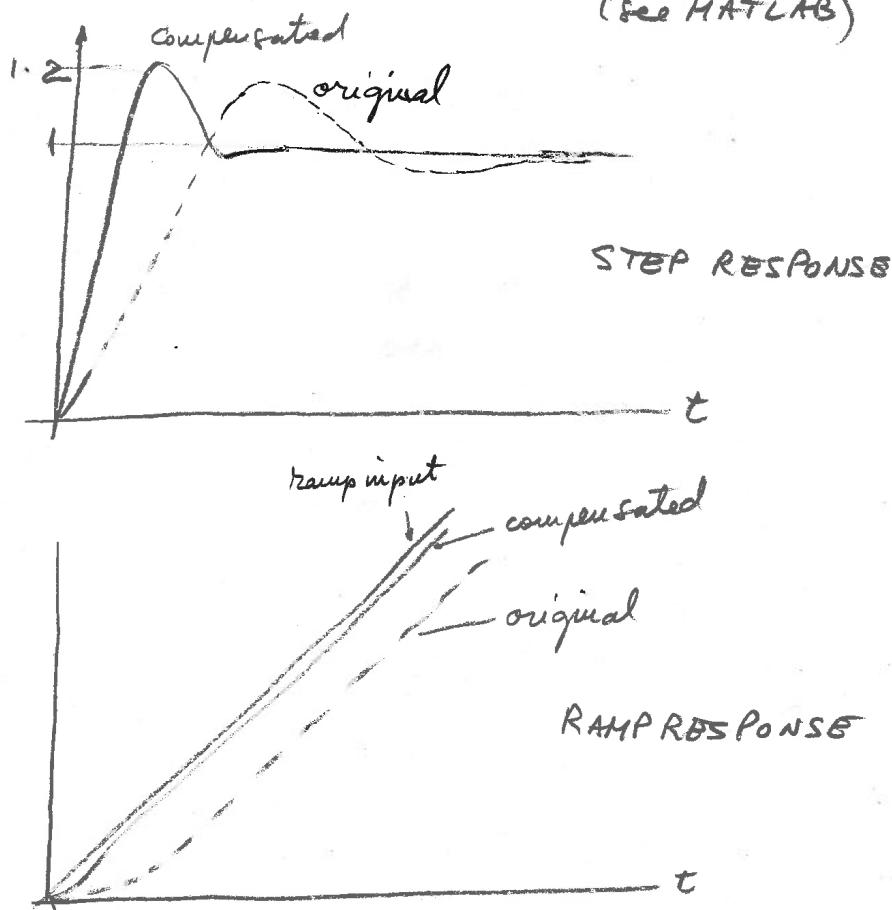


Fig.5b zoom-in Bode plot of KG &amp; adjusted Gc1KG(s); Example 11.6



C11 Plot step response and ramp response of the original and compensated systems (see MATLAB)



### Discussion

Ramp response has improved dramatically  
Step response has a faster rise time and  
a faster settling time, but it has a  
slightly higher overshoot

C 11 a

Fig.6a Step response Example 11.6

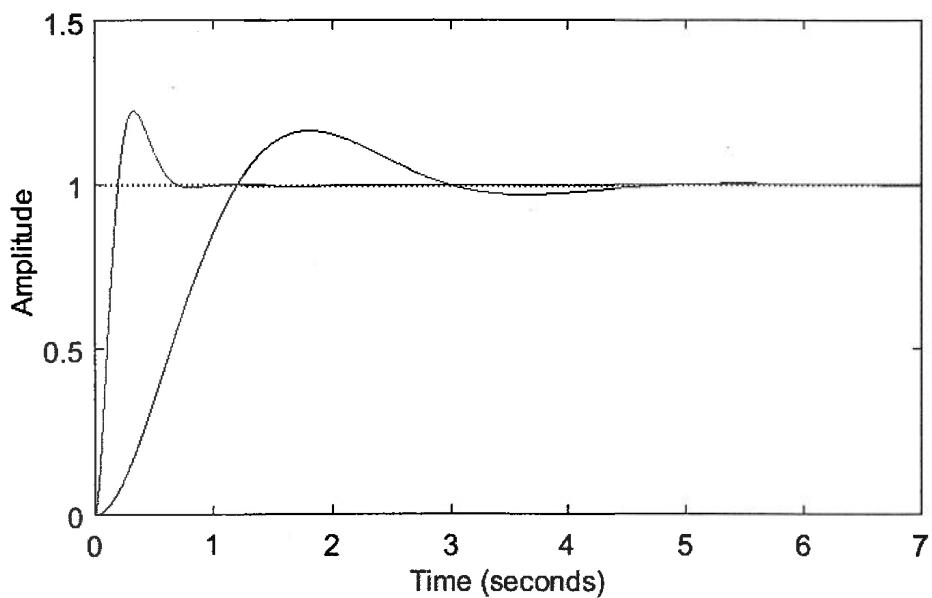
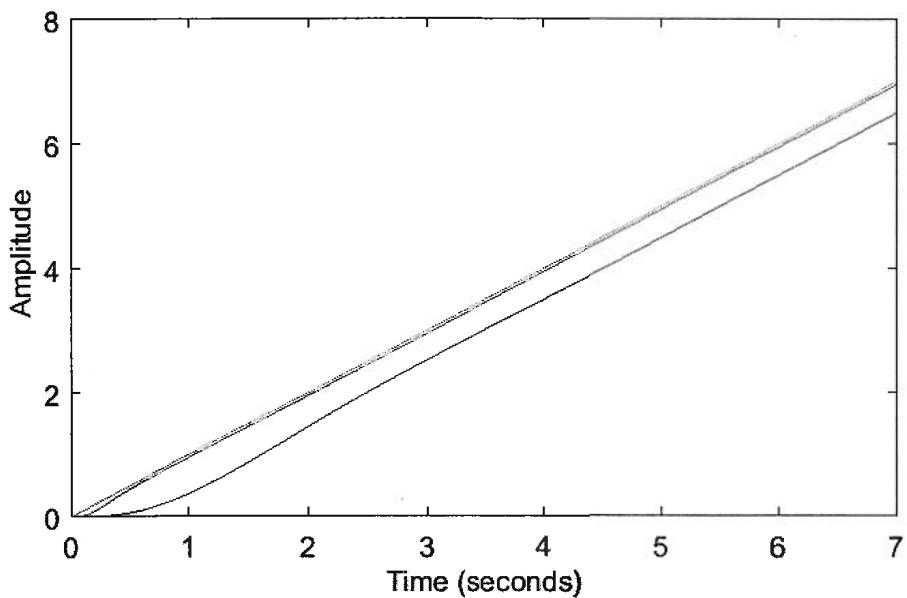


Fig.6b Ramp response Example 11.6



## 8.8 Lag Compensators

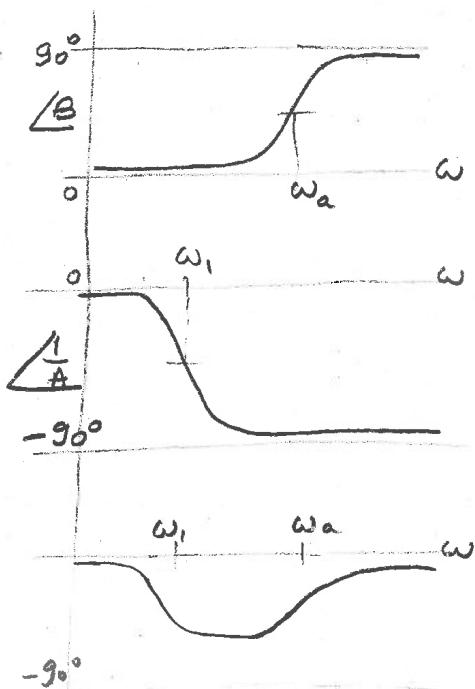
C4

LAG COMPENSATOR

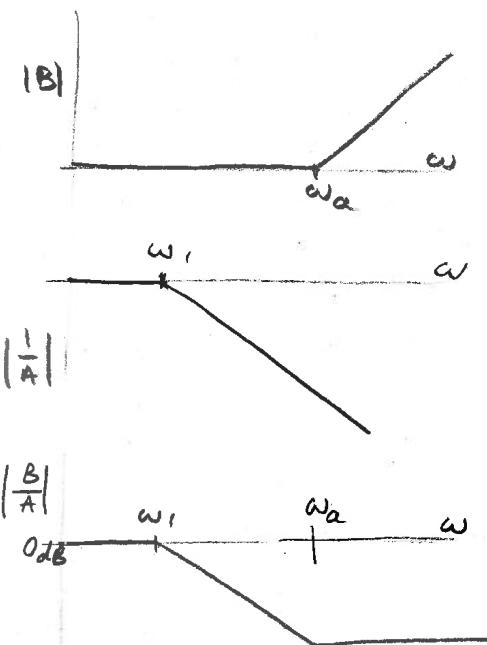
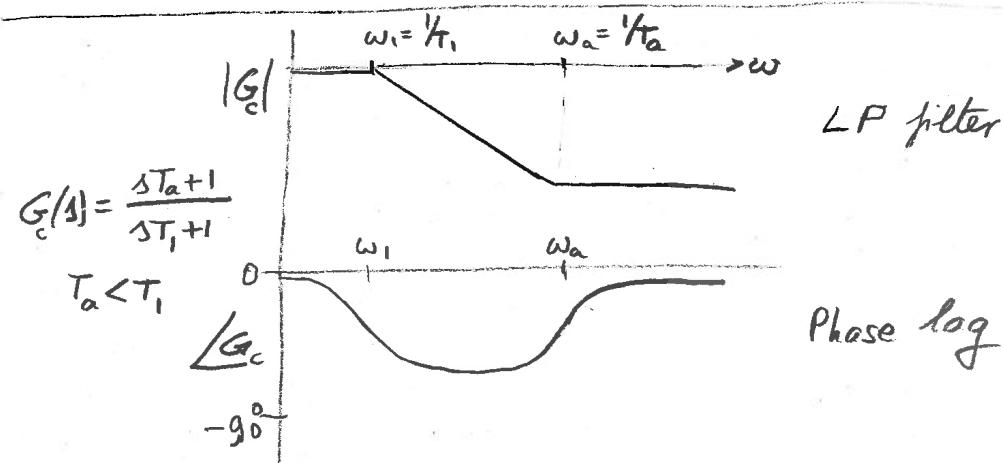
$$G_c = \frac{sT_a + 1}{sT_i + 1}$$

$$\omega_1 < \omega_a, \quad T_a < T_i$$

- Phase -



- Gain -

Adds "lag" between  $\omega_1$  &  $\omega_a$ low pass filter  
(higher freq. are reduced)

c5

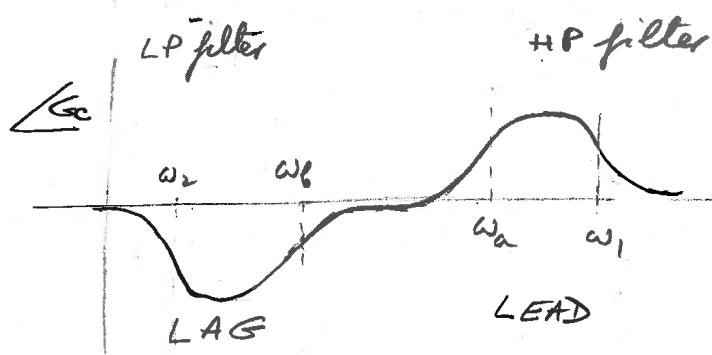
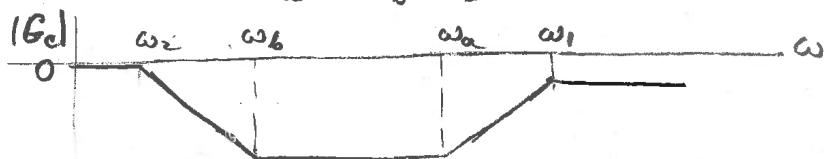
### LAG-LEAD COMPENSATOR (NOTCH FILTER)

$$G_c = \frac{1/T_a + 1}{1/T_1 + 1} \cdot \frac{1/T_b + 1}{1/T_2 + 1}$$

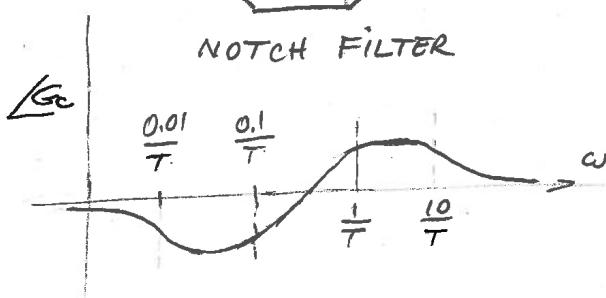
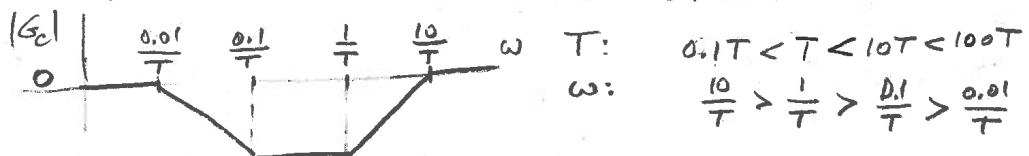
lead      lag

$$T_1 < T_a < T_b < T_2$$

$$\omega_1 > \omega_a > \omega_b > \omega_2$$



Example (Fig 11.35, p 648).  $G_c = \frac{1+1}{0.1T+1} \cdot \frac{10T+1+1}{100T+1+1}$



$$G(i\omega) = 1 \quad (0 \text{ dB})$$

$$G(i\omega) = 1 \quad (0 \text{ dB})$$

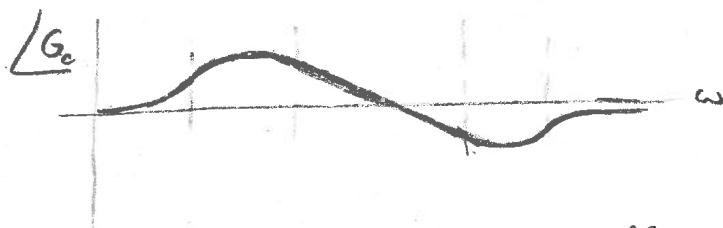
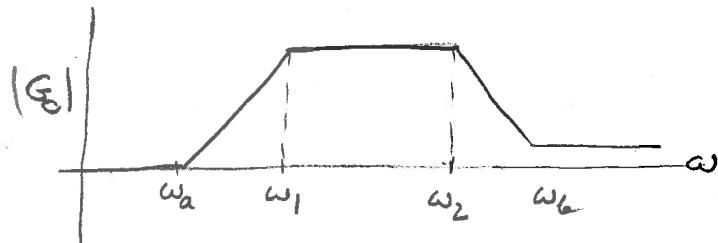
C6

**LEAD-LAG COMPENSATOR  
(BAND PASS FILTER)**

$$G_c = \frac{sT_a + 1}{sT_1 + 1} \cdot \frac{sT_b + 1}{sT_2 + 1}$$

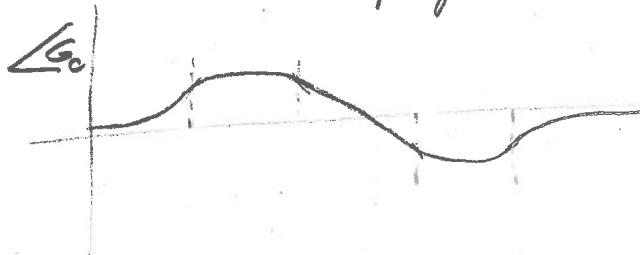
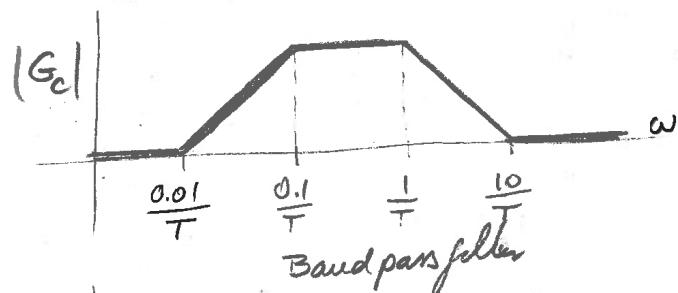
$$\omega_a < \omega_1 < \omega_2 < \omega_b$$

$$T_a > T_1 > T_2 > T_b$$



Example : Band pass filter

$$G_c = \frac{100T_1 + 1}{10T_1 + 1} \cdot \frac{0.1T_2 + 1}{T_2 + 1}$$



~~C6a~~ Notch filter  
Example

$$G_c = \frac{T_s + 1}{0.1T_s + 1} \cdot \frac{10T_s + 1}{100T_s + 1}$$

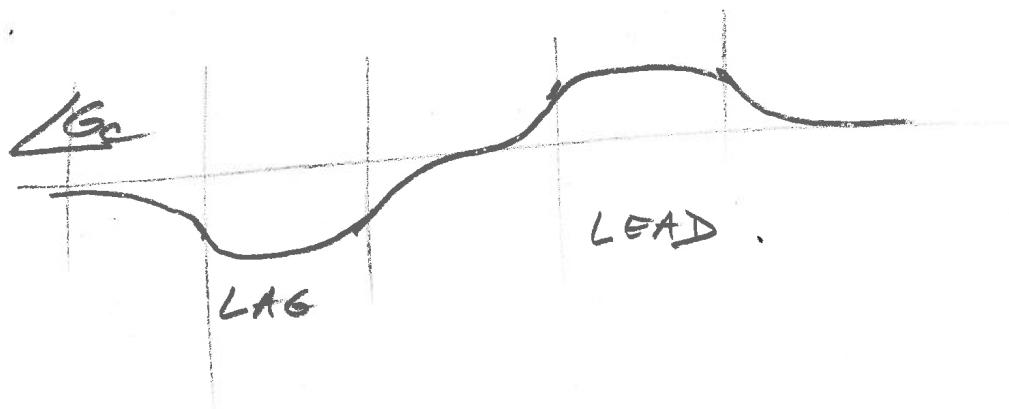
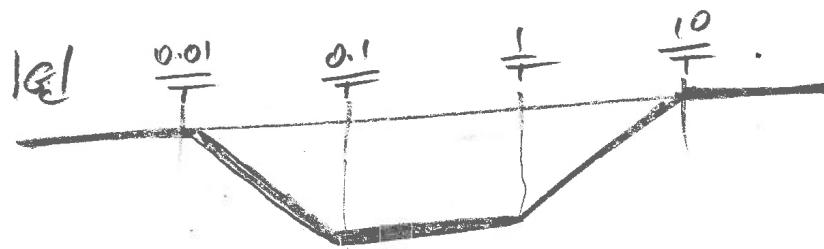
$$0.1T < T < \frac{T_1}{10} < \frac{T_2}{100T}$$

$$\omega_1 > \omega_a > \omega_b > \omega_2$$

$$\frac{10}{T} > \frac{1}{T} > \frac{0.1}{T} > \frac{0.01}{T}$$

$$\omega_2 < \omega_b < \omega_a < \omega_1$$

$$\frac{0.01}{T} \quad \frac{0.1}{T} \quad \frac{1}{T} \quad \frac{10}{T}$$



## 9 Combined Analysis

## Combined analysis

Purpose : use all the tools to evaluate the system stability

Time domain:

- Root loci of  $G$
- Step response of  $G_{CL}$

Frequency domain:

- Nyquist plot of  $G$
- Bode plots of  $G$ : margins  $\angle$  gain  $K_g$  phase  $\phi$

Given:  $G(s)$

Find: combined analysis: Quad plot

Nyquist plot	Margins on Bode plot
Root locus for Gain=1	Step response

Examples :

open MATLAB codes

aircraft roll mode
• Nyquist plot : STABLE
• root locus
• step response

Bode plot : INSUFFICIENT phase margin

B11.9 - STABLE : N-plot

root locus

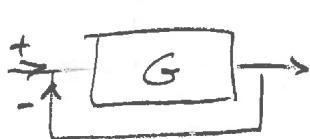
step response

- INSUFFICIENT phase margin

454/523

B11.8 UNSTABLE

Combined analysis  
of aircraft roll model.

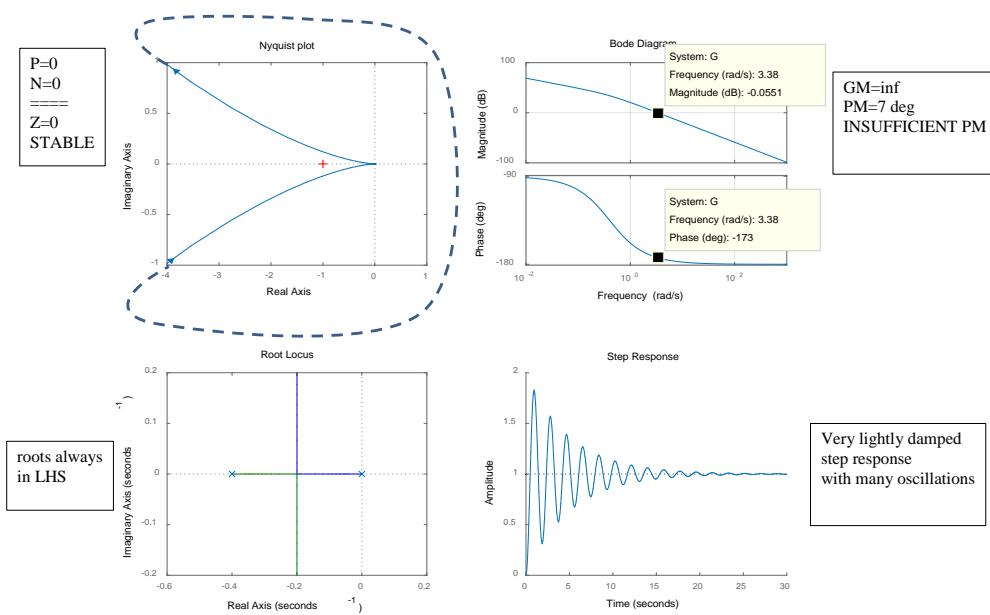


$$G(s) = \frac{K}{Js^2 + Cs} = \frac{114}{10s^2 + 4s}$$

$$\rho_1 = 0$$

$$\rho_2 = -0.4$$

See plot



2017 04 17

## Aircraft roll model combined analysis (cont.)

Nyquist.

✓ stable

 $Z=0$ 

Bode

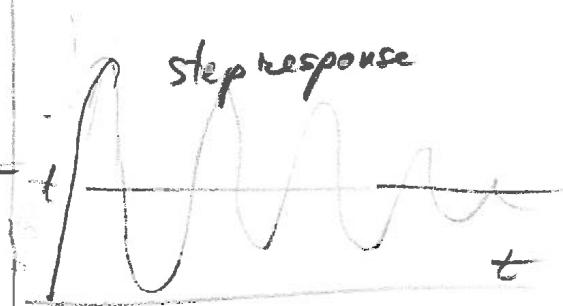
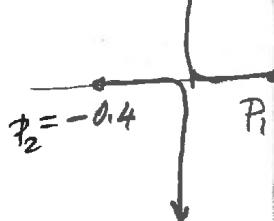
Margins

$$(K_3)_{dB} = \infty \quad \checkmark$$

$$\gamma^\circ = 7^\circ \text{ Not sufficient}$$

Root locus

No problem

Conclusion

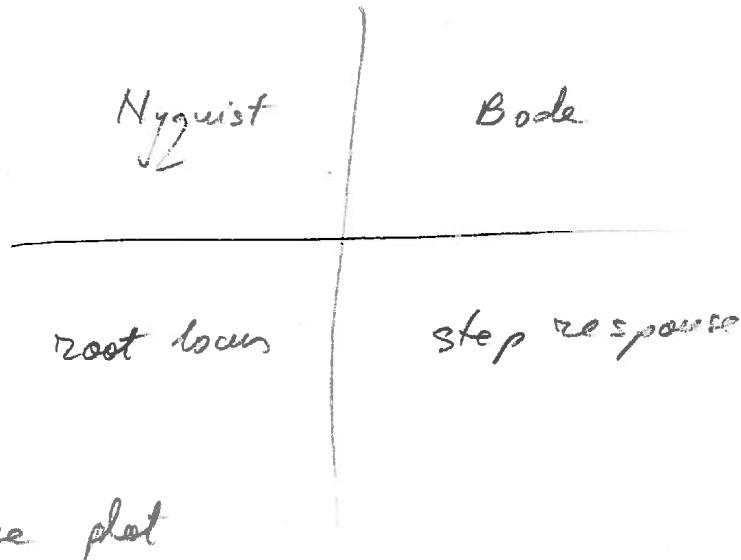
- Nyquist and root locus indicate stable system
- MARGIN ANALYSIS WARNS about low phase margin;  $\rightarrow$  COMPENSATOR need
- step response — oscillatory

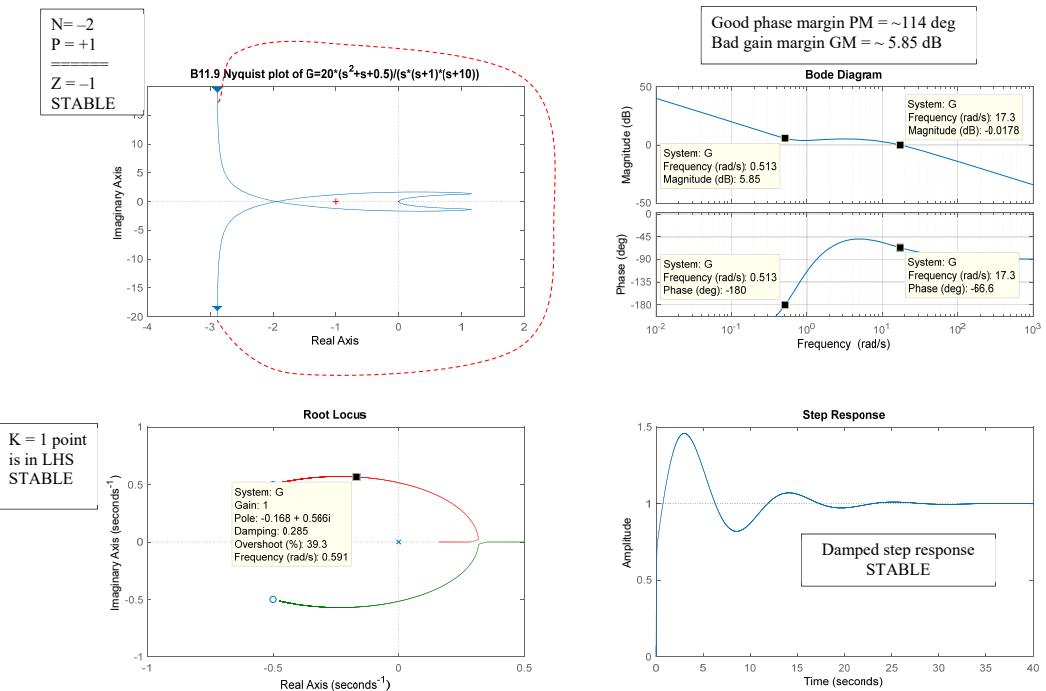
Problem B11.9



$$G(s) = \frac{20(s^2 + s + 0.5)}{s(s-1)(s+10)}$$

$$\left. \begin{array}{l} p_1 = 0 \\ p_2 = +1 \\ p_3 = -10 \end{array} \right\} P = 1$$

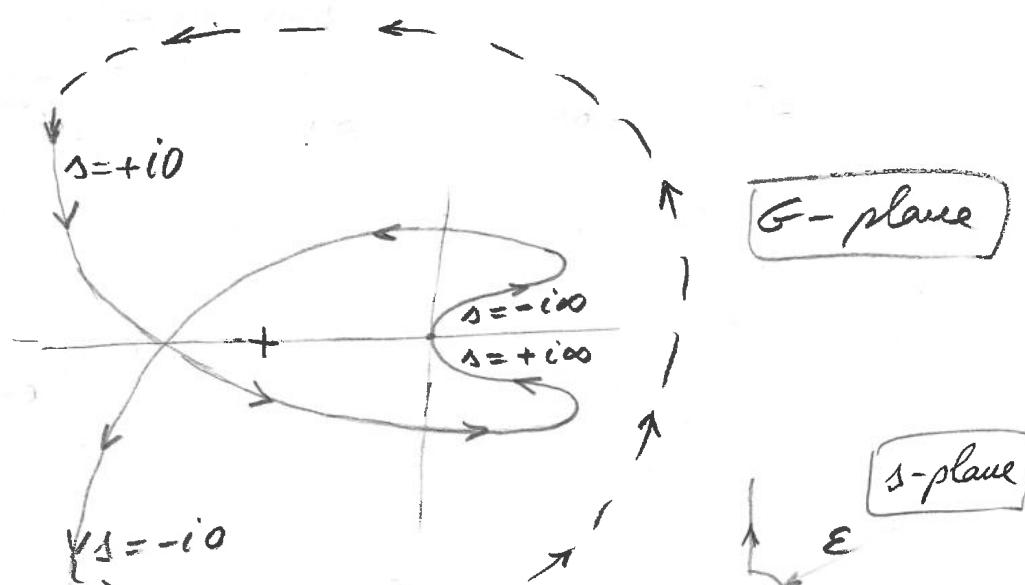




459/523

## Problem B11.9 (cont.)

Nyquist plot



$$G^o = \frac{\bar{z}_0}{-1} = -2 \frac{1}{\varepsilon} e^{-i\varphi}$$

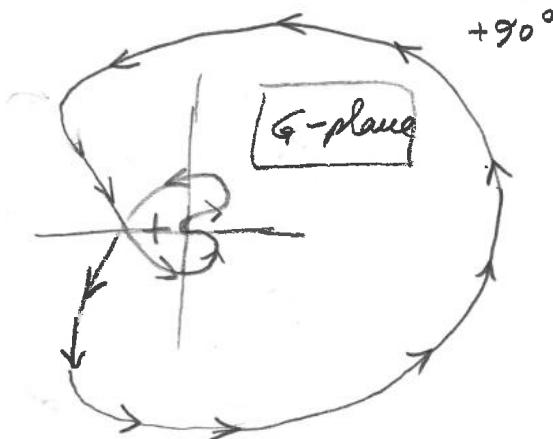
$$G^o = \frac{2}{\varepsilon} e^{i(180^\circ - \varphi)}$$

$$-90^\circ < \varphi < 90^\circ$$

$$s = \varepsilon e^{i\varphi}$$

$$-90^\circ < \varphi < 90^\circ$$

$$-90^\circ < \angle G^o < -270^\circ$$



$$\begin{aligned} N &= -2 \\ P &= 1 \\ \hline Z &= -1 \\ \boxed{\text{Stable}} \end{aligned}$$

(B11.9 cont.)

Bode plotGood phase margin:  $\gamma \approx 114^\circ \text{ at } 16.6 \frac{\text{rad}}{\text{sec}}$ 

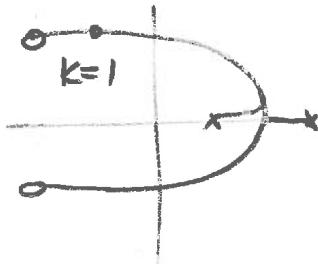
Bad gain margin:  $K_g = -5.3 \text{ dB} \text{ at } 0.55 \frac{\text{rad}}{\text{sec}}$   
 at low freq.

Good gain margin at higher  $\omega > 16 \frac{\text{rad}}{\text{sec}}$ ,  $\angle \gamma$   
 ( $\gamma$  does not cross  $-180^\circ$ )

There could be problems at low  $\omega$   
 $(\omega \approx 0.5 \text{ rad/sec})$  due to imperfections

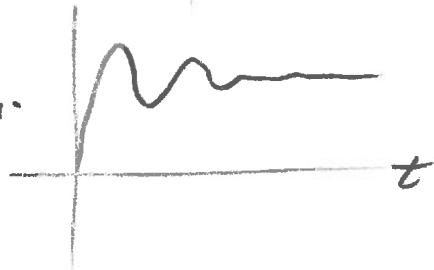
Root locus $K=1$  in LHS

stable

Step response

damped oscillation

STABLE

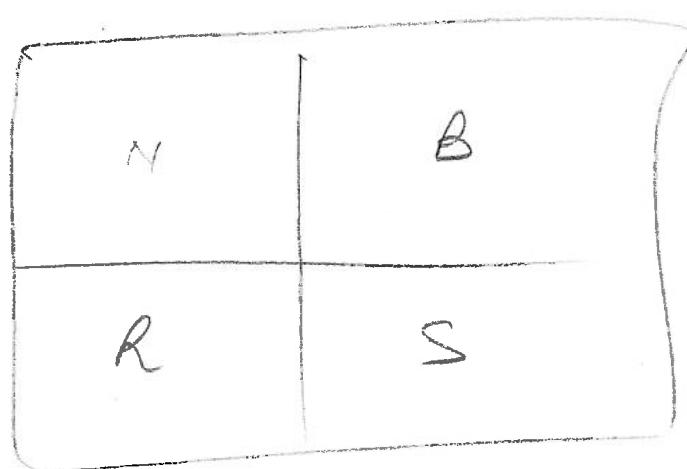


Problem B11.8



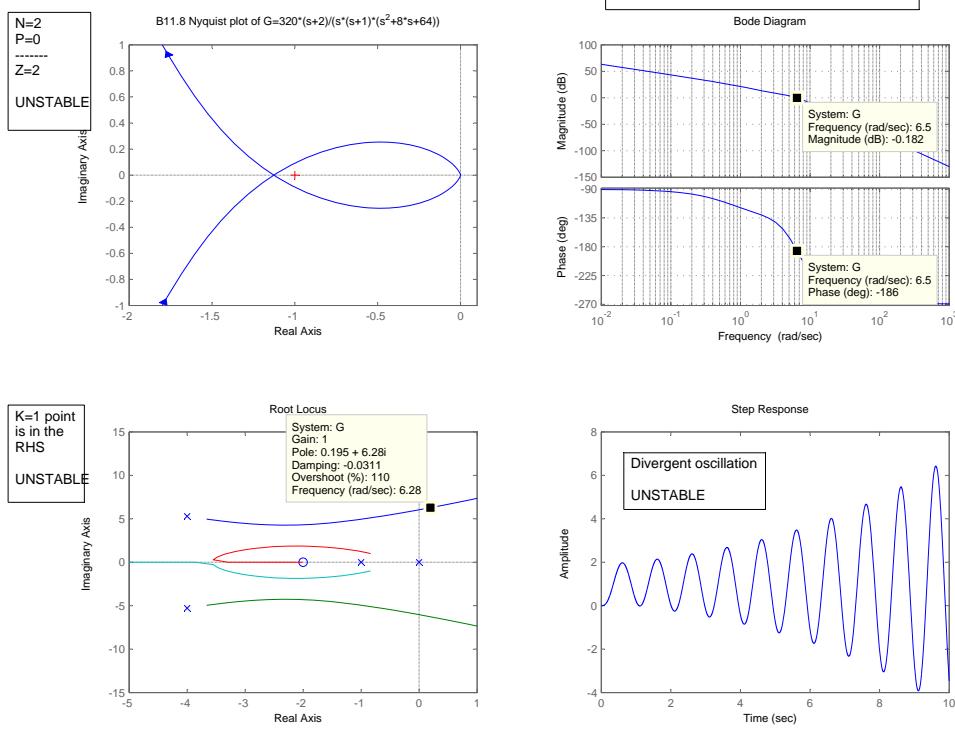
$$G(s) = \frac{320(s+2)}{s(s+1)(s^2 + 8s + 44)}$$

4 Poles:  $\begin{array}{l} 0 \\ -1 \\ -4 \pm i5.3 \end{array}$   
all in LHS

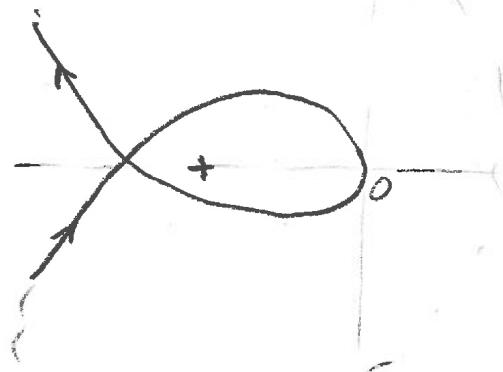


see plot

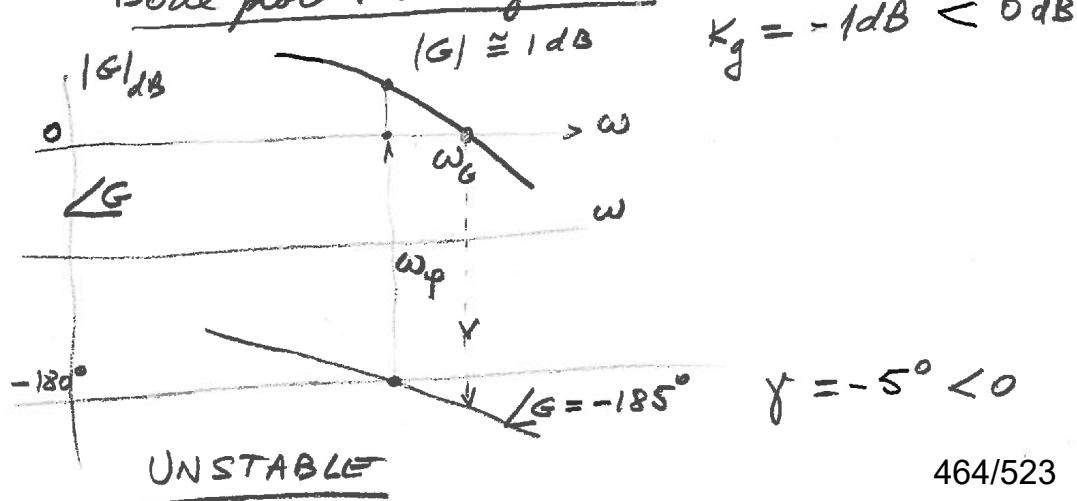
462/523



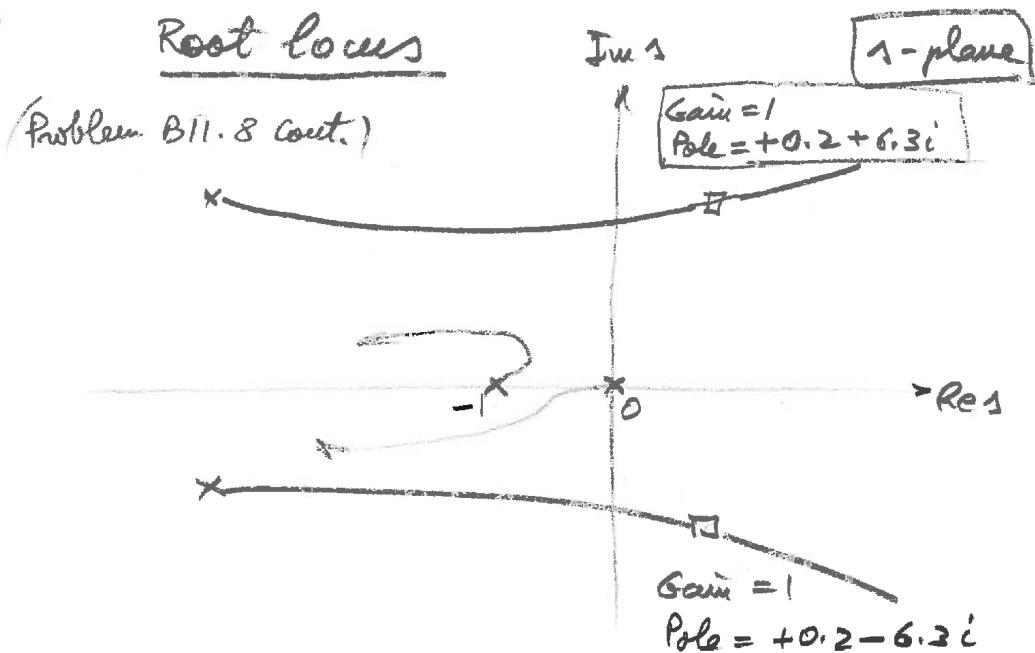
Problem B11.8 (cont.)

Nyquist plot $P = 0$  no poles in RHS $N = 2$ 
 $Z = P + N = 2$  zeros  
 in RHS of  $1 + G(s)$   
 because UNSTABLE

$$G_{CL} = \frac{G(s)}{1 + G(s)}$$

Bode plot : margins

464/523

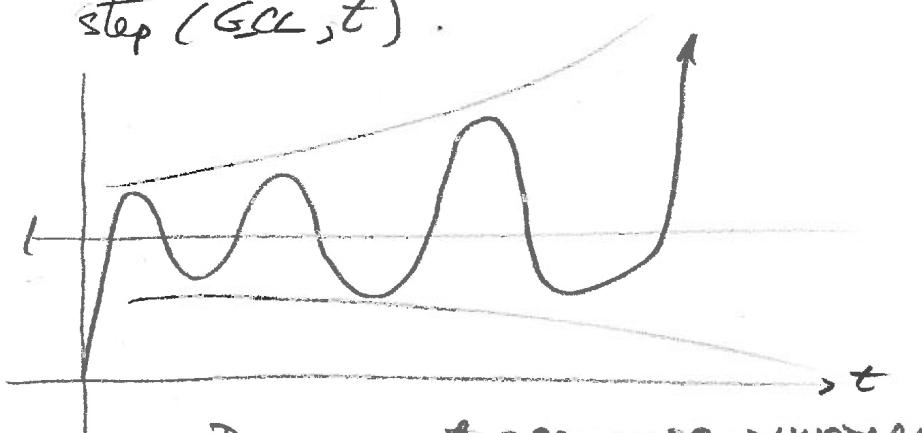


Roots in RHS  $\rightarrow$  UNSTABLE.

### Step response

$$G_{CL} = \text{feedback}(G, 1).$$

$$\text{step}(G_{CL}, t).$$



Divergent response  $\rightarrow$  UNSTABLE

All four criteria predicted an

UNSTABLE closed loop response

ACTION: DO NOT close the loop! 465/523

## 10 Control Systems Designer

The Control System Designer is a MATLAB SISO Tool.

CSD

Control System Designer (CSD) tool  
is an interactive GUI environment  
for analyzing and modifying feedback  
control systems.

- CSD was introduced in R2015a
- previous, we used "SISO Tool"

To start CSD, type in Command Window:

>> controlSystemDesigner

The legacy siso command is also  
recognized:

>> sisotool

Before starting CSD, run a MATLAB  
program to create the basic system

G(s).

- 
- Single input, single output (SISO)
  - Graphical User Interface (GUI)
  - Requires separate definition of plant transfer  
function G(s) through an m file

v  
CSD

---

Control System Designer

## Control System Designer

Design single-input, single-output (SISO) controllers

### Description

The **Control System Designer** app lets you design single-input, single-output (SISO) controllers for feedback systems modeled in MATLAB or Simulink (requires Simulink Control Design™ software).

Using this app, you can:

- Design controllers using:
  - Interactive Bode, root locus, and Nichols graphical editors for adding, modifying, and removing controller poles, zeros, and gains.
  - Automated PID, LQG, or IMC tuning.
  - Optimization-based tuning (requires Simulink Design Optimization™ software).
  - Automated loop shaping (requires Robust Control Toolbox™ software).
- Tune compensators for single-loop or multiloop control architectures.
- Analyze control system designs using time-domain and frequency-domain responses, such as step responses and pole-zero maps.
- Compare response plots for multiple control system designs.
- Design controllers for multimodel control applications.

### Open the Control System Designer App

- MATLAB Toolstrip: On the Apps tab, under **Control System Design and Analysis**, click the app icon.
- MATLAB command prompt: Enter `controlSystemDesigner`.
- Simulink model editor: Select **Analysis > Control Design > Control System Designer**.

### Examples

- “Control System Designer Tuning Methods”

2-135

468/523

3  
CSD

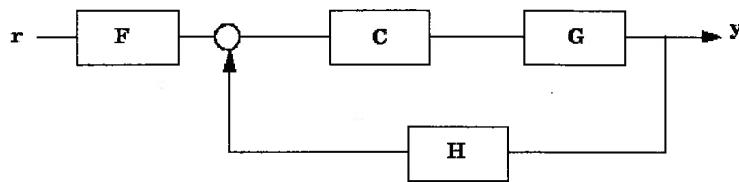
## 2 Functions — Alphabetical List

---

- "Bode Diagram Design"
- "Root Locus Design"
- "Design Compensator Using Automated Tuning Methods"
- "Design Multiloop Control System"
- "Analyze Designs Using Response Plots"
- "Compare Performance of Multiple Designs"
- "Multimodel Control Design"

### Programmatic Use

`controlSystemDesigner` opens the **Control System Designer** app using the following default control architecture:



The architecture consists of the LTI objects:

- *G* — Plant model
- *C* — Compensator
- *H* — Sensor model
- *F* — Prefilter

By default, the app configures each of these models as a unit gain.

`controlSystemDesigner(plant)` initializes the plant, *G*, to `plant`. `plant` can be any SISO LTI model created with `ss`, `tf`, `zpk` or `frd`, or an array of such models.

`controlSystemDesigner(plant,comp)` initializes the compensator, *C*, to the SISO LTI model `comp`.

u<sub>c</sub>s>

---

Control System Designer

`controlSystemDesigner(plant,comp,sensor)` initializes the sensor model,  $H$ , to `sensor`. `sensor` can be any SISO LTI model or an array of such models. If you specify both `plant` and `sensor` as LTI model arrays, the lengths of the arrays must match.

`controlSystemDesigner(plant,comp,sensor,prefilt)` initializes the prefilter model,  $F$ , to the SISO LTI model `prefilt`.

`controlSystemDesigner(vIEWS)` opens the app and specifies the initial graphical editor configuration. `vIEWS` can be any of the following character vectors, or a cell array of multiple character vectors.

- '`rlocus`' — Root locus editor
- '`bode`' — Open-loop Bode Editor
- '`nichols`' — Open-loop Nichols Editor
- '`filter`' — Bode Editor for the closed-loop response from prefilter input to the plant output

In addition to opening the specified graphical editors, the app plots the closed-loop, input-output step response.

`controlSystemDesigner(vIEWS,plant,comp,sensor,prefilt)` specifies the initial plot configuration and initializes the plant, compensator, sensor, and prefilter using the specified models. If a model is omitted, the app uses the default value.

`controlSystemDesigner(initData)` opens the app and initializes the system configuration using the initialization data structure `initdata`. To create `initdata`, use `sisoinit`.

`controlSystemDesigner(sessionFile)` opens the app and loads a previously saved session. `sessionFile` is the name of a session data file on the MATLAB path. This data includes the current system architecture and plot configuration, and any designs and responses saved in the **Data Browser**.

To save a session, in the **Control System Designer** app, on the **Control System** tab, click  **Save Session**.

**See Also**

**Apps**  
Control System Tuner

2-137

470/523

5  
c>>

**2 Functions — Alphabetical List**

---

**Functions**  
`pidTuner | sisoinit`

**Introduced in R2015a**

## **'Control System Designer' MATLAB Tool**

### **Aircraft Roll Motion Autopilot Development PD Example**

Table of Contents:

1	Control System Design Objectives .....	2
2	Create Plant Transfer Function G .....	2
3	Create 'Control System Designer' Model.....	3
3.1	Open 'Control System Designer' Software Tool .....	3
3.2	Import Plant Transfer Function G .....	4
3.3	Add Design Specifications .....	6
3.4	Store the Baseline Design as G .....	8
3.5	Record the Baseline Performance and Stability Reading of G Design.....	9
3.6	Verify Design Specifications .....	9
4	PID Controller using Automatic PID Tuning to Generate Design1 .....	10
4.1	Choose PD Controller and Adjust 'Response Time' Range .....	10
5	Evaluate Design1 Model.....	12
5.1	Stability Margins of Design2 Model .....	13
5.2	Performance Indicators of Design2 Model .....	13
6	Compare the step response of the initial and final designs.....	14

## 1 CONTROL SYSTEM DESIGN OBJECTIVES

$$\text{Consider the aircraft roll transfer function } G(s) = \frac{K}{Js^2 + cs} = \frac{114}{10s^2 + 4s}$$

Design a feedback control system to control the aircraft roll motion with the following control design objectives:

Design Objective 1: Control the unconstraint aircraft motion resulting from an aileron input. Have a autopilot system that can maintain the aircraft at a constant bank angle

We will achieve this objective through feedback (FB)

Design Objective 2: Achieve a reasonable aircraft roll response.

Design Objective 3: Ensure safe and stable operation of the feedback control system

We define ‘reasonable response’ using two control design specifications:

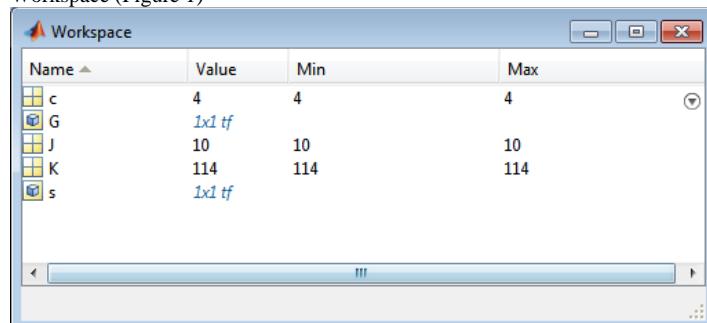
- DS1: Fast response time as measured by rise time  
 $t_r \leq 1.5 \text{ sec}$
- DS2: maximum percentage overshoot for step input less than 20%  
 $M_p \leq 20\%$

We ensue safe and stable operation of the feedback control system by meeting a third design specification based on stability margins:

- DS3:  $GM = 10 \text{ dB}$ ,  $PM = 60^\circ$

## 2 CREATE PLANT TRANSFER FUNCTION G

Run m-file “aircraft\_roll\_model.m” to create the plant transfer function G in the Workspace (Figure 1)



The screenshot shows the MATLAB workspace window with the following data:

Name	Value	Min	Max
c	4	4	4
G	<code>1x1 tf</code>		
J	10	10	10
K	114	114	114
s	<code>1x1 tf</code>		

Figure 1

### 3 CREATE ‘CONTROL SYSTEM DESIGNER’ MODEL

#### 3.1 OPEN ‘CONTROL SYSTEM DESIGNER’ SOFTWARE TOOL

In the Command Window, type “controlSystemDesigner”

>>controlSystemDesigner  
(legacy name “sisotool”, also works, i.e., >>sisotool )

The Control System Designer GUI opens as shown in Figure 2.

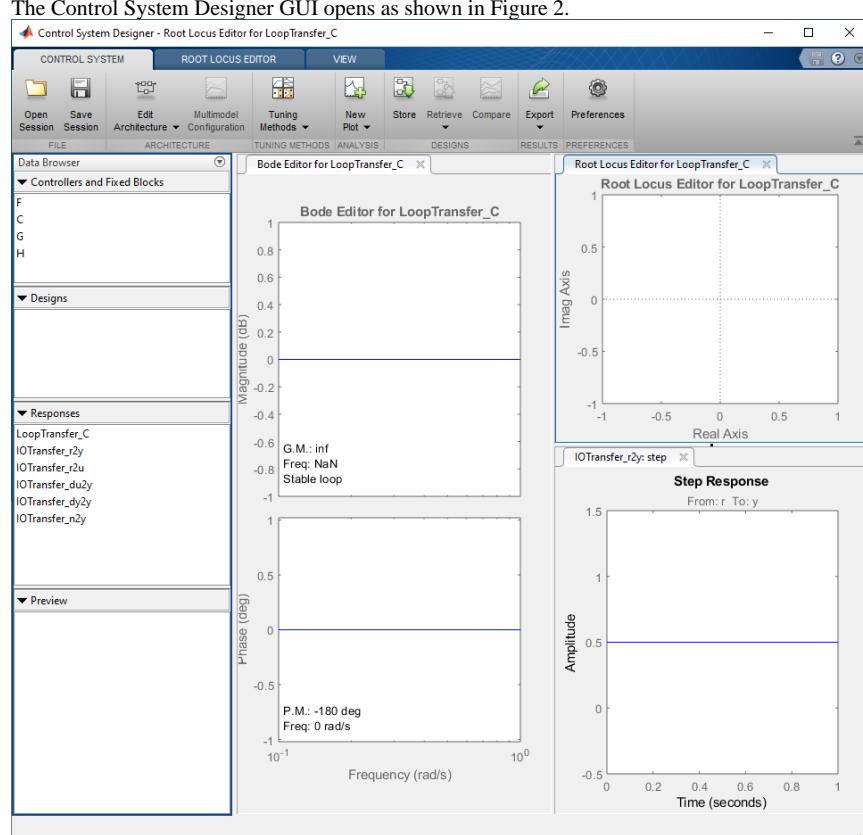


Figure 2

The default configuration has four blocks (F, C, G, H) as indicated in the LHS pane. All the blocks in this configuration are set to be unitary transfer functions by default (the plant block G, which is selected in the upper LHS pane, is shown in the lower LHS pane with Value: 1).

### 3.2 IMPORT PLANT TRANSFER FUNCTION G

Press “Edit Architecture” button to import the system G data from the Workspace (Figure 3).

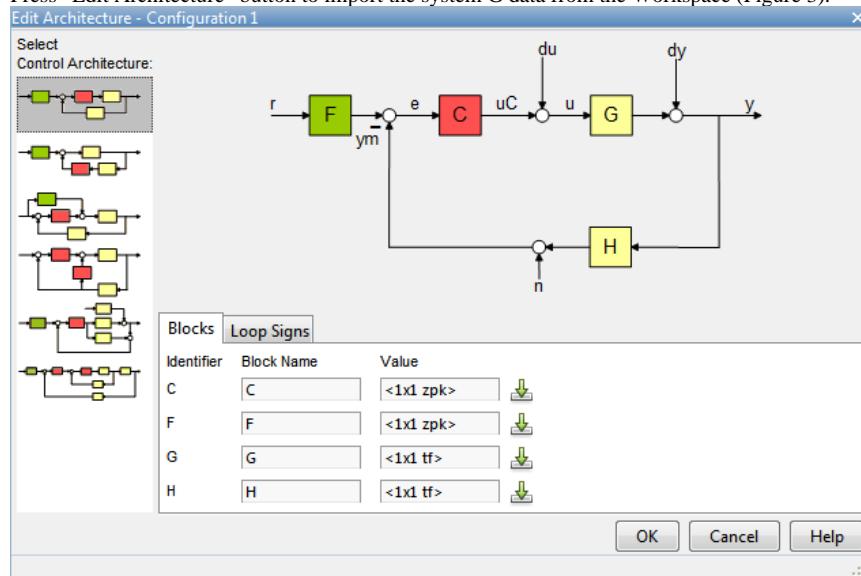


Figure 3

Note that the first control system configuration is selected by default (this is the configuration we work with, no need to change it.)

Press the down arrow icon at the RHS end of the G row to open the “Import Data for G” dialog box and select ‘G’ from the Base Workspace (Figure 4).

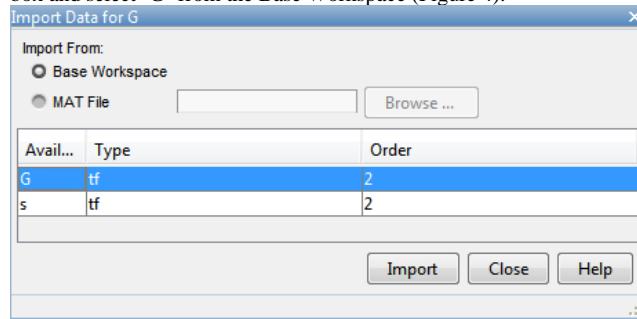


Figure 4

Press ‘Import’ to return to ‘Edit Architecture’ dialog box and press OK to return to the main view.

Now, our system  $G(s) = \frac{114}{10s^2 + 2s}$  is shown in the lower LHS pane of the model (Figure 5)

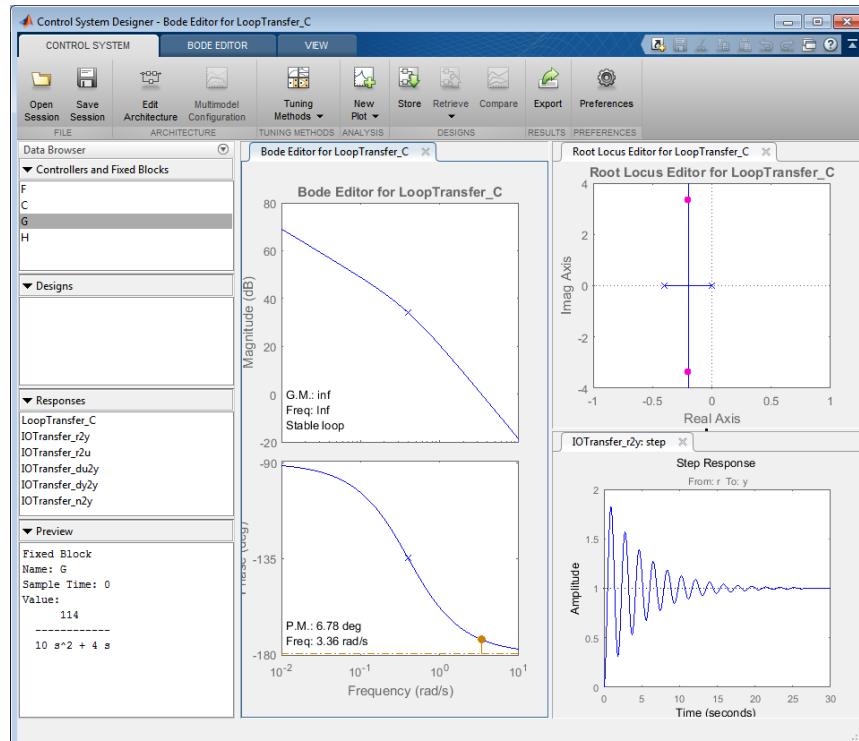


Figure 5

### 3.3 ADD DESIGN SPECIFICATIONS

Right-click on the Bode Editor window and select Design Requirements → New. The following window shown in Figure 6 appears:

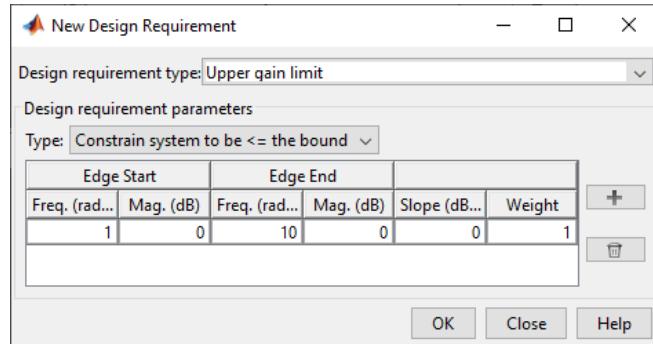


Figure 6

From the pull-down menu "Design requirement type" select "Gain & Phase margins" and enter the required DS3 values 10dB and 60deg to make the window look like shown in Figure 7.

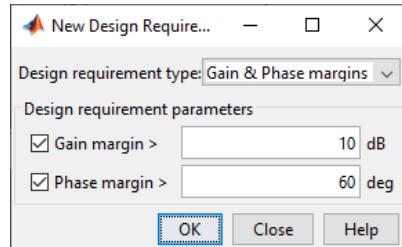


Figure 7

Press OK. Your screen should look like Figure 8.

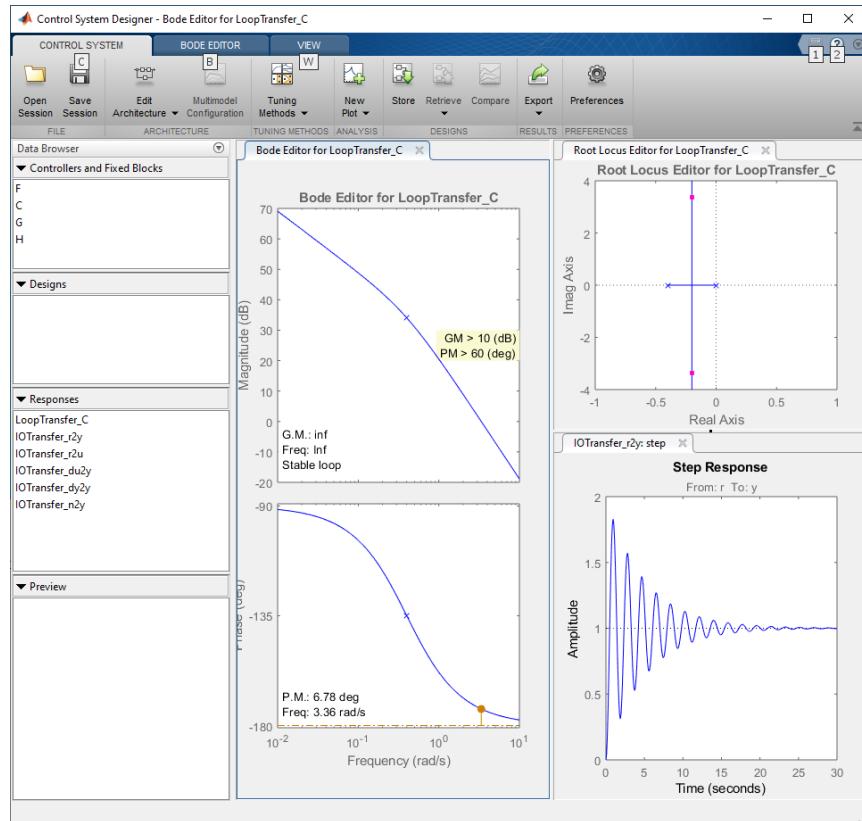


Figure 8

### 3.4 STORE THE BASELINE DESIGN AS G

Press ‘Store’ button to store this design; it will shown as ‘Design1’ in the second LHS pane. Save the CSD Session: press ‘Save Session’ button and save your session with the name ‘CSD\_aircraft\_Session\_20200116’. The extension of the file, if shown, should be ‘.mat’. Your screen should look like Figure 9.

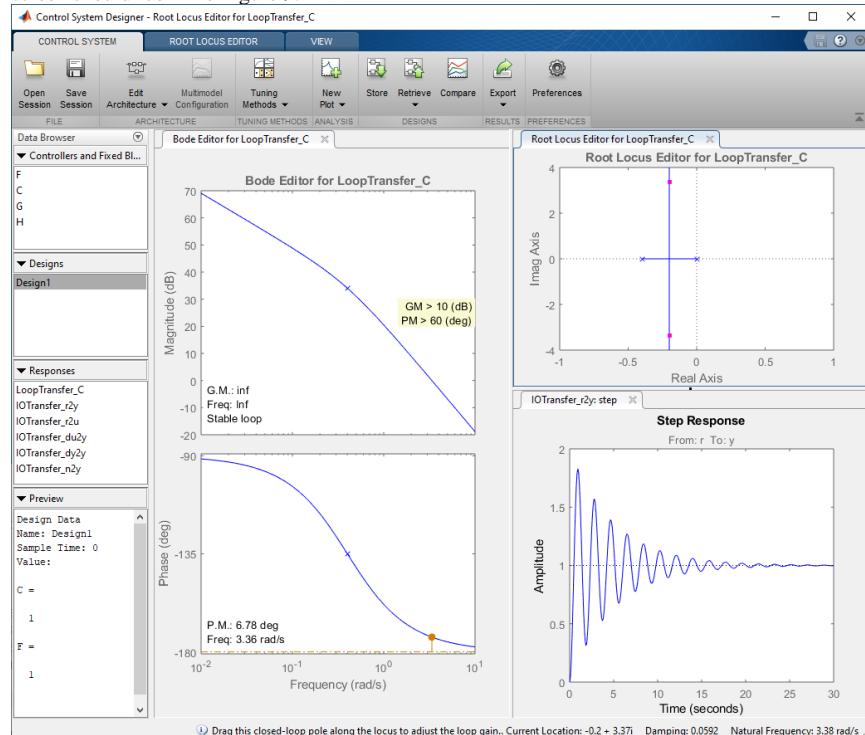


Figure 9

### 3.5 RECORD THE BASELINE PERFORMANCE AND STABILITY READING OF G DESIGN

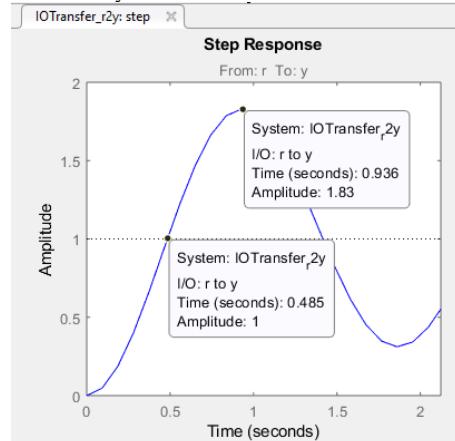
Record the following readings from this CSD windows:

- GM: inf
- PM: 6.78 deg at 3.36 rad/s
- CL poles  $-0.2 + 3.37i$  and  $-0.2 - 3.37i$  shown as the two red dots in the Root Locus Editor. Click on them and read the values at the bottom of the window.
- A lightly damped step response 'r2y: step' (i.e., from input r to output y) with a large overshoot

It is apparent that the response is unsatisfactory because:

- many oscillations until it settles down to  $x_{ss} = 1$
- large overshoot
- insufficient phase margin

Zoom into the 'r2y: step' window and use data tips to mark the rise time  $t_r$  and overshoot  $M_p$  and make the r2y window look as shown below



### 3.6 VERIFY DESIGN SPECIFICATIONS

It is apparent from the reading taken so far that:

- DS1:  $t_r = 0.485$  sec < 1.5 sec indicating that DS1 is satisfied
- DS2:  $M_p = 83\% > 20\%$  indicating that DS2 is NOT satisfied
- DS3: GM = inf; PM = 6.78 deg indicating that DS3 is NOT satisfied

A controller must be design to improve system performance and satisfy all the design specifications.

#### 4 PID CONTROLLER USING AUTOMATIC PID TUNING TO GENERATE DESIGN1

In this section we explain how a PD controller can be added and tuned to give a better system performance that satisfies all the design specifications.

Press ‘Tuning Methods’ and select ‘PID’ from the pull-down menu. The ‘PID Tuning’ window opens up (Figure 10).

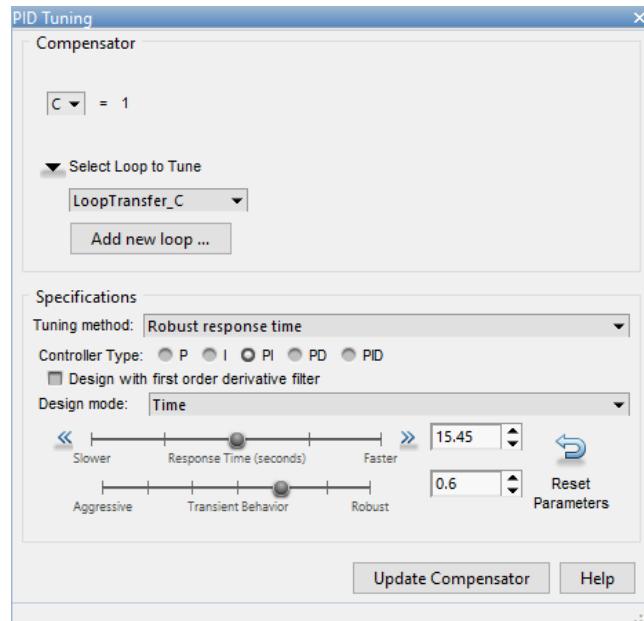


Figure 10

##### 4.1 CHOOSE PD CONTROLLER AND ADJUST ‘RESPONSE TIME’ RANGE

Select ‘PD’ on ‘Controller Type’ line. Press the >> arrow on the RHS of the ‘Response Time’ bar to see ‘1.545’ in the RHS window. Use the up/down scroll (or just type in the value) until it shows the value 1.5 (see Figure 11).

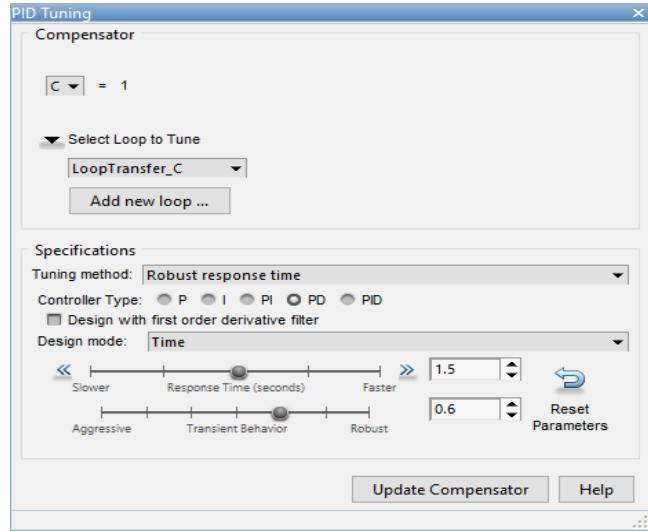


Figure 11

Press 'Update Compensator' button. The 'PID Tuning' window looks as shown in Figure 12.

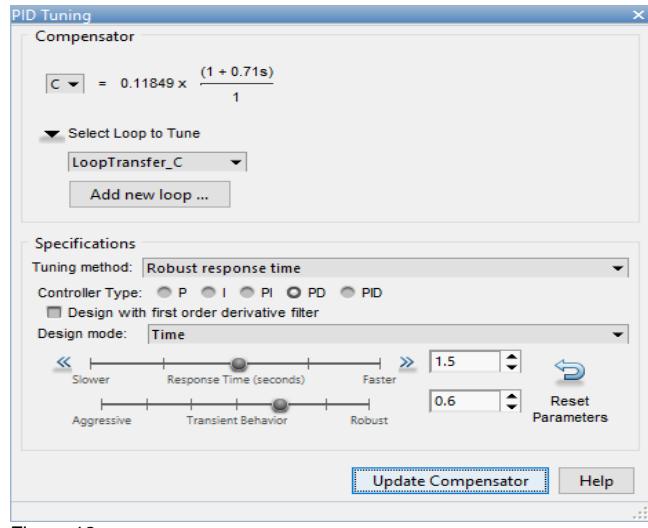


Figure 12

## 5 EVALUATE DESIGN1 MODEL

Close ‘PID Tuning’ widow and return to the ‘Control System Designer’ main window. Press ‘Store’. A new name appears in the ‘Designs’ LHS window; the new name is ‘Design2’. Restore the original view in the step response window using the zoom -out button. Now, the main window looks as shown in Figure 13.

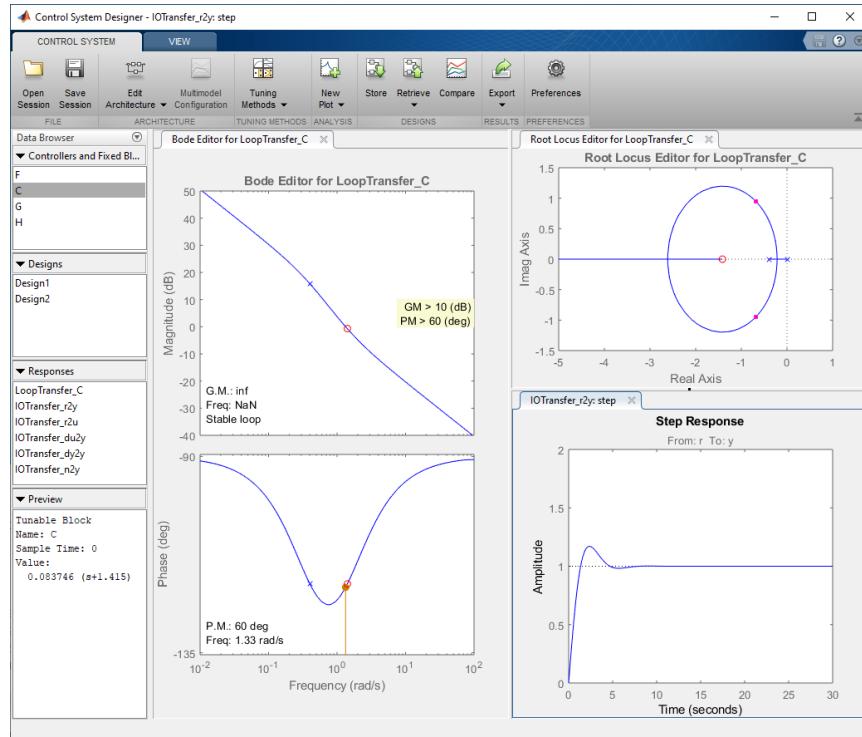


Figure 13

Note that the controller formula appear as

```
Tunable Block
Name: C
Sample Time: 0
Value:
0.083746 (s+1.415)
```

### 5.1 PERFORMANCE INDICATORS OF DESIGN2 MODEL

Next, verify the status of DS1 and DS2 conditions. Recall

- DS1: Fast response time as measured by rise time  
 $t_r \leq 1.5$  sec
- DS2: maximum percentage overshoot for step input less than 20%  
 $M_p \leq 20\%$

Use datatips to read  $t_r$  and  $M_p$  on the ‘Step Response’ plot for the closed loop system ‘IOTransfer\_r2y: step’ (Figure 14).

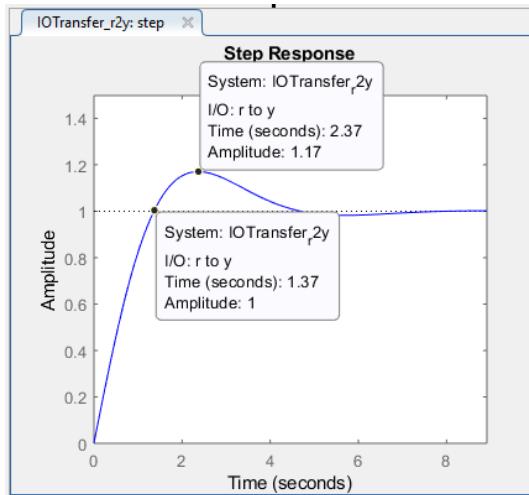


Figure 14

It is apparent that:

- $t_r = 1.37$  sec < 1.5 sec which means that DS1 is satisfied
- $M_p = 17\% < 20\%$  which means that DS2 is satisfied.

### 5.2 STABILITY MARGINS OF DESIGN2 MODEL

The stability margins can be read in the Bode plot of Figure 13; they are much better than in the original design:

- GM: inf which is better than  $GM = 10$  dB required by DS3
- PM: 60 deg, which meets the value  $PM = 60^\circ$  required by DS3

The stability margin criteria are satisfied which means that DS3 is satisfied.

We state that all three design specification DS1, DS2, DS3, have been met and the control system design process has completed successfully.

## 6 COMPARE THE STEP RESPONSE OF THE INITIAL AND FINAL DESIGNS

To compare the initial and final designs, do the following:

Press ‘Compare’ button and a ‘Compare Designs’ window pops up (Figure 15).

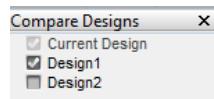


Figure 15

Check box in front of ‘Design1’. The step response plot contains the two responses overlapped (Figure 16).

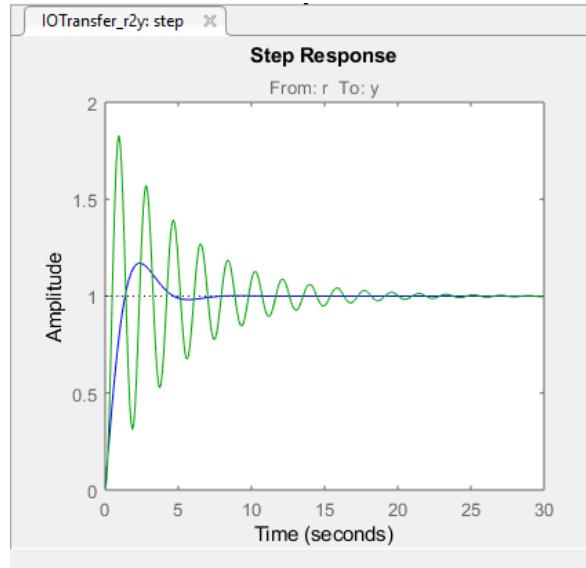


Figure 16

The initial design ‘Design1’ is shown in green, whereas the current design is shown in blue.

The response improvements achieved through this controller design process are quite apparent.

```
aircraft_roll_model.m × +  
1 %% description  
2 %{  
3 aircraft roll model for use with CSD Tool  
4 %}  
5 %% initialization  
6 - clc % clear command window  
7 - clear % clear workspace  
8 - % close all % close all plots  
9 - format compact  
10 - s=tf('s');  
11 - %% aircraft model  
12 - K=114;  
13 - J=10; % inertia  
14 - c=4; % damping  
15 - G=K/(J*s^2+c*s) % plant  
16 - controlSystemDesigner  
17
```

## **'Control System Designer' MATLAB Tool**

### **Aircraft Roll Motion Autopilot Development PID Example**

Table of Contents:

1	Control System Design Objectives .....	2
2	Create Plant Transfer Function G .....	2
3	Create 'Control System Designer' Model.....	3
3.1	Open 'Control System Designer' Software Tool .....	3
3.2	Import Plant Transfer Function G .....	4
3.3	Store the Baseline Design as G .....	6
3.4	Save the CSD Sesion.....	6
3.5	Record the Baseline Performance and Stability Reading of G Design .....	7
4	PID Controller using Automatic PID Tuning to Generate Design1 .....	7
4.1	Adjust 'Response Time' Range.....	8
4.2	Adjust the 'Time Response' and 'Transient Behavior' .....	9
5	Evaluate Design1 Model.....	10
5.1	Stability Margins of Design1 Model.....	10
5.2	Performance Indicators of Design 1 Model .....	11
6	Manual Adjustment of the Controller – Design2 Model .....	12
6.1	Grab and Move Compensator Zero to Improve Design.....	12
6.2	Adjust Compensator in Compensator Editor .....	13
6.3	Save the CSD Sesion.....	14
7	Evaluate Design2 Model.....	14
7.1	Stability Margins of Design2 Model.....	14
7.2	Performance Indicators of Design2 Model .....	15
8	Compare the step response of the initial and final designs.....	16

## 1 CONTROL SYSTEM DESIGN OBJECTIVES

$$\text{Consider the aircraft roll transfer function } G(s) = \frac{K}{Js^2 + cs} = \frac{114}{10s^2 + 4s}$$

Design a feedback control system to control the aircraft roll motion with the following control design objectives:

Design Objective 1: Control the unconstraint aircraft motion resulting from an aileron input. Have a autopilot system that can maintain the aircraft at a constant bank angle

We will achieve this objective through feedback (FB)

Design Objective 2: Achieve a reasonable aircraft roll response.

Design Objective 3: Ensure safe and stable operation of the feedback control system

We define ‘reasonable response’ using two control design specifications:

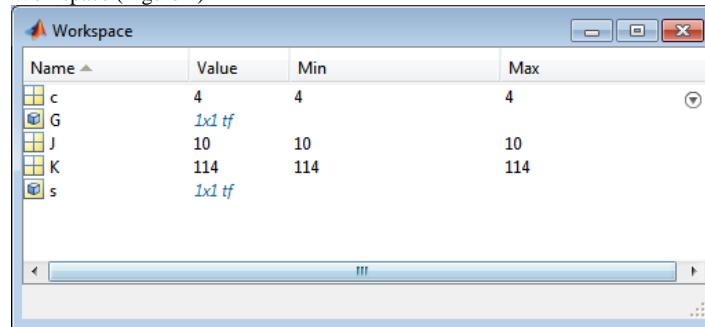
- DS1: Fast response time as measured by rise time  
 $t_r \leq 1.5 \text{ sec}$
- DS2: maximum percentage overshoot for step input less than 20%  
 $M_p \leq 20\%$

We ensure safe and stable operation of the feedback control system by meeting a third design specification based on stability margins:

- DS3:  $GM = 10 \text{ dB}$ ,  $PM = 60^\circ$

## 2 CREATE PLANT TRANSFER FUNCTION G

Run m-file “aircraft\_roll\_model.m” to create the plant transfer function G in the Workspace (Figure 1)



The screenshot shows the MATLAB workspace window with the following data:

Name	Value	Min	Max
c	4	4	4
G	1x1 tf		
J	10	10	10
K	114	114	114
s	1x1 tf		

Figure 1

### 3 CREATE ‘CONTROL SYSTEM DESIGNER’ MODEL

#### 3.1 OPEN ‘CONTROL SYSTEM DESIGNER’ SOFTWARE TOOL

In the Command Window, type “controlSystemDesigner”

```
>>controlSystemDesigner
```

(legacy name “sisotool”, also works, i.e., >>sisotool )

The Control System Designer GUI opens as shown in Figure 2.

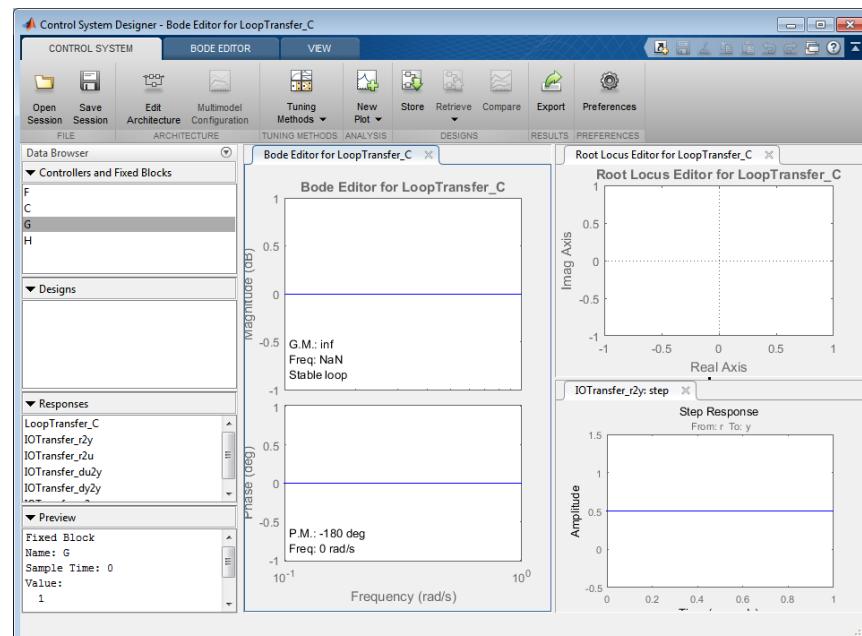


Figure 2

The default configuration has four blocks (F, C, G, H) as indicated in the LHS pane. All the blocks in this configuration are set to be unitary transfer functions by default (the plant block G, which is selected in the upper LHS pane, is shown in the lower LHS pane with Value: 1).

### 3.2 IMPORT PLANT TRANSFER FUNCTION G

Press “Edit Architecture” button to import the system G data from the Workspace (Figure 3).

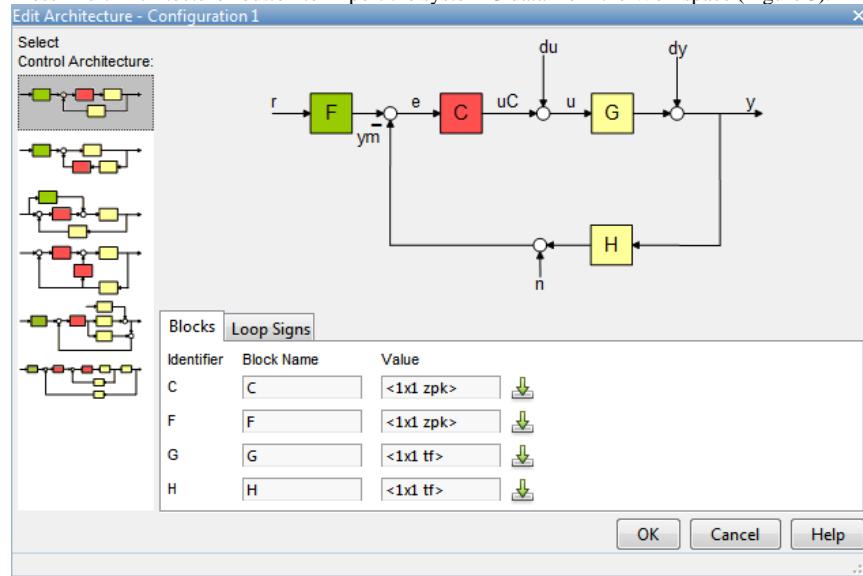


Figure 3

Note that the first control system configuration is selected by default (this is the configuration we work with, no need to change it.)

Press the down arrow icon at the RHS end of the row G to open the “Import Data for G” dialog box and select ‘G’ from the Base Workspace (Figure 4).

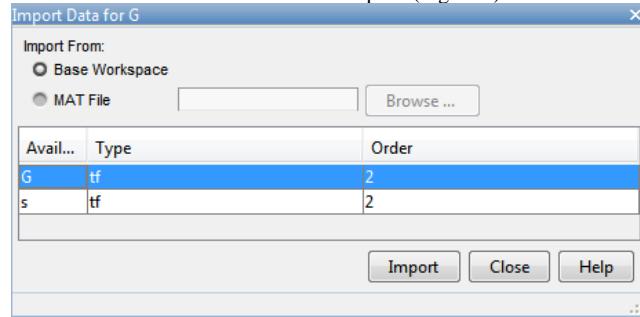


Figure 4

Press 'Import' to return to 'Edit Architecture' dialog box and press OK to return to the main view.

Now, our system  $G(s) = \frac{114}{10s^2 + 2s}$  is shown in the lower LHS pane of the model (Figure 5)

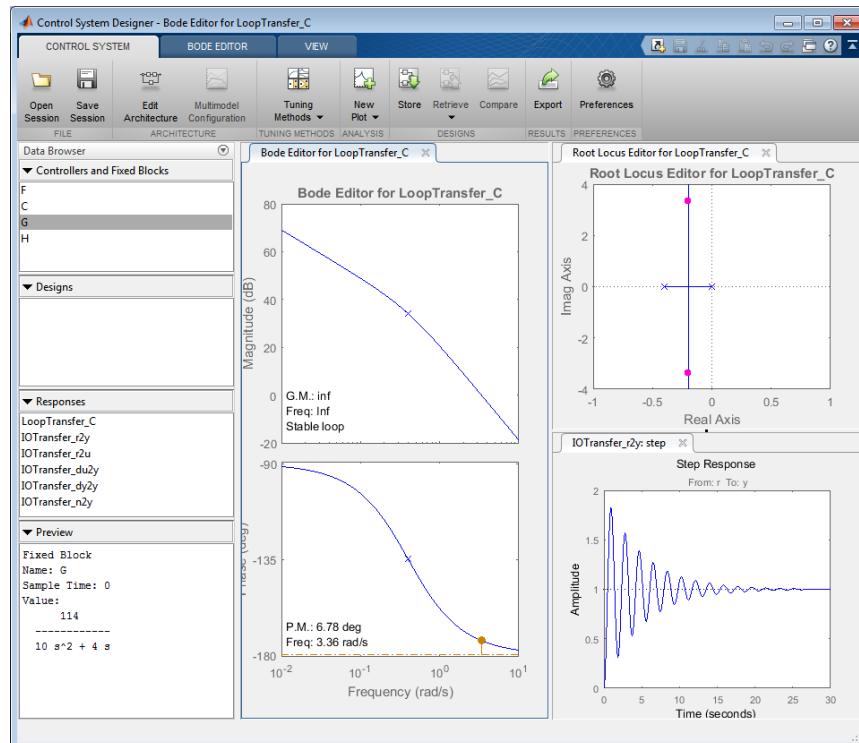


Figure 5

### 3.3 STORE THE BASELINE DESIGN AS G

Press ‘Store’ button to store this design; it will shown as ‘Design1’ in the second LHS pane; double click on it to allow name edit and rename it as ‘G’. Your screen should look like Figure 6.

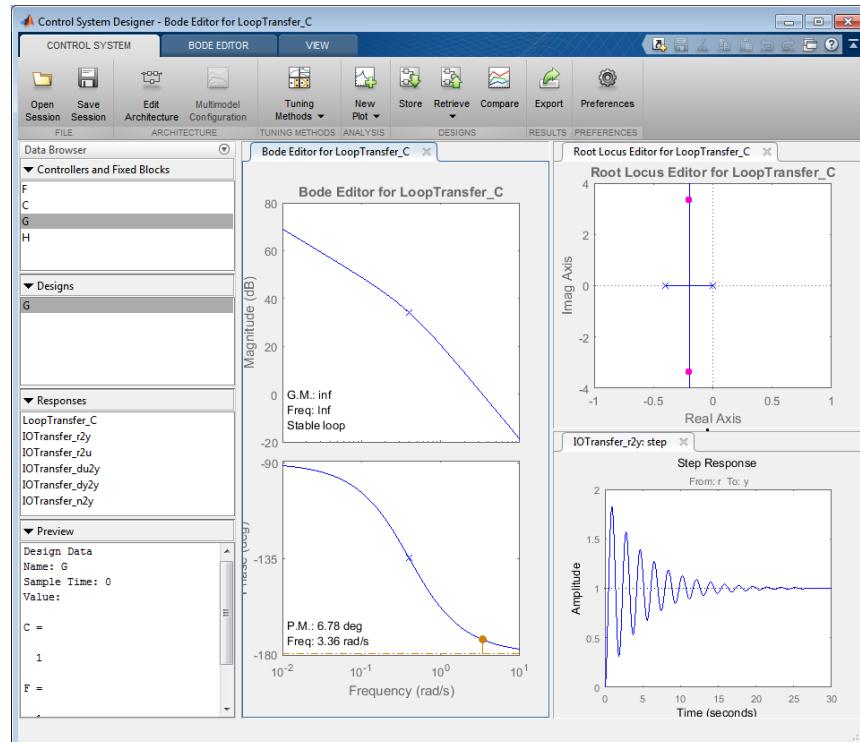


Figure 6

### 3.4 SAVE THE CSD SESSION

Press ‘Save Session’ button and save your session with the name ‘CSD\_aircraft\_Session\_20161219’. The extension of the file, if shown, should be ‘.mat’.

### 3.5 RECORD THE BASELINE PERFORMANCE AND STABILITY READING OF G DESIGN

Record the following readings from this window:

- GM: inf
- PM: 6.78 deg at 3.36 rad/s
- CL poles  $-0.2 + 3.37i$  and  $-0.2 - 3.37i$  shown as the two red dots in the Root Locus Editor
- A lightly damped step response 'r2y' (i.e., from input r to output y) with a large overshoot

It is apparent that the response is unsatisfactory because:

- many oscillations until it settles down to  $x_{ss} = 1$
- large overshoot ( $x_p \approx 1.8$ ,  $M_p \approx 80\%$ )
- insufficient phase margin

### 4 PID CONTROLLER USING AUTOMATIC PID TUNING TO GENERATE DESIGN1

Press 'Tuning Methods' and select 'PID' from the pull-down menu. The 'PID Tuning' window opens up (Figure 7).

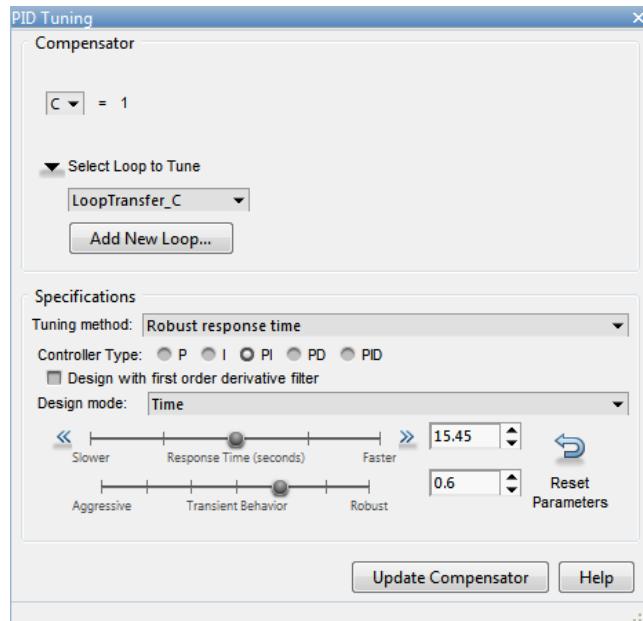


Figure 7

## 4.1 ADJUST ‘RESPONSE TIME’ RANGE

Select ‘PID’ on ‘Controller Type’ line. Press the >> arrow on the RHS of the ‘Response Time’ bar to see ‘1.545’ in the RHS window (Figure 8).

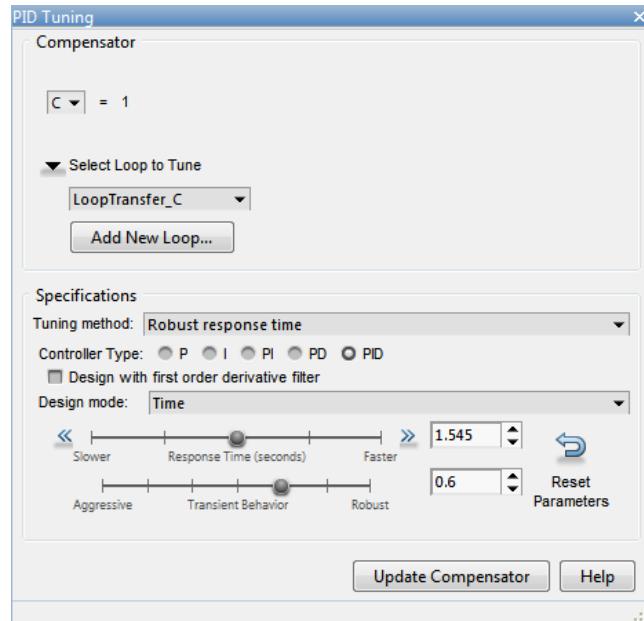


Figure 8

#### 4.2 ADJUST THE ‘TIME RESPONSE’ AND ‘TRANSIENT BEHAVIOR’

Use the up/down arrows to make the ‘Time Response’ reading in the RHS window ‘1.422’ and the ‘Transient Behavior’ reading in the RHS window be ‘0.6’.

Press ‘Update Compensator’button. The ‘PID Tuning’ window looks as shown in Figure 9.

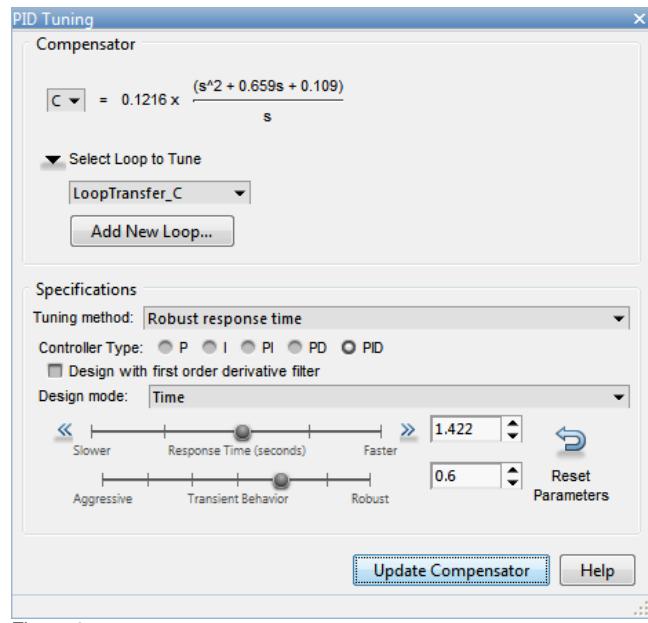


Figure 9

## 5 EVALUATE DESIGN1 MODEL

Close ‘PID Tuning’ widow and return to the ‘Control System Designer’ main window. Press ‘Store’. A new name appears in the ‘Designs’ LHS window; the new name is ‘Design1’. Now, the main window looks as shown in Figure 10.

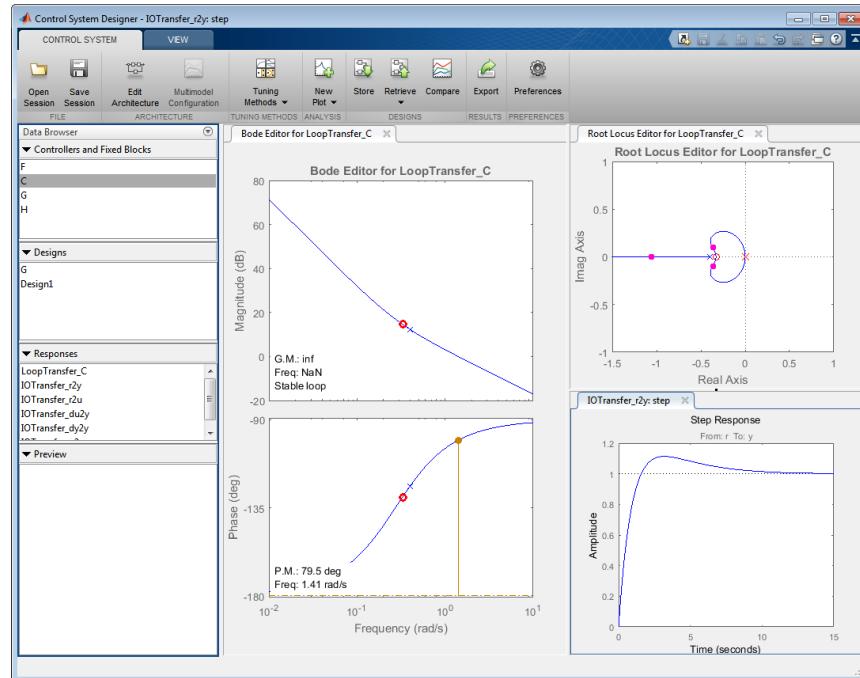


Figure 10

Note that the compensator formula appear as

$$C = \frac{0.11236 (s+0.3118)^2}{s}$$

### 5.1 STABILITY MARGINS OF DESIGN1 MODEL

The stability margins can be read in the Bode plot; they are much better:

- GM: inf which is better than  $GM = 10\text{ dB}$  required by DS3
- PM: 79.5 deg, which is better tan ,  $PM = 60^\circ$  required by DS3

The stability margin criteria are satisfied and hence we can say that DS3 condition has been met.

## 5.2 PERFORMANCE INDICATORS OF DESIGN 1 MODEL

Next, verify the status of DS1 and DS2 conditions. Recall

- DS1: Fast response time as measured by rise time  
 $t_r \leq 1.5$  sec
- DS2: maximum percentage overshoot for step input less than 20%  
 $M_p \leq 20\%$

We use datatips to read  $t_r$  and  $M_p$  on the 'Step Response' plot for the closed loop system 'IOTransfer\_r2y' (Figure 11).

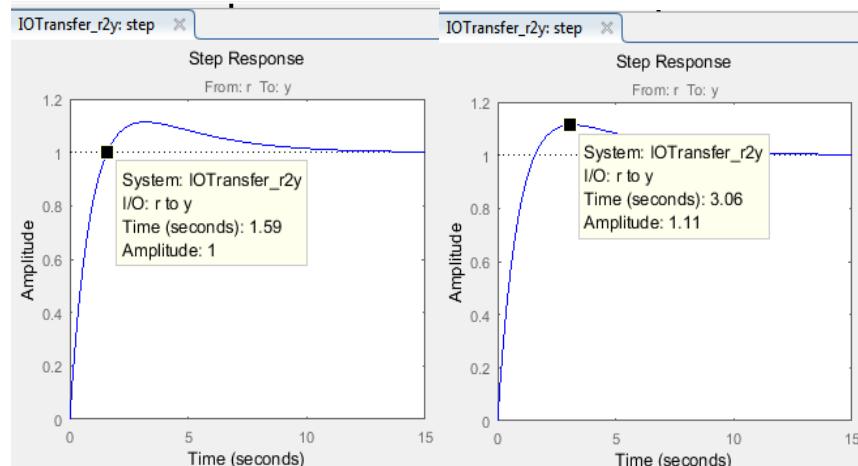


Figure 11

It is apparent that  $t_r = 1.59$  sec ,  $M_p = 11\%$

It seems that DS2 has been also satisfied by DS1 is not yet fully satisfied (though it is very close to it).

## 6 MANUAL ADJUSTMENT OF THE CONTROLLER – DESIGN2 MODEL

To further improve the system performance and meet the design specifications, we perform manual tuning in the Bode diagram.

### 6.1 GRAB AND MOVE COMPENSATOR ZERO TO IMPROVE DESIGN

Grab the compensator Zero shown by a red circle on the Bode Phase plot (Figure 12).

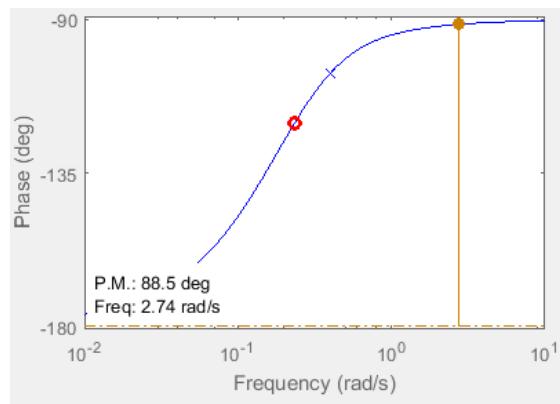


Figure 12

Move this red circle slightly to improve the response time  $t_r$  while maintaining GM: inf and a large PM.

This manual adjustment operation must be performed with great delicacy and in very small steps. Continuous monitoring of the changes in stability margin should be done continuously. If stability is lost, than it should be restored immediately through a backward step and fine adjustment should be continued with delicacy. Note that the gain margin GM is prone of jumping from 'inf', which is OK, to a large negative value which is not OK!

When a satisfactory situation has been met, press 'Store'; a new design named 'Design2' appears in the 'Designs' LHS pane. The overall picture is shown in Figure 13.

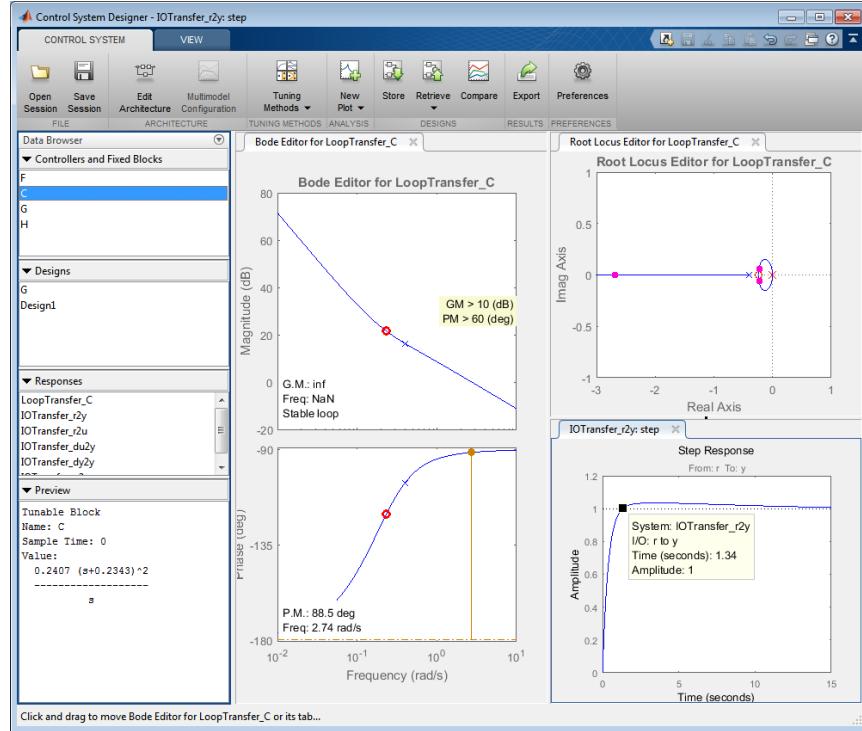


Figure 13

## 6.2 ADJUST COMPENSATOR IN COMPENSATOR EDITOR

One can also adjust the compensator directly by entering values for its properties. Right click on 'C' in the upper LHS pane and choose 'Open Selection' from the pulldown menu. The 'Compensator Editor' window opens as shown in Figure 14. The compensator properties, e.g., the 'Real Part' of the Complex Zero can be entered here.

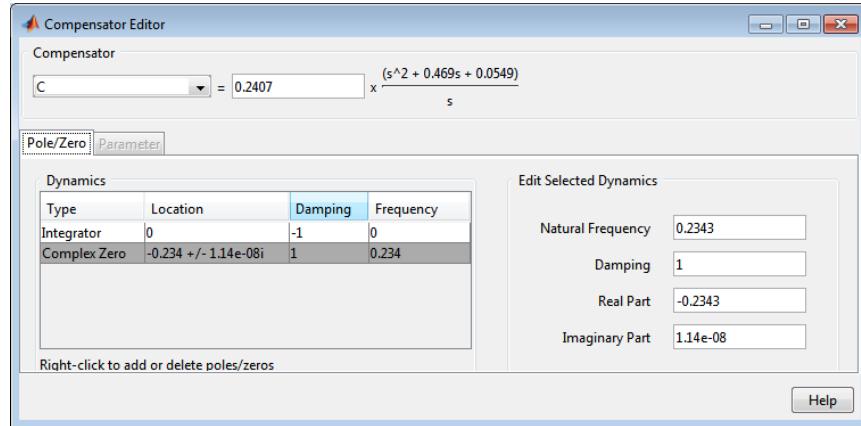


Figure 14

### 6.3 SAVE THE CSD SESSION

To save the latest designs, press the ‘Save Session’ button and save your session with the name ‘CSD\_aircraft\_Session\_20161219’. The file extension, if shown, should be ‘.mat’.

## 7 EVALUATE DESIGN2 MODEL

### 7.1 STABILITY MARGINS OF DESIGN2 MODEL

The stability margins can be read on the Bode plots of Figure 13; they are very good:

- GM: inf which is better than  $GM = 10\text{ dB}$  required by DS3
- PM:  $88.5\text{ deg}$ , which is better than  $PM = 60^\circ$  required by DS3

The stability margin criteria are satisfied and DS3 condition has been met.

## 7.2 PERFORMANCE INDICATORS OF DESIGN2 MODEL

Figure 13 indicates that the step response has a faster response time and a smaller overshoot than before (Figure 15).

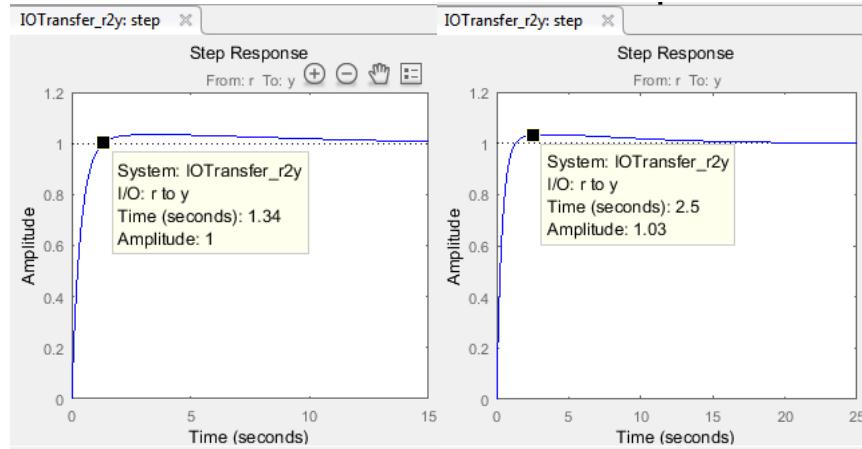


Figure 15

From Figure 15, we read:  $t_r = 1.34 \text{ sec}$ ,  $M_p = 3\%$ .

It seems that DS1 and DS2 have been met.

The only drawback this this Design2 situation could be that the system seems to take longer than before to settle down to the steady state value.

Further tweaking of the controller could be attempted to overcome this aspect if considered important.

However, for now, we state that all three design specification DS1, DS2, DS3 have been met and the control system design process can be considered complete.

## 8 COMPARE THE STEP RESPONSE OF THE INITIAL AND FINAL DESIGNS

To compare the initial and final designs, do the following:

Press ‘Compare’ button and a ‘Compare Designs’ window pops up (Figure 16).

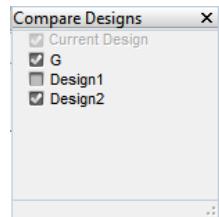


Figure 16

Check boxes in front of ‘G’ and ‘Design2’. The step response plot contains the two responses overlapped (Figure 17).

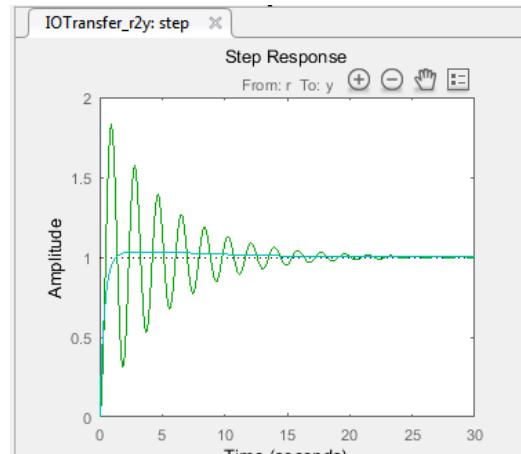


Figure 17

The initial design ‘G’ is shown in green, whereas the final design ‘Design2’ is shown in blue.

The response improvements achieved through this controller design process are quite apparent.

## 12 State-Space Representations

ss (def)

## STATE-SPACE REPRESENTATION

State-space representation is a time domain model of the form:

$$\frac{d}{dt} z = Az + Bu \quad (\text{dynamic system}) \quad (1a)$$

$$y = Cz + Du \quad (\text{output}) \quad (1b)$$

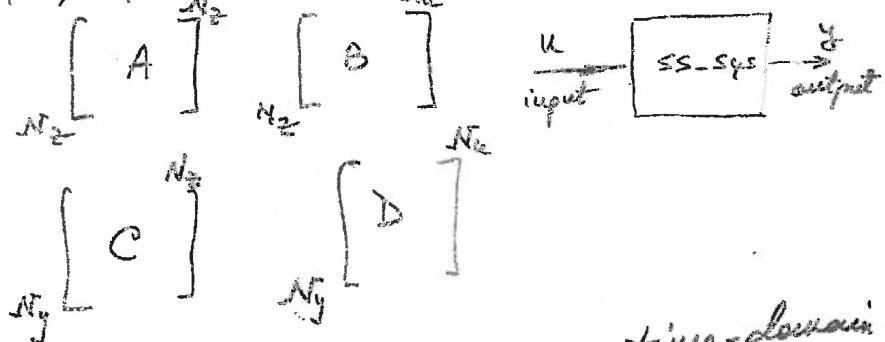
The variables involved in Eq.(1) are:

$z$  = column of state variables, length  $N_z$

$u$  = column of input variables, —  $N_u$

$y$  = column of output variables —  $N_y$

$A, B, C, D$  = state space matrices



The SS representation is a time-domain representation, whereas the TF representation was a Laplace-domain representation.

MATLAB accepts both TF and SS representations.

(a)  
ss

1<sup>st</sup> order system  
State-space representation

Recall

$$T \ddot{x}(t) + x(t) = f(t) \quad (1)$$

Rewrite (1) as :

$$\dot{z} = -\frac{1}{T} z + f$$

or

$$\frac{d}{dt} z = -\frac{1}{T} z + \frac{1}{T} f \quad (2)$$

$$\begin{matrix} \uparrow & \uparrow & \uparrow & \uparrow \\ z & A & z & B & f \end{matrix}$$

$$\left\{ \begin{array}{l} \frac{d}{dt} z = Az + Bu \\ y = Cz + Du \end{array} \right. \quad \left\{ \begin{array}{l} z = x \\ u = f \\ y = x \\ A = -1/T \\ B = 1/T \\ C = 1 \\ D = 0 \end{array} \right.$$

## 2nd Order System

state-space representation

$\overset{?}{\text{SS def}}$  To derive the SS representation of a 2nd order system, recall its 2nd order ordinary differential equation in standard form, i.e.,

$$\ddot{x} + 2\zeta\omega_n \dot{x} + \omega_n^2 x = \omega_n^2 f(t) \quad (2)$$

Write (2) in terms of  $x$  and  $\dot{x}$  only, i.e.,

$$\frac{d}{dt} \dot{x} + 2\zeta\omega_n \dot{x} + \omega_n^2 x = \omega_n^2 f(t) \quad (3)$$

Add the identity  $\frac{d}{dt}x = \dot{x}$  and write (3) as an extended 1st order system, i.e.,

$$\begin{cases} \frac{d}{dt}x = \dot{x} \\ \frac{d}{dt}\dot{x} = -2\zeta\omega_n \dot{x} - \omega_n^2 x + \omega_n^2 f(t) \end{cases} \quad (4)$$

Express (4) in matrix form, i.e.,

$$\frac{d}{dt} \begin{bmatrix} x \\ \dot{x} \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ -\omega_n^2 & -2\zeta\omega_n \end{bmatrix} \begin{bmatrix} x \\ \dot{x} \end{bmatrix} + \begin{bmatrix} 0 \\ \omega_n^2 \end{bmatrix} f(t) \quad (\text{dynamic system}) \quad (5)$$

$$y = [x] \quad (\text{output})$$

ss1dt Define :

$$\begin{aligned} z &= \begin{bmatrix} x \\ \dot{x} \end{bmatrix}, \quad u = [f(t)] \quad y = [x] \\ N_z &= 2 \quad N_u = 1 \quad N_y = 1 \end{aligned}$$

$$A = \begin{bmatrix} 0 & 1 \\ -\omega_n^2 & -2j\omega_n \end{bmatrix} \quad B = \begin{bmatrix} 0 \\ \omega_n^2 \end{bmatrix} \quad (6)$$

$$C = [1 \ 0] \quad D = [0]$$

(6)  $\rightarrow$  (5) :

$$\left\{ \begin{array}{l} \frac{d}{dt}z = Az + Bu \\ y = Cz + Du \end{array} \right. \quad (7)$$

Eqs. (6), (7) form the SS representation of the dynamic system (2). The TF representation of the same system is

$$G(j\omega) = \frac{\omega_n^2}{s^2 + 2j\omega_n s + \omega_n^2}$$

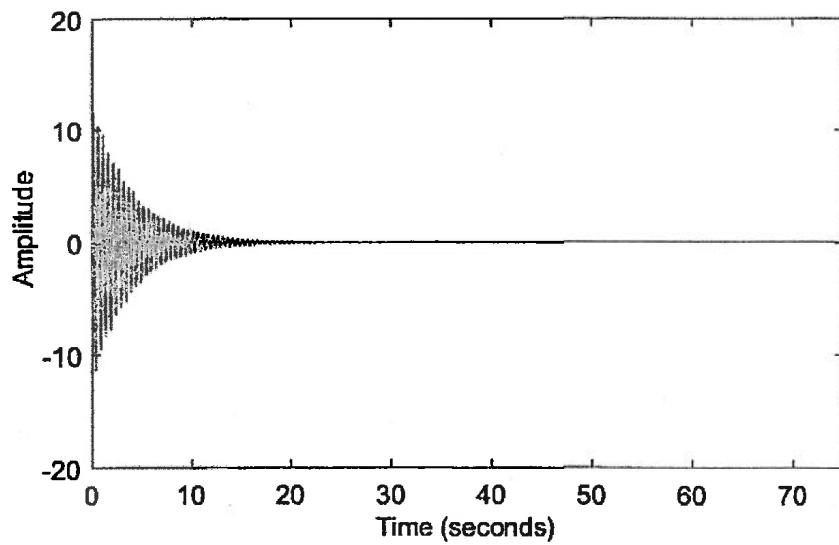
The TF and SS representations are interchangeable in MATLAB

4  
SS

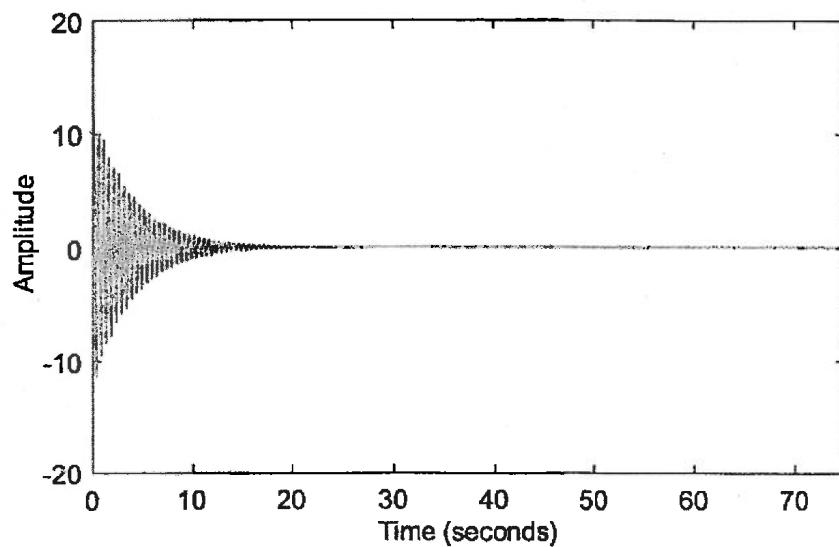
```
Comparison of TF and SS representations
initial data: fn, Hz; z% =
    2      2
TF poles =
-0.2513 +12.5639i
-0.2513 -12.5639i
f_TF, Hz; z_TF% =
1.9996   2.0000
1.9996   2.0000
SS poles =
-0.2513 +12.5639i
-0.2513 -12.5639i
f_SS, Hz; z_SS% =
1.9996   2.0000
1.9996   2.0000
```

5  
ss/det

impulse response of TF system: z TF= 2%



impulse response of SS system: z SS= 2%



G:\EMCH516\EMCH516\_2...\\single\_dof\_time\_response\_TF\_SS.m      Page 1

```
1 %% DESCRIPTION
2 %
3 SISO system time response
4 Comparison of TF and SS representations
5 %
6 %% initialization
7 opengl hardwarebasic % switch to basic hardware graphics functions
8 clc % clear command window
9 clear % clear workspace
10 % close all % close all plots
11 format compact
12 tol=1e-10; % tolerance for discarding machine zero
13 %% DEFINE PARAMETERS
14 % ----- data for class -----
15 fn=2; wn=2*pi*fn; % plunge frequency, Hz
16 z=2e-2; % plunge damping ratio
17 %% DEFINE TIME RANGE
18 T=1/fn; % time scale
19 % dt=1e-3;
20 % tmax=50*T;
21 tmax=150*T;
22 Nt=1000; dt=tmax/Nt;
23 t=0:dt:tmax; % time range
24 %% CALCULATE SISO TRANSFER MATRIX
25 G=tf([wn^2],[1 2*z*wn wn^2]);
26 %% CALCULATE POLES OF G
27 [~,poles_TF,~]=zpkdata(G); s_TF=poles_TF{1,1}; % poles
28 % display(s_TF,'poles of G')
29 f_TF=abs(imag(s_TF)/(2*pi)); % frequencies
30 zz=-real(s_TF)./abs(s_TF); z_TF=zz.*((abs(zz)>tol)); % damping
31 %% CALCULATE IMPULSE RESPONSE OF TF SYSTEM
32 figure(1);
33 subplot(2,1,1); impulse(G,t);
34 title(['impulse response of TF system: z_TF= ' num2str(z_TF(1)*100) '%'])
35 %% DISPLAY TF RESULTS
36 display('Comparison of TF and SS representations')
37 display([fn z*100], 'initial data: fn, Hz; z');
38 display(s_TF, 'TF poles');
39 display([f_TF z_TF*100], 'f_TF, Hz; z_TF');
40 %% CALCULATE SISO STATE SPACE MODEL
41 %----- state space matrices -----
42 A=[ 0 1 ;
43 -wn^2 -2*z*wn];
44 B=[ 0 ;
45 wn^2];
46 C=[1 0];
47 D=[0];
48 ss_sys=ss(A,B,C,D);
49 %% EXTRACT POLES, FREQUENCY, DAMPING
50 [~,zz,poles_SS]=damp(ss_sys);
```

G:\EMCH516\EMCH516\_2...\\single\_dof\_time\_response\_TF\_SS.m      Page 2

```
51 % display(ss,'poles of ss_sys')
52 s_SS=poles_SS;                                       % poles
53 f_SS=abs(imag(poles_SS))/(2*pi);               % frequencies
54 z_SS=zz.*((abs(zz)>tol);                       % damping
55 %% DISPLAY SS RESULTS
56 % display(' ');
57 display(s_SS, 'SS poles');
58 display([f_SS z_SS*100], 'f_SS, Hz; z_SS');
59 %% CALCULATE IMPULSE RESPONSE OF SS SYSTEM
60 subplot(2,1,2); impulse(ss_sys,t);
61 title(['impulse response of SS system: z_SS= ' num2str(z_SS(1)*100) '%'])
62
63
64
65
66
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```

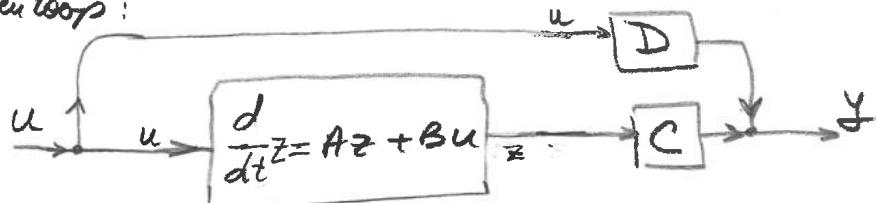
510/523

<sup>2018 10/24</sup> FB of SS system

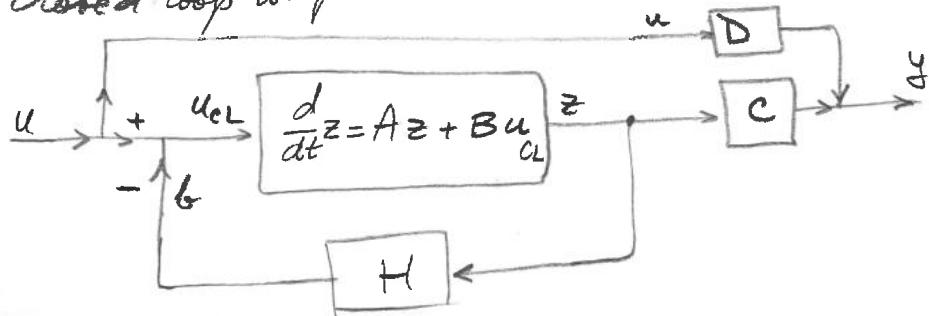
$$\frac{d}{dt} z = Az + Bu \quad (1a)$$

$$y = Cz + Du \quad (1b)$$

open loops:



closed loop w. feedback H:



$$u_{CL} = u - b = u - Hz$$

$$\frac{d}{dt} z = Az + Bu_{CL} = Az + Bu - BHz$$

$$\frac{d}{dt} z = (A - BH)z + Bu$$

$$\boxed{\begin{array}{l} A_{CL} \\ = A - BH \end{array}}$$

same  $A_{CL}$  as for  
simplified model

$$\left\{ \begin{array}{l} \frac{d}{dt} z = A_{CL}z + Bu \quad \text{ss sys} \\ y = Cz + Du \quad \text{w. feedback } H \end{array} \right.$$

2  
2016.08.01

## H for velocity feedback

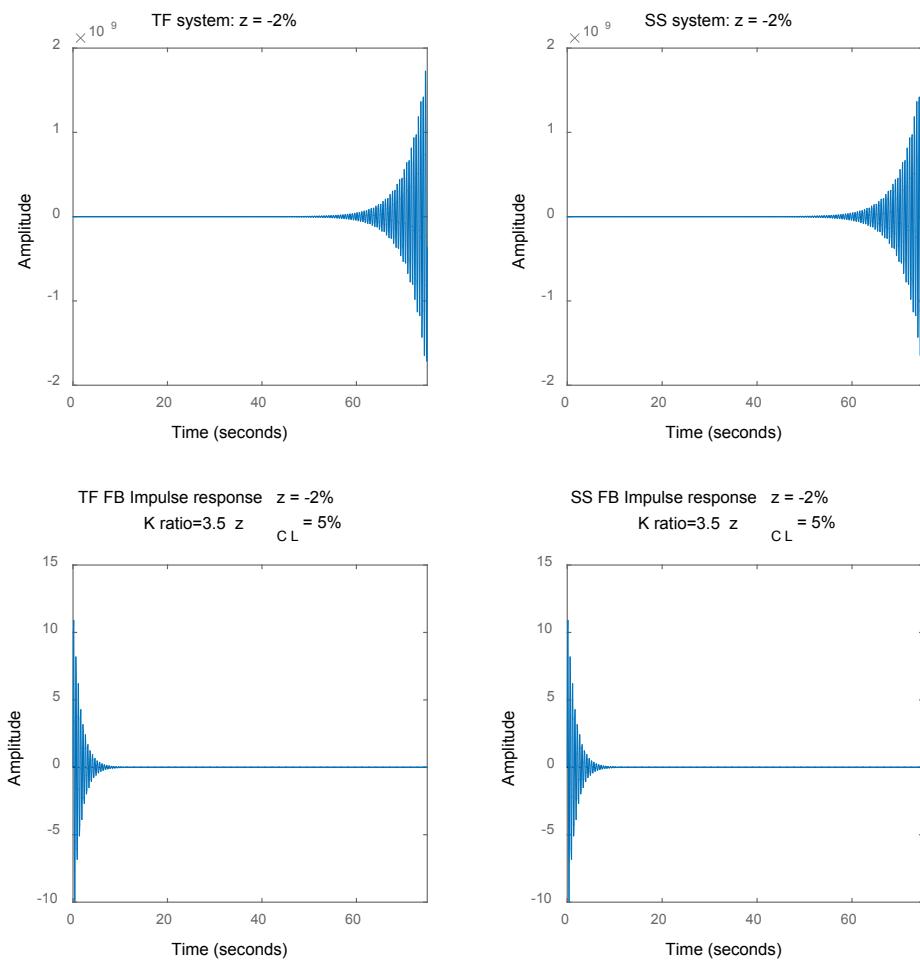
$$\text{Recall } z = \begin{bmatrix} x \\ \dot{x} \end{bmatrix} \quad (6)$$

For velocity feedback use

$$H = \begin{bmatrix} 0 & K \end{bmatrix} \quad (7)$$

$$\text{The } b(t) = H z(t) = \begin{bmatrix} 0 & K \end{bmatrix} \begin{bmatrix} x(t) \\ \dot{x}(t) \end{bmatrix} = K \dot{x}(t) \quad (8)$$

feedback signal  $b(t)$   
is proportional to velocity  $\dot{x}(t)$ .



513/523

```
1 %% DESCRIPTION
2 %{
3 SISO system time response with feedback
4 Comparison of TF and SS representations
5 use feedback function in TF
6 use direct definition of A_CL in SS
7 %}
8 %% initialization
9 opengl hardwarebasic % switch to basic hardware graphics functions
10 clc % clear command window
11 clear % clear workspace
12 % close all % close all plots
13 format compact;
14 tol=1e-10; % tolerance for discarding machine zero
15 %% DEFINE PARAMETERS
16 % ---- data for example ----
17 % ---- data for HW ----
18 m=2; % mass, kg
19 fn=2; wn=2*pi*fn; % plunge frequency, Hz
20 z=-2e-2; % plunge damping ratio
21 z_ratio=2.5; % desired z_ratio defined as zCL/|z|
22 display('HW06a')
23 display([fn z*100], 'initial data: fn, Hz; z%');
24 %% CALCULATE critical FB gain
25 K_cr=2*abs(z)/wn; % critical FB gain
26 %% DEFINE TIME RANGE
27 T=1/fn; % time scale
28 % dt=1e-3;
29 % tmax=50*T;
30 tmax=150*T;
31 Nt=1000; dt=tmax/Nt;
32 t=0:dt:tmax; % time range
33 %% CALCULATE SISO TRANSFER MATRIX
34 G=tf(wn^2,[1 2*z*wn wn^2]); % transfer function G
35 %% EXTRACT TF POLES, FREQUENCY, DAMPING OF G
36 [~,poles_TF,~]=zpkdata(G); s_TF=poles_TF{1,1}; % poles
37 % display(s_TF,'poles of G')
38 f_TF=abs(imag(s_TF)/(2*pi)); % frequencies
39 zz=real(s_TF)./abs(s_TF); z_TF=zz.*((abs(zz)>tol)); % damping
40 %% CALCULATE IMPULSE RESPONSE OF TF SYSTEM
41 figure(1);
42 subplot(2,2,1); impulse(G,t);
43 title(['TF system: z = ' num2str(z_TF(1)*100) '%'])
44 %% CALCULATE SISO STATE SPACE MODEL
45 %----- state space matrices -----
46 AA=[ 0 1 ;
47 -wn^2 -2*z*wn];
48 BB=[ 0 ;
49 wn^2];
50 CC=[1 0];
```

G:\EMCH516\EMCH51...\\FB\_single\_dof\_time\_response\_TF\_SS.m      Page 2

```
51 DD=[0];
52 ss_sys=ss(AA,BB,CC,DD);
53 %% EXTRACT SS POLES, FREQUENCY, DAMPING
54 [~,zz,poles_SS]=damp(ss_sys);
55 % display(ss,'poles of ss_sys')
56 s_SS=poles_SS; % poles
57 f_SS=abs(imag(poles_SS))/(2*pi); % frequencies
58 z_SS=zz.*((abs(zz)>tol)); % damping
59 %% CALCULATE IMPULSE RESPONSE OF SS SYSTEM
60 subplot(2,2,2); impulse(ss_sys,t);
61 title(['SS system: z = ' num2str(z_SS(1)*100) '%'])
62 %% ADD VELOCITY FEEDBACK TO IMPROVE DAMPING
63 K_ratio=1+z_ratio;
64 K=K_ratio*K_cr; % FB gain
65 display([z_ratio K_ratio K])
66 H=K*tf([1 0],1);
67 %% TF SYSTEM WITH FB
68 G_CL=feedback(G,H);
69 [~,p_CL_TF,~]=zpksdata(G_CL); s_CL_TF=p_CL_TF{1,1};
70 f_CL_TF=imag(s_CL_TF)/(2*pi); % frequencies
71 zz=-real(s_CL_TF)./abs(s_CL_TF); z_CL_TF=zz.*((abs(zz)>tol)); % damping
72 %% CALCULATE TF IMPULSE RESPONSE WITH FB
73 subplot(2,2,3); impulse(G_CL,t)
74 T1=['TF FB Impulse response'];
75 T2=[' z = ' num2str(z*100) '%'];
76 T3=['K ratio=' num2str(K_ratio)];
77 T4=[' z_C_L= ' num2str(z_CL_TF(1)*100) '%'];
78 line1=[T1 T2];
79 line2=[T3 T4];
80 title({line1; line2})
81 %% SS SYSTEM WITH FB
82 % ---- SS velocity feedback MATRIX-----
83 H=[0 K]; % FB matrix
84 % ===== state space matrices with FB =====
85 AA_CL=AA-BB*H; % state matrix AA with FB
86 % AA_CL=feedback(ss_sys,H); % this does not work; H should be also ss
87 CC_CL=CC-DD*H; % state matrix CC with FB
88 ss_sys_CL=ss(AA_CL, BB, CC_CL, DD);
89 %% EXTRACT POLES, FREQUENCY, DAMPING
90 [~,zz_CL_SS,poles_CL_SS]=damp(ss_sys_CL);
91 %ss_CL=eig(AA0);
92 % display(ss_CL,'poles of G_CL')
93 s_CL_SS=poles_CL_SS; % poles
94 f_CL_SS=abs(imag(poles_CL_SS))/(2*pi); % frequencies
95 z_CL_SS=zz_CL_SS.*((abs(zz_CL_SS)>tol)); % damping
96 %% CALCULATE IMPULSE RESPONSE OF SS SYSTEM WITH FB
97 subplot(2,2,4); impulse(ss_sys_CL,t);
98 title(['impulse response of SS FB system: zCL_SS= ' num2str(z_CL_SS(1)*100) '%'])
99 T1_SS=['SS FB Impulse response'];
100 T2_SS=[' z = ' num2str(z*100) '%'];
```

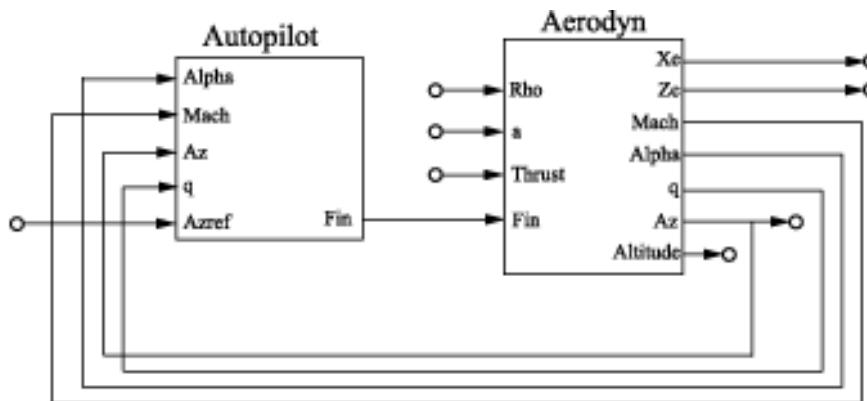
G:\EMCH516\EMCH51... \FB\_single\_dof\_time\_response\_TF\_SS.m      Page 3

```
101 T3_SS=['K_ratio=' num2str(K_ratio)];
102 T4_SS=[' z_C_L= ' num2str(z_CL_TF(1)*100) '%'];
103 line1_SS=[T1_SS T2_SS];
104 line2_SS=[T3_SS T4_SS];
105 title({line1_SS; line2_SS})
106 %% DISPLAY TF RESULTS
107 display(s_TF, 'TF poles');
108 display([f_TF z_TF*100], 'f_TF, Hz; z_TF%');
109 %% DISPLAY TF FB RESULTS
110 display(s_CL_TF, 'TF FB poles');
111 display([f_CL_TF z_CL_TF*100], 'f_CL, Hz; zh_CL %');
112 %% DISPLAY SS RESULTS
113 % display(' ');
114 display(s_SS, 'SS poles');
115 display([f_SS z_SS*100], 'f_SS, Hz; z_SS%');
116 %% DISPLAY SS FB RESULTS
117 % display(' ');
118 display(s_CL_SS, 'SS FB poles');
119 display([f_CL_SS z_CL_SS*100], 'f_CL_SS, Hz; z_CL_SS%');
120
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```

## MIMO Feedback Loop

This example shows how to obtain the closed-loop response of a MIMO feedback loop in three different ways.

In this example, you obtain the response from Azref to Az of the MIMO feedback loop of the following block diagram.



You can compute the closed-loop response using one of the following three approaches:

- Name-based interconnection with `connect`
- Name-based interconnection with `feedback`
- Index-based interconnection with `feedback`

You can use whichever of these approaches is most convenient for your application.

Load the plant Aerodyn and the controller Autopilot into the MATLAB® workspace. These models are stored in the datafile `MIMOfeedback.mat`.

```
load(fullfile(matlabroot, 'examples', 'control', 'MIMOfeedback.mat'))
```

Aerodyn is a 4-input, 7-output state-space (`ss`) model. Autopilot is a 5-input, 1-output `ss` model. The inputs and outputs of both models names appear as shown in the block diagram.

Compute the closed-loop response from Azref to Az using `connect`.

```
T1 = connect(Autopilot,Aerodyn,'Azref','Az');
```

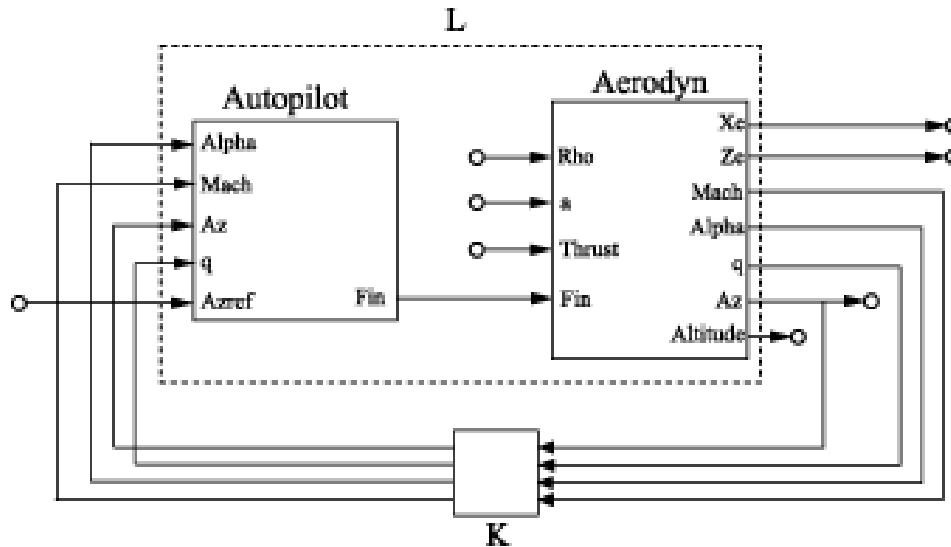
```
Warning: The following block inputs are not used: Rho,a,Thrust.
Warning: The following block outputs are not used: Xe,Ze,Altitude.
```

The `connect` function combines the models by joining the inputs and outputs that have matching names. The last two arguments to `connect` specify the input and output signals of the resulting model. Therefore, `T1` is

a state-space model with input Azref and output Az. The connect function ignores the other inputs and outputs in Autopilot and Aerodyn.

Compute the closed-loop response from Azref to Az using name-based interconnection with the feedback command. Use the model input and output names to specify the interconnections between Aerodyn and Autopilot.

When you use the feedback function, think of the closed-loop system as a feedback interconnection between an open-loop plant-controller combination L and a diagonal unity-gain feedback element K. The following block diagram shows this interconnection.



```
L = series(Autopilot,Aerodyn,'Fin');

FeedbackChannels = {'Alpha','Mach','Az','q'};
K = ss(eye(4), 'InputName', FeedbackChannels, ...
       'OutputName', FeedbackChannels);

T2 = feedback(L,K, 'name', +1);
```

The closed-loop model T2 represents the positive feedback interconnection of L and K. The 'name' option causes feedback to connect L and K by matching their input and output names.

T2 is a 5-input, 7-output state-space model. The closed-loop response from Azref to Az is T2('Az', 'Azref').

Compute the closed-loop response from Azref to Az using feedback, using indices to specify the interconnections between Aerodyn and Autopilot.

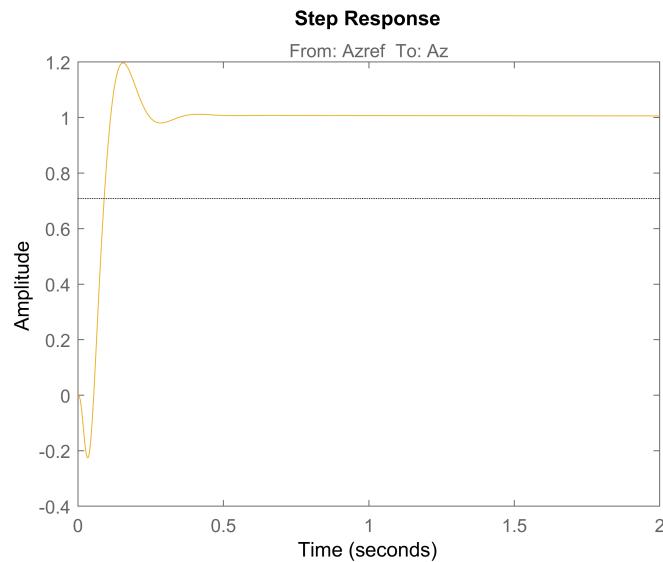
```
L = series(Autopilot,Aerodyn,1,4);
K = ss(eye(4));
T3 = feedback(L,K,[1 2 3 4],[4 3 6 5],+1);
```

The vectors [1 2 3 4] and [4 3 6 5] specify which inputs and outputs, respectively, complete the feedback interconnection. For example, feedback uses output 4 and input 1 of L to create the first feedback interconnection. The function uses output 3 and input 2 to create the second interconnection, and so on.

T3 is a 5-input, 7-output state-space model. The closed-loop response from Azref to Az is T3(6,5).

Compare the step response from Azref to Az to confirm that the three approaches yield the same results.

```
step(T1,T2('Az','Azref'),T3(6,5),2)
```



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MIMO Feedback Loop - MATLAB & Simulink

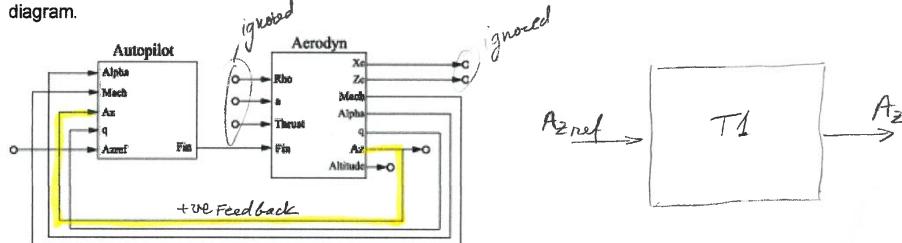
<http://www.mathworks.com/help/control/ug/build-a-model-of-a-multi-i...>

## MIMO Feedback Loop

This example shows how to obtain the closed-loop response of a MIMO feedback loop in three different ways.

[Open This Example](#)

In this example, you obtain the response from  $Az_{ref}$  to  $Az$  of the MIMO feedback loop of the following block diagram.



Compute the closed-loop response from  $Az_{ref}$  to  $Az$  using `connect`.

T1

*Input      Output*  
`T1 = connect(Autopilot,Aerodyn, 'Azref', 'Az');`

Warning: The following block inputs are not used: Rho,a,Thrust.

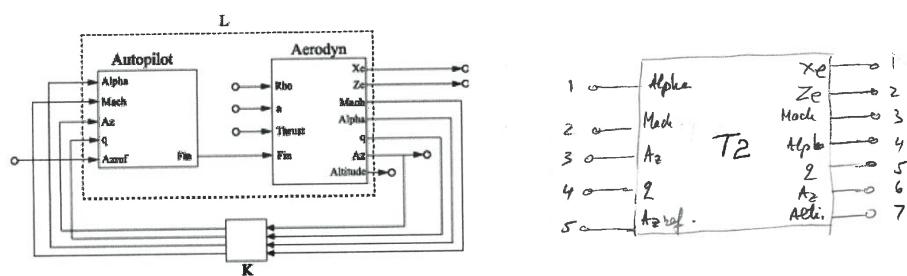
Warning: The following block outputs are not used: Xe,Ze,Altitude.

The `connect` function combines the models by joining the inputs and outputs that have matching names. The last two arguments to `connect` specify the input and output signals of the resulting model. Therefore, T1 is a

T2

Compute the closed-loop response from  $Az_{ref}$  to  $Az$  using name-based interconnection with the feedback command. Use the model input and output names to specify the interconnections between Aerodyn and Autopilot.

When you use the feedback function, think of the closed-loop system as a feedback interconnection between an open-loop plant-controller combination  $L$  and a diagonal unity-gain feedback element  $K$ . The following block diagram shows this interconnection.



$L = \text{series}(\text{Autopilot}, \text{Aerodyn}, \text{'Fin'})$ ; Series construct using variable Fin

```
FeedbackChannels = {'Alpha', 'Mach', 'Az', 'q'};
K = ss(eye(4), 'InputName', FeedbackChannels, ... unit ss system from Alpha, Mach, Az, q
      'OutputName', FeedbackChannels);
T2 = feedback(L, K, 'name', Az);  
+ve feedback
```

match names  
T2 = feedback(L, K, 'name', Az); +ve feedback

521/523

T3

Compute the closed-loop response from Azref to Az using feedback, using indices to specify the interconnections between Aerodyn and Autopilot.

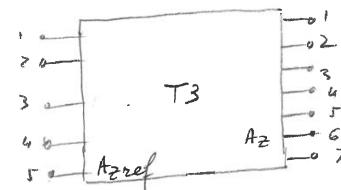
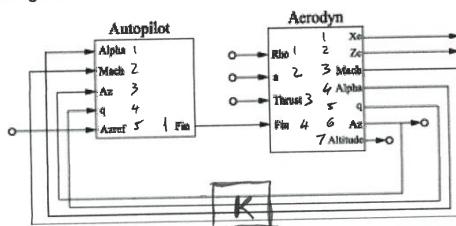
```

L = series(Autopilot,Aerodyn,1,4);
K = ss(eye(4));
T3 = feedback(L,K,[1 2 3 4],[4 3 6 5],-1);
    
```

+ve feedback  
feedback ( $G, H, \text{input}, \text{output}, \pm 1$ )

The vectors [1 2 3 4] and [4 3 6 5] specify which inputs and outputs, respectively, complete the feedback interconnection. For example, feedback uses output 4 and input 1 of L to create the first feedback interconnection. The function uses output 3 and input 2 to create the second interconnection, and so on.

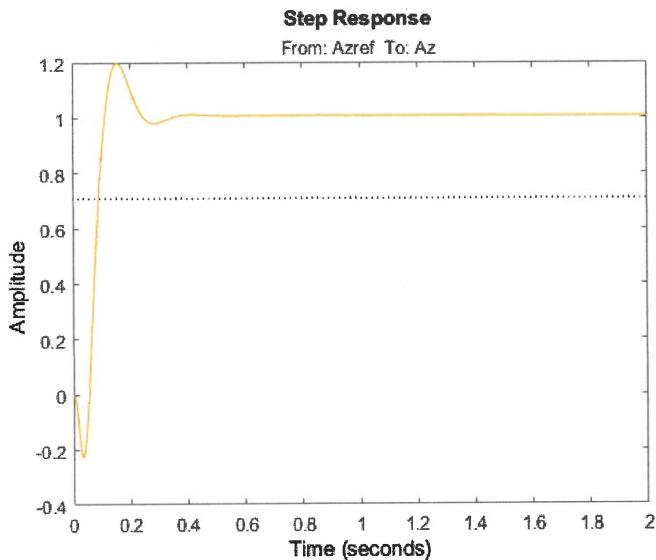
T3 is a 5-input, 7-output state-space model. The closed-loop response from Azref to Az is T3(6,5).



522/523

Compare the step response from Azref to Az to confirm that the three approaches yield the same results.

```
step(T1,T2('Az','Azref'),T3(6,5),2)
```



523/523

## 12 Machine Learning Control

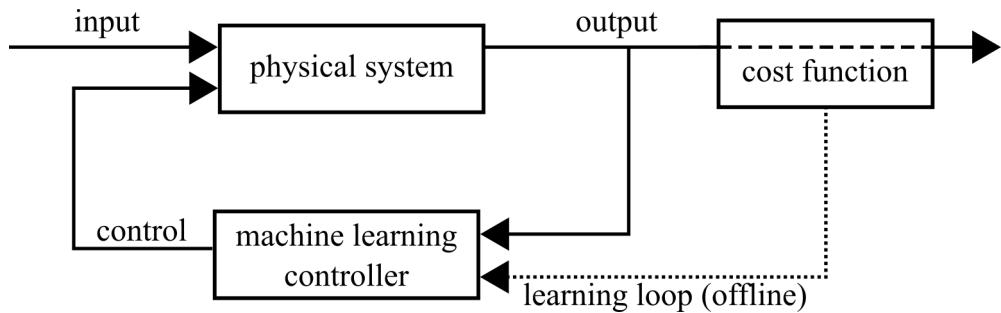


Figure 12.1: Simplified overview of a control system leveraging a machine learning controller.