Lecture Notes

Fundamentals of Control Systems

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Chapter 6

DESIGN OF CONTINUOUS CONTROL SYSTEMS
Content

- Introduction
- Effect of controllers on system performance
- Control systems design using the root locus method
- Control systems design in the frequency domain
- Design of PID controllers
- Control systems design in state-space
- Design of state estimators
Introduction
Introduction to design process

- Design is a process of adding/configuring hardware as well as software in a system so that the new system satisfies the desired specifications.
Series compensator

- The controller is connected in series with the plant.

![Block diagram of series compensator]

- Controllers: phase lead, phase lag, lead-lag compensator, P, PD, PI, PID,…

- Design method: root locus, frequency response

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State feedback control

- All the states of the system are fed back to calculate the control rule.

\[ \dot{x}(t) = Ax(t) + Bu(t) \]

- State feedback controller: \( u(t) = r(t) - Kx(t) \)
  \[ K = \begin{bmatrix} k_1 & k_2 & \ldots & k_n \end{bmatrix} \]

- Design method: pole placement, LQR, …
Effects of controller on system performance
The addition of a pole (in the left-half s-plane) to the open-loop transfer function has the effect of pushing the root locus to the right, tending to lower the system’s relative stability and to slow down the settling of the response.
The addition of a zero (in the left-half s-plane) to the open-loop transfer function has the effect of pulling the root locus to the left, tending to make the system more stable and to speed up the settling of the response.
Effects of lead compensators

- **Transfer function:**
  \[ G_C(s) = K_C \frac{1 + \alpha Ts}{1 + Ts} \quad (\alpha > 1) \]

- **Frequency response:**
  \[ G_C(j\omega) = K_C \frac{1 + \alpha Tj\omega}{1 + Tj\omega} \]

- **Characteristics of the Bode plots:**
  \[ \varphi_{\text{max}} = \sin^{-1}\left(\frac{\alpha - 1}{\alpha + 1}\right) \]
  \[ \omega_{\text{max}} = \frac{1}{T\sqrt{\alpha}} \]
  \[ L(\omega_{\text{max}}) = 20\log K_C + 10\log \alpha \]

- **The lead compensators improve the transient response (POT, t_s,..)**
Lead compensator implementation

Lead compensator transfer function:

\[
\frac{U(s)}{E(s)} = \frac{R_2 R_4}{R_1 R_3} \frac{1 + R_1 C_1 s}{1 + R_2 C_2 s} = K_c \frac{1 + \alpha T_s}{1 + T_s}
\]

\((\alpha > 1 \Leftrightarrow R_1 C_1 > R_2 C_2)\)

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Effects of lag compensators

- **Transfer function:**
  \[ G_C(s) = K_C \frac{1 + \alpha T_s}{1 + T_s} \quad (\alpha < 1) \]

- **Frequency response:**
  \[ G_C(j\omega) = K_C \frac{1 + \alpha T_j \omega}{1 + T_j \omega} \]

- **Characteristics of the Bode plots:**
  \[ \varphi_{\min} = \sin^{-1} \left( \frac{\alpha - 1}{\alpha + 1} \right) \]
  \[ \omega_{\min} = \frac{1}{T \sqrt{\alpha}} \]
  \[ L(\omega_{\min}) = 20 \log_2 K_C + 10 \log_2 \alpha \]

- **The lag compensators reduce the steady-state error.**
Lag compensator implementation

- Lag compensator transfer function:

\[
\frac{U(s)}{E(s)} = \frac{R_2 R_4}{R_1 R_3} \frac{1 + R_1 C_1 s}{1 + R_2 C_2 s} = K_c \frac{1 + \alpha T s}{1 + T s} \quad (\alpha < 1 \iff R_1 C_1 < R_2 C_2)
\]
Effects of lead-lag compensators

Transfer function: 
\[ G_C(s) = K_C \left( \frac{1 + \alpha_1 T_1 s}{1 + T_1 s} \right) \left( \frac{1 + \alpha_2 T_2 s}{1 + T_2 s} \right) \]  
\((\alpha_1 < 1, \alpha_2 > 1)\)

Bode diagram

The lead-lag compensators improve transient response and reduces the steady-state error.
Effects of proportional controller (P)

- Transfer function: \( G_c(s) = K_p \)

- Increasing proportional gain leads to decreasing steady-state error, however, the system become less stable, and the POT increases.

- Ex: response of a proportional control system whose plant has the transfer function below:

\[
G(s) = \frac{10}{(s + 2)(s + 3)}
\]
Transfer function:

\[ G_C(s) = K_P + K_D s = K_P (1 + T_D s) \]

The PD controller is a special case of phase lead compensator, the maximum phase lead is \( \phi_{\text{max}} = 90^\circ \) at the frequency \( \omega_{\text{max}} = +\infty \).

The PD controller speed up the response of the system, however it also makes the system more sensitive to high frequency noise.
Effects of proportional derivative controller (PD)

* Note: The larger the derivative constant, the faster the response of the system.
PD controller implementation

- PD controller transfer function:

\[
\frac{U(s)}{E(s)} = \frac{R_2 R_4}{R_1 R_3} (1 + R_1 C_1 s) = K_P + K_D s
\]
Effects of proportional integral controller (PI)

- **Transfer function:**
  \[ G_C(s) = K_p + \frac{K_i}{s} = K_p \left(1 + \frac{1}{T_i s}\right) \]

- **Bode diagram**

- The PI controller is a special case of phase lag compensator, the minimum phase lag is \( \phi_{\text{min}} = -90^0 \) at the frequency \( \omega_{\text{min}} = +\infty \).

- PI controllers eliminate steady state error to step input, however it can increase POT and settling time.
Effects of proportional integral controller (PI)

Note: The larger the integral constant, the larger the POT of response of the system.
**PI controller implementation**

- **PI controller transfer function:**

\[
\frac{U(s)}{E(s)} = \frac{R_2 R_4}{R_1 R_3} \frac{R_2 C_2 s + 1}{R_2 C_2 s} = K_P + \frac{K_I}{s}
\]
Effects of proportional integral controller (PID)

★ Transfer function:

\[ G_C(s) = K_P + \frac{K_I}{s} + K_Ds \]

\[ \equiv G_C(s) = K_P\left(1 + \frac{1}{T_I s}\right)(1 + T_D s) \]

\[ \equiv G_C(s) = K_P\left(1 + \frac{1}{T_I s}\right)(1 + T_D s) \]

★ Bode diagram

★ Effects of PID controllers:

▸ speed up response of the system

▸ Eliminate steady-state error to step input.
Comparison of PI, PD and PID controllers

![Graph comparing PI, PD, and PID controllers](image)

- **PI**
- **PD**
- **PID**
- **Uncompensated**

The graph illustrates the response of different control systems over time (t). The y-axis represents the system output y(t), and the x-axis represents time t.
Control systems design using the root locus method
Procedure for designing lead compensator using the root locus

**Lead compensator:** \[ G_C(s) = K_C \frac{s + (1/\alpha T)}{s + (1/T)} \quad (\alpha > 1) \]

- **Step 1:** Determine the dominant poles \( s_{1,2}^* \) from desired transient response specification:
  - Overshoot (POT)
  - Settling time \( t_s \)

  \[
  \begin{cases}
  \text{Overshoot (POT)} & \Rightarrow \begin{cases} 
  \xi \\
  \omega_n 
  \end{cases} \\
  \text{Settling time } t_s & \Rightarrow s_{1,2}^* = -\xi \omega_n \pm j\omega_n \sqrt{1 - \xi^2}
  \end{cases}
  \]

- **Step 2:** Determine the deficiency angle so that the dominant poles \( s_{1,2}^* \) lie on the root locus of the compensated system:

  \[
  \phi^* = -180^0 + \sum_{i=1}^n \arg(s_i^* - p_i) - \sum_{i=1}^m \arg(s_i^* - z_i)
  \]

  where \( p_i \) and \( z_i \) are poles & zeros of \( G(s) \) before compensation.

  \[
  \phi^* = -180^0 + \sum \text{angle from } p_i \text{ to } s_1^* - \sum \text{angle from } z_i \text{ to } s_1^*
  \]
Procedure for designing lead compensator using the root locus

**Step 3: Determine the pole & zero** of the lead compensator

Draw 2 arbitrarily rays starting from the dominant pole $s_1^*$ such that the angle between the two rays equal to $\phi^*$. The intersection between the two rays and the real axis are the positions of the pole and the zero of the lead compensator.

Two methods often used for drawing the rays:

- Bisector method
- Pole elimination method

**Step 4: Calculate the gain $K_C$** using the formula:

$$\left| G_C(s)G(s) \right|_{s=s_1^*} = 1$$
**Objective:** design the compensator \( G_C(s) \) so that the response of the compensated system satisfies: \( \text{POT} < 20\%; \) \( t_s < 0.5 \text{sec} \) (2% criterion).

**Solution:**

Because the design objective is to improve the transient response, we need to design a lead compensator:

\[
G_C(s) = K_c \frac{s + \left( \frac{1}{\alpha T} \right)}{s + \left( \frac{1}{T} \right)} \quad (\alpha > 1)
\]
**Step 1:** Determine the dominant poles:

\[ \text{POT} = \exp \left( -\frac{\xi \pi}{\sqrt{1 - \xi^2}} \right) < 0.2 \quad \Rightarrow \quad -\frac{\xi \pi}{\sqrt{1 - \xi^2}} < \ln 0.2 = -1.6 \quad \Rightarrow \quad \xi > 0.45 \]

Chose \( \xi = 0.707 \)

\[ t_{q^d} = \frac{4}{\xi \omega_n} < 0.5 \quad \Rightarrow \quad \omega_n > \frac{4}{0.5 \times \xi} \quad \Rightarrow \quad \omega_n > 11.4 \]

Chose \( \omega_n = 15 \)

The dominant poles are:

\[ s_{1,2}^* = -\xi \omega_n \pm j\omega_n \sqrt{1 - \xi^2} = -0.707 \times 15 \pm j15\sqrt{1 - 0.707^2} \]

\[ s_{1,2}^* = -10.5 \pm j10.5 \]
**Step 2**: Determine the deficiency angle:

**Method 1:**

\[ \phi^* = -180^0 + \{ \arg((-10,5 + j10,5) - 0) + \arg((-10,5 + j10,5) - (-5)) \} \]

\[ = -180^0 + \left\{ \arctan\left(\frac{10,5}{-10,5}\right) + \arctan\left(\frac{10,5}{-5,5}\right) \right\} \]

\[ = -180^0 + (135 + 117,6) \]

\[ \Rightarrow \phi^* = 72,6^0 \]

**Method 2:**

\[ \phi^* = -180^0 + (\beta_1 + \beta_2) \]

\[ = -180^0 + (135^0 + 117,6^0) \]

\[ \Rightarrow \phi^* = 72,6^0 \]
**Step 3:** Determine the pole and the zero of the compensator (bisector method)

\[ OB = OP \sin \left( \frac{O\hat{P}x + \phi^*}{2} \right) = 28.12 \]

\[ OC = OP \sin \left( \frac{O\hat{P}x - \phi^*}{2} \right) = 8.0 \]

\[ \Rightarrow G_C(s) = K_C \frac{s + 8}{s + 28} \]
Example of designing a lead compensator using RL (cont’)

**Step 4:** Determine the gain of the compensator:

\[
|G_C(s)G(s)|_{s=s^*} = 1
\]

\[
\Leftrightarrow K_C \frac{-10,5 + j10,5 + 8}{-10,5 + j10,5 + 28 \cdot (-10,5 + j10,5)(-10,5 + j10,5 + 5)} = 1
\]

\[
\Leftrightarrow K_C \frac{10,79 \times 50}{20,41 \times 15 \times 11,85} = 1
\]

\[
\Leftrightarrow K_C = 6,7
\]

**Conclusion:** The transfer function of the lead compensator is:

\[
G_C(s) = 6,7 \frac{s + 8}{s + 28}
\]
Root locus of the system

Root locus of the uncompensated system

Root locus of the compensated system
Transient response of the system

\[ y(t) \]

![Graph of transient response](image)

Transient response of the system

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Procedure for designing lag compensator using the root locus

**Lag compensator:**

\[ G_C(s) = K_C \frac{s + (1/\beta T)}{s + (1/T)} \quad (\beta < 1) \]

- **Step 1:** Determine \( \beta \) to meet the steady-state error requirement:

  \[ \beta = \frac{K_P}{K_P^*} \quad \text{or} \quad \beta = \frac{K_V}{K_V^*} \quad \text{or} \quad \beta = \frac{K_a}{K_a^*} \]

- **Step 2:** Chose the zero of the lag compensator:

  \[ \frac{1}{\beta T} << |\text{Re}(s_{1,2}^*)| \]

- **Step 3:** Calculate the pole of the compensator:

  \[ \frac{1}{T} = \beta \cdot \frac{1}{\beta T} \]

- **Step 4:** Calculate \( K_C \) satisfying the condition:

  \[ |G_C(s)G(s)|_{s=s_{1,2}^*} = 1 \]
Objective: design the compensator $G_C(s)$ so that the compensated system satisfies the following performances: steady state error to ramp input is 0.02 and transient response of the compensated system is nearly unchanged.

Solution:

The compensator to be design is a lag compensator:

$$G_C(s) = K_C \frac{s + (1/\beta T)}{s + (1/T)} \quad (\beta < 1)$$
**Step 1:** Determine $\beta$

The velocity constant of uncompensated system:

$$K_V = \lim_{s \to 0} sG(s) = \lim_{s \to 0} \frac{10}{s(s + 3)(s + 4)} = 0.83$$

The desired velocity constant:

$$K_V^* = \frac{1}{e_{x_l}^*} = \frac{1}{0.02} = 50$$

Then:

$$\beta = \frac{K_V}{K_V^*} = \frac{0.83}{50} = 0.017$$
Example of designing a lag compensator using RL (cont’)

★ **Step 2:** Chose the zero of the lag compensator

The pole of the uncompensated system:

\[ 1 + G(s) = 0 \iff 1 + \frac{10}{s(s + 3)(s + 4)} = 0 \iff \begin{cases} s_{1,2} = -1 \pm j \\ s_3 = -5 \end{cases} \]

⇒ The dominant poles of the uncompensated system: \( s_{1,2} = -1 \pm j \)

Chose: \( \frac{1}{\beta T} \ll \left| \text{Re}\{s_1\} \right| = 1 \implies \frac{1}{\beta T} = 0,1 \)

★ **Step 3:** Calculate the pole of the compensator:

\[
\frac{1}{T} = \beta \frac{1}{\beta T} = (0,017)(0,1) \implies \frac{1}{T} = 0,0017
\]

⇒ \( G_C(s) = K_C \frac{s + 0,1}{s + 0,0017} \)
**Step 4:** Determine the gain of the compensator

\[
|G_C(s)G(s)|_{s=s^*} = 1
\]

\[
\Leftrightarrow \quad K_C \left. \frac{s + 0.1}{s + 0.0017 \cdot s(s + 3)(s + 4)} \right|_{s = -1 \pm j} = 1
\]

\[
\Rightarrow \quad K_C \left. \frac{-1 + j + 0.1}{(-1 + j + 0.0017) \cdot (-1 + j)(-1 + j + 3)(-1 + j + 4)} \right| = 1
\]

\[
K_C = 1.0042 \approx 1
\]

\[
\Rightarrow \quad G_C(s) = \frac{s + 0.1}{s + 0.0017}
\]
Root locus of the system

Root locus of the uncompensated system

Root locus of the compensated system
Transient response of the system

\[ y(t) \]

uncompensated
compensated

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The compensator to be designed

\[ G_{C}(s) = G_{C1}(s)G_{C2}(s) \]

- **Step 1:** Design the lead compensator \( G_{C1}(s) \) to satisfy the transient response performances.

- **Step 2:** Let \( G_{1}(s) = G(s) \cdot G_{C1}(s) \)

  Design the lag compensator \( G_{C2}(s) \) in series with \( G_{1}(s) \) to satisfy the steady-state performances (and not to degrade the transient response obtained after phase lead compensating)
Example of designing a lead lag compensator using RL

★ **Objective:** design the compensator \( G_C(s) \) so that the compensated system has the dominant poles with \( \zeta = 0.5 \), \( \omega_n = 5 \) (rad/sec) and the velocity constant \( K_v = 80 \).

★ **Solution**

★ The compensator to be designed is a lead lag compensator because the design objective is to improve the transient response and to reduce the steady-state error.

\[
G_C(s) = G_{C1}(s)G_{C2}(s)
\]
**Step 1**: Design the lead compensator $G_{C1}(s)$

The dominant poles:

$$s_{1,2} = -\xi \omega_n \pm j \omega_n \sqrt{1 - \xi^2} = -0.5 \times 5 \pm j5\sqrt{1 - 0.5^2}$$

$$s_{1,2}^* = -2,5 \pm j4,33$$

The deficiency angle:

$$\phi^* = -180^0 + (\beta_1 + \beta_2)$$

$$= -180^0 + (120^0 + 115^0)$$

$$\phi^* = 55^0$$
Example of designing a lead lag compensator using RL (cont’)

Chose the zero of the lead compensator so that it eliminates the pole at $-0.5$ of $G(s)$ (pole elimination method)

\[
\frac{1}{\alpha T_1} = 0.5
\]

\[OA = 0.5\]

\[AB = PA \frac{\sin A\hat{P}B}{\sin PAB} = 4.76 \frac{\sin 55^0}{\sin 60^0} = 4.5\]

\[
\frac{1}{T_1} = OA + AB = 5
\]

\[G_{C1}(s) = K_{C1} \frac{s + 0.5}{s + 5}\]
Example of designing a lead lag compensator using RL (cont’)

Calculate $K_{C1}$: 
$$ \left| G_{C1}(s)G(s) \right|_{s=s^*} = 1 $$

$$ K_{C1} \left. \begin{array}{c}
\frac{s + 0.5}{s + 5} \\
\frac{4}{s(s + 0.5)}
\end{array} \right|_{s=2.5 + j4.33} = 1 $$

$K_{C1} = 6.25$

$$ \Rightarrow G_{C1}(s) = 6.25 \frac{s + 0.5}{s + 5} $$

The lead-compensated open-loop system:

$$ G_1(s) = G_{C1}(s)G(s) = \frac{25}{s(s + 5)} $$
Example of designing a lead lag compensator using RL (cont’)

**Step 2**: Design the lag compensator $G_{C2}(s)$

$$G_{C2}(s) = K_{C2} \frac{1}{s + \frac{\beta T_2}{T_2}}$$

- Determine $\beta$:

$$K_V = \lim_{s \to 0} sG_1(s) = \lim_{s \to 0} s \frac{25}{s(s + 5)} = 5$$

$$K_V^* = 80$$

$$\Rightarrow \quad \beta = \frac{K_V}{K_V^*} = \frac{5}{80} = \frac{1}{16}$$
Example of designing a lead lag compensator using RL (cont’)

– Determine the zero of the lag compensator:

\[
\frac{1}{\beta T_2} \ll \left| \text{Re}(s^*) \right| = \left| \text{Re}(-2.5 + j4.33) \right| = 2.5
\]

Chose: \( \frac{1}{\beta T_2} = 0.16 \)

– Calculate the pole of the lag compensator:

\[
\frac{1}{T_2} = \beta \cdot \frac{1}{\beta T_2} = \frac{1}{16}(0.16)
\]

\[\Rightarrow \quad \frac{1}{T_2} = 0.01\]
Example of designing a lead lag compensator using RL (cont’)

- Calculate \( K_{C2} \) using the gain condition: \( \left| G_{C2}(s)G_1(s) \right|_{s=s^*} = 1 \)

\[
\Rightarrow \left( \left| G_{C2}(s) \right|_{s=s^*} \left| G_1(s) \right|_{s=s^*} \right) = 1
\]

\[
\Rightarrow \left| K_{C2} \frac{-2.5 + j4.33 + 0.16}{-2.5 + j4.33 + 0.01} \right| = 1
\]

\[
\Rightarrow K_{C2} = 1.01
\]

The transfer function of the lag compensator:

\[
G_{C2}(s) = 1.01 \frac{(s + 0.16)}{(s + 0.01)}
\]

Final result:

\[
G_C(s) = G_{C1}(s)G_{C2}(s) = 6.31 \frac{(s + 0.5)(s + 0.16)}{(s + 5)(s + 0.01)}
\]
Control system design in frequency domain
Procedure for designing lead compensators in frequency domain

The lead compensator: 

\[ G_C(s) = K_C \frac{\alpha Ts + 1}{Ts + 1} \quad (\alpha > 1) \]

**Step 1:** Determine \( K_C \) to meet the steady-state error requirement: 

\[ K_C = \frac{K_P^*}{K_P} \quad \text{or} \quad K_C = \frac{K_V^*}{K_V} \quad \text{or} \quad K_C = \frac{K_a^*}{K_a} \]

**Step 2:** Let \( G_1(s) = K_C G(s) \). Plot the Bode diagram of \( G_1(s) \)

**Step 3:** Determine the gain crossover frequency of \( G_1(s) \):

\[ L_1(\omega_C) = 0 \quad \text{or} \quad |G_1(j\omega_C)| = 1 \]

**Step 4:** Determine the phase margin of \( G_1(s) \) (phase margin of uncompensated system):

\[ \Phi M = 180 + \varphi_1(\omega_C) \]

**Step 5:** Determine the necessary phase lead angle to be added to the system:

\[ \varphi_{\text{max}} = \Phi M^* - \Phi M + \theta \]

\( \Phi M^* \) is the desired phase margin, \( \theta = 5^0 \div 20^0 \)
Procedure for designing lead compensators in frequency domain

- **Step 6:** Calculate \( \alpha \):
  \[ \alpha = \frac{1 + \sin \varphi_{\text{max}}}{1 - \sin \varphi_{\text{max}}} \]

- **Step 7:** Determine the **new gain crossover frequency** (of the compensated open-loop system) using the conditions:
  \[ L_1(\omega_C') = -10 \log \alpha \quad \text{or} \quad |G_1(j\omega_C')| = 1/\sqrt{\alpha} \]

- **Step 8:** Calculate the **time constant** \( T \):
  \[ T = \frac{1}{\omega_C' \sqrt{\alpha}} \]

- **Step 9:** Check if the compensated system satisfies the gain margin? If not, repeat the design procedure from step 5.

**Note:** It is possible to determine \( \omega_C \) (step 3), \( \Phi_M \) (step 4) and \( \omega_C' \) (step 7) by using Bode diagram instead of using analytic calculation.
Objective: Design the compensator $G_C(s)$ so that the compensated system satisfies the performances:

$$K_V^* = 20; \quad \Phi M^* \geq 50^0; \quad GM^* \geq 10\text{dB}$$

Solution:

The transfer function of the lead compensator to be designed:

$$G_C(s) = K_C \frac{1 + \alpha Ts}{1 + Ts}$$

$(\alpha > 1)$
**Step 1:** Determine $K_C$

The velocity constant of the uncompensated system:

$$K_V = \lim_{s \to 0} sG(s) = \lim_{s \to 0} s \frac{4}{s(s + 2)} = 2$$

The desired velocity constant: $K_V^* = 20$

$$\Rightarrow K_C = \frac{K_V^*}{K_V} = \frac{20}{2} \Rightarrow K_C = 10$$

**Step 2:** Denote $G_1(s) = K_C G(s) = 10 \cdot \frac{4}{s(s + 2)}$

$$\Rightarrow G_1(s) = \frac{20}{s(0.5s + 1)}$$

Draw the Bode diagram of $G_1(s)$.
Design lead compensator in frequency domain – Example (cont’)

-20dB/dec

-40dB/dec

\( \omega_c = 6 \)

\( \Phi M \)

+180

+90

0

-90

-160

10^0 10^1 10^2

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**Step 3:** The gain crossover frequency of $G_1(s)$

According to the Bode diagram: $\omega_c \approx 6$ (rad/sec)

**Step 4:** The phase margin of $G_1(s)$

According to the Bode diagram:

$$\varphi_1(\omega_c) \approx -160^0$$

$$\Rightarrow \Phi M = 180 + \varphi_1(\omega_c) \approx 20^0$$

**Step 5:** The necessary phase lead angle to be added:

$$\varphi_{max} = \Phi M^* - \Phi M + \theta$$ (chose $\theta = 7$)

$$\Rightarrow \varphi_{max} = 50^0 - 20^0 + 7^0$$

$$\Rightarrow \varphi_{max} = 37^0$$
**Step 6:** Calculate \( \alpha \)

\[
\alpha = \frac{1 + \sin \varphi_{\text{max}}}{1 - \sin \varphi_{\text{max}}} = \frac{1 + \sin 37^0}{1 - \sin 37^0} \quad \Rightarrow \quad \alpha = 4
\]

**Step 7:** Determine the new gain crossover frequency using Bode plot

\[
L_1(\omega'_C) = -10 \log \alpha = -10 \log 4 = -6 \text{dB}
\]

The abscissa of the intersection between Bode magnitude diagram and the horizontal line with ordinate of 6dB is the new gain crossover frequency. According to the plot (in slide 54), we have:

\( \omega'_C \approx 9 \) (rad/sec)

**Step 8:** Calculate \( T \)

\[
T = \frac{1}{\omega'_C \sqrt{\alpha}} = \frac{1}{(9)(\sqrt{4})} \quad \Rightarrow \quad T = 0.056 \quad \Rightarrow \quad \alpha T = 0.224
\]
Design lead compensator in frequency domain – Example (cont’)

\[ -20\text{dB/dec} \]

\[ 40\text{dB/dec} \]

\[ +20\text{dB/dec} \]

\[ -20\text{dB/dec} \]

\[ -40\text{dB/dec} \]

\[ \omega_c = 6 \]

\[ 1/\alpha T = 4.5 \]

\[ \omega_c' = 9 \]

\[ 1/T = 18 \]

\[ \Phi M \]

\[ \Phi M' \]
**Step 9:** Check the gain margin of the compensated system

According to the compensated Bode diagram, $GM^* = +\infty$, then the compensated system fulfills the design requirements.

**Conclusion:** The designed lead compensator is:

$$G_C(s) = 10 \frac{1 + 0.224s}{1 + 0.056s}$$
Objective: Design the compensator $G_c(s)$ so that the compensated system has: $\Phi M^* \geq 50^0$, $GM^* \geq 10\, dB$ and steady-state error to unit step input $e_{ss}^* \leq 0.05$;
Procedure for designing lag compensators in frequency domain

The lag compensator:

\[ G_C(s) = K_C \frac{\alpha T_s + 1}{T_s + 1} \quad (\alpha < 1) \]

- **Step 1**: Determine \( K_C \) to meet the steady-state error requirement:
  \[ K_C = K_P^*/K_P \quad \text{or} \quad K_C = K_V^*/K_V \quad \text{or} \quad K_C = K_a^*/K_a \]

- **Step 2**: Let \( G_1(s) = K_C G(s) \). Plot the Bode diagram of \( G_1(s) \)

- **Step 3**: Determine the new gain crossover frequency \( \omega'_C \) satisfying the following condition:
  \[ \varphi_1(\omega'_C) = -180^0 + \Phi M^* + \theta \]
  \( \Phi M^* \) is the desired phase margin, \( \theta = 5^0 \div 20^0 \)

- **Step 4**: Calculate \( \alpha \) using the condition:
  \[ L_1(\omega'_C) = -20 \log \alpha \quad \text{or} \quad \left| G_1(j \omega'_C) \right| = \frac{1}{\alpha} \]
Procedure for designing lag compensators in frequency domain

**Step 5**: Chose the zero of the lag compensator so that:

\[
\frac{1}{\alpha T} \ll \omega'_C \quad \Rightarrow \quad \alpha T
\]

**Step 6**: Calculate the **time constant** \( T \):

\[
\frac{1}{T} = \alpha \frac{1}{\alpha T} \quad \Rightarrow \quad T
\]

**Step 7**: Check if the compensated system satisfies the gain margin? If not, repeat the design procedure from step 3.

**Note**: It is possible to determine \( \varphi_1(\omega'_C) \), \( \omega'_C \) (step 3), \( L_1(\omega'_C) \) (step 4) by using Bode diagram instead of using analytic calculation.
Objective: design the lag compensator $G_C(s)$ so that the compensated system satisfies the following performances:

$$K_V^* = 5; \quad \Phi M^* \geq 40^0; \quad GM^* \geq 10 dB$$

Solution

The transfer function of the lag compensator to be designed:

$$G_C(s) = K_C \frac{1 + \alpha Ts}{1 + Ts} \quad (\alpha < 1)$$
Step 1: Determine $K_C$

The velocity constant of the uncompensated system:

$$K_V = \lim_{s \to 0} s G(s) = \lim_{s \to 0} s \frac{1}{s(s + 1)(0.5s + 1)} = 1$$

The desired velocity constant: $K_V^* = 5$

$$\Rightarrow \quad K_C = \frac{K_V^*}{K_V} = 5$$

Step 2: Denote $G_1(s) = K_C G(s)$

$$\Rightarrow \quad G_1(s) = \frac{5}{s(s + 1)(0.5s + 1)}$$

Draw the Bode diagram of $G_1(s)$
Design lag compensator in frequency domain – Example (cont’)

-20 dB/dec
-40 dB/dec
-60 dB/dec

-90°
-180°
-270°

10^{-3} 10^{-2} 10^{-1} 10^{0} 10^{1}
Step 3: Determine the new gain crossover frequency:

\[ \varphi_1(\omega_C') = -180^0 + \Phi M^* + \theta \]

\[ \Rightarrow \varphi_1(\omega_C') = -180^0 + 40^0 + 5^0 \]

\[ \Rightarrow \varphi_1(\omega_C') = -135^0 \]

According to the Bode diagram: \( \omega_C' \approx 0.5 \) (rad/sec)

Step 4: Calculate \( \alpha \) using the condition:

\[ L_1(\omega_C') = -20 \log \alpha \]

According the Bode diagram: \( L_1(\omega_C') \approx 18 \) (dB)

\[ \Rightarrow 18 = -20 \log \alpha \Rightarrow \log \alpha = -0.9 \Rightarrow \alpha = 10^{-0.9} \]

\[ \Rightarrow \alpha = 0.126 \]
**Step 5:** Chose the zero of the lag compensator:

\[
\frac{1}{\alpha T} \ll \omega'_c = 0.5
\]

Chose \( \frac{1}{\alpha T} = 0.05 \) \( \Rightarrow \) \( \alpha T = 20 \)

**Step 6:** Calculate the time constant \( T \)

\[
\frac{1}{T} = \alpha \frac{1}{\alpha T} = 0.126 \times 0.05 = 0.0063 \Rightarrow T = 159
\]

**Step 7:** It can be verified in the Bode diagram that the compensated system satisfies the gain margin requirement.

**Conclusion**

\[
G_C(s) = 5 \frac{(20s + 1)}{(159s + 1)}
\]
Design lag compensator in frequency domain – Example (cont’)

\[ L_1(\omega'_c) = 14 \]

\[ L'(\omega'_{-\pi}) = -135 \]

-20dB/dec
-40dB/dec
-60dB/dec

\[ \omega'_c = 0.5 \]
\[ \omega'_{-\pi} = 2 \]
Objective: Design the compensator $G_c(s)$ so that the compensated system has: $\Phi M^* \geq 50^0; GM^* \geq 10dB$ and steady-state error to unit step input $e_{ss}^* \leq 0.05$;

Solution:
### Comparison of phase lead and phase lag compensator

<table>
<thead>
<tr>
<th></th>
<th>Phase-Lead</th>
<th>Phase-Lag</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Approach</strong></td>
<td>Addition of phase-lead angle near crossover frequency on Bode diagram.</td>
<td>Addition of phase-lag to yield an increased error constant while maintaining desired dominant roots in s-plane or phase margin on Bode diagram</td>
</tr>
<tr>
<td></td>
<td>Add lead network to yield desired dominant roots in s-plane.</td>
<td></td>
</tr>
<tr>
<td><strong>Results</strong></td>
<td>1. Increases system bandwidth</td>
<td>1. Decreases system bandwidth</td>
</tr>
<tr>
<td></td>
<td>2. Increases gain at higher frequencies</td>
<td></td>
</tr>
<tr>
<td><strong>Advantages</strong></td>
<td>1. Yields desired response</td>
<td>1. Suppresses high-frequency noise</td>
</tr>
<tr>
<td></td>
<td>2. Improves dynamic response</td>
<td>2. Reduces steady-state error</td>
</tr>
<tr>
<td><strong>Disadvantages</strong></td>
<td>1. Requires additional amplifier gain</td>
<td>1. Slows down transient response</td>
</tr>
<tr>
<td></td>
<td>2. Increases bandwidth and thus susceptibility to noise</td>
<td>2. May require large values of components for $RC$ network</td>
</tr>
<tr>
<td></td>
<td>3. May require large values of components for $RC$ network</td>
<td></td>
</tr>
<tr>
<td><strong>Applications</strong></td>
<td>1. When fast transient response is desired</td>
<td>1. When error constants are specified</td>
</tr>
<tr>
<td><strong>Situations not applicable</strong></td>
<td>1. When phase decreases rapidly near crossover frequency</td>
<td>1. When no low-frequency range exists where phase is equal to desired phase margin</td>
</tr>
</tbody>
</table>

(Dorf and Bishop (2008), Modern control system –p.729)
Design of PID controllers
Determine the PID parameters based on the step response of the open-loop system.

\[ u(t) \rightarrow \text{Plant} \rightarrow y(t) \]

\[ y(t) \]

\[ K \]

\[ T_1 \quad T_2 \]
Zeigler – Nichols method 1 (cont’)

PID controller:

\[ G_C(s) = K_P \left( 1 + \frac{1}{T_I s} + T_D s \right) \]

<table>
<thead>
<tr>
<th>Controller</th>
<th>( K_P )</th>
<th>( T_I )</th>
<th>( T_D )</th>
</tr>
</thead>
<tbody>
<tr>
<td>P</td>
<td>( \frac{T_2}{(T_1K)} )</td>
<td>( \infty )</td>
<td>0</td>
</tr>
<tr>
<td>PI</td>
<td>0.9( \frac{T_2}{(T_1K)} )</td>
<td>0.3( T_1 )</td>
<td>0</td>
</tr>
<tr>
<td>PID</td>
<td>1.2( \frac{T_2}{(T_1K)} )</td>
<td>2( T_1 )</td>
<td>0.5( T_1 )</td>
</tr>
</tbody>
</table>
∗ **Problem:** Design a PID controller to control a furnace providing the open-loop characteristic of the furnace obtained from an experiment beside.

\[ K = 150 \]

\[ T_1 = 8 \text{ min} = 480 \text{ sec} \]

\[ T_2 = 24 \text{ min} = 1440 \text{ sec} \]

\[ K_P = 1.2 \frac{T_2}{T_1 K} = 1.2 \frac{1440}{480 \times 150} = 0.024 \]

\[ T_I = 2T_1 = 2 \times 480 = 960 \text{ sec} \]

\[ T_D = 0.5T_1 = 0.5 \times 480 = 240 \text{ sec} \]

\[ G_{PID}(s) = 0.024 \left(1 + \frac{1}{960s} + 240s\right) \]
Determine the PID parameters based on the response of the closed-loop system at the stability boundary.
Zeigler – Nichols method 2 (cont’)

**PID controller:**

\[ G_C(s) = K_P \left( 1 + \frac{1}{T_I s} + T_D s \right) \]

<table>
<thead>
<tr>
<th>Controller</th>
<th>( K_P )</th>
<th>( T_I )</th>
<th>( T_D )</th>
</tr>
</thead>
<tbody>
<tr>
<td>P</td>
<td>0.5( K_{cr} )</td>
<td>( \infty )</td>
<td>0</td>
</tr>
<tr>
<td>PI</td>
<td>0.45( K_{cr} )</td>
<td>0.83( T_{cr} )</td>
<td>0</td>
</tr>
<tr>
<td>PID</td>
<td>0.6( K_{cr} )</td>
<td>0.5( T_{cr} )</td>
<td>0.125( T_{cr} )</td>
</tr>
</tbody>
</table>
Problem: Design a PID controller to control the angle position of a DC motor, providing that by experiment the critical gain of the system is 20 and the critical cycle is $T=1$ sec.

Solution:

According to the given data:

$$K_{cr} = 20$$
$$T_{cr} = 1\text{ sec}$$

Applying Zeigler – Nichols method 2:

$$K_P = 0.6K_{cr} = 0.6 \times 20 = 12$$
$$T_I = 0.5T_{cr} = 0.5 \times 1 = 0.5\text{ sec}$$
$$T_D = 0.125T_{cr} = 0.125 \times 1 = 0.125\text{ sec}$$

$$G_{PID}(s) = 12 \left(1 + \frac{1}{0.125s} + 0.5s\right)$$
Analytical method for designing PID controller

- **Step 1:** Establish equation(s) representing the relationship between the controller to be designed and the desired performances.

- **Step 2:** Solve the equation(s) obtained in step 1 for the parameter(s) of the controller.
Example: Design PID controller so that the control system satisfies the following requirements:

- Closed-loop complex poles with $\xi=0.5$ and $\omega_n=8$.
- Velocity constant $K_V = 100$.

Solution: The transfer function of the PID controller to be designed

$$G_C(s) = K_P + \frac{K_I}{s} + K_D s$$
Analytical method for designing PID controller (cont’)

velocity constant of the controlled system:

\[ K_V = \lim_{{s \to 0}} s G_C(s) G(s) = \lim_{{s \to 0}} s \left( K_P + \frac{K_I}{s} + K_D s \right) \left( \frac{100}{s^2 + 10s + 100} \right) \]

\[ \Rightarrow K_V = K_I \]

According to the design requirement: \( K_V = 100 \)

\[ \Rightarrow K_I = 100 \]

The characteristic equation of the controlled system:

\[ 1 + \left( K_P + \frac{K_I}{s} + K_D s \right) \left( \frac{100}{s^2 + 10s + 100} \right) = 0 \]

\[ \Rightarrow s^3 + (10 + 100K_D)s^2 + (100 + 100K_P)s + 100K_I = 0 \quad (1) \]
Analytical method for designing PID controller (cont’)

★ The desired characteristic equation:

\[(s + a)(s^2 + 2\xi\omega_n s + \omega_n^2) = 0\]

⇒ \[(s + a)(s^2 + 8s + 64) = 0\]

⇒ \[s^3 + (a + 8)s^2 + (8a + 64)s + 64a = 0\] (2)

★ Balancing the coefficients of the equations (1) and (2), we have:

\[
\begin{align*}
10 + 100K_D &= a + 8 \\
100 + 100K_P &= 8a + 64 \\
100K_I &= 64a
\end{align*}
\]

⇒ \[
\begin{align*}
a &= 156.25 \\
K_P &= 12.14 \\
K_D &= 1.54
\end{align*}
\]

Conclusion: \[G_C(s) = 12.64 + \frac{100}{s} + 1.54s\]
Manual tuning of PID controllers

- Effect of increasing a parameter of PID controller independently on closed-loop performance:

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Rise time</th>
<th>POT</th>
<th>Settling time</th>
<th>Steady-state error</th>
<th>Stability</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_P$</td>
<td>Decrease</td>
<td>Increase</td>
<td>Small change</td>
<td>Decrease</td>
<td>Degrade</td>
</tr>
<tr>
<td>$K_I$</td>
<td>Decrease</td>
<td>Increase</td>
<td>Increase</td>
<td>Eliminate</td>
<td>Degrade</td>
</tr>
<tr>
<td>$K_D$</td>
<td>Minor change</td>
<td>Decrease</td>
<td>Decrease</td>
<td>No effect</td>
<td>Improve if $K_D$ small</td>
</tr>
</tbody>
</table>
Manual tuning of PID controllers (cont.)

A procedure for manual tuning of PID controllers:

1. Set $K_I$ and $K_D$ to 0, gradually increase $K_P$ to the critical gain $K_{cr}$ (i.e. the gain makes the closed-loop system oscillate)

2. Set $K_P \approx K_{cr}/2$

3. Gradually increase $K_I$ until the steady-state error is eliminated in a sufficient time for the process (Note that too much $K_I$ will cause instability).

4. Increase $K_D$ if needed to reduce POT and settling time (Note that too much $K_D$ will cause excessive response and overshoot)
Control systems design in state-space using pole placement method
Controllability

- Consider a system:
  \[
  \begin{align*}
  \dot{x}(t) &= Ax(t) + Bu(t) \\
  y(t) &=Cx(t)
  \end{align*}
  \]

- The system is complete state controllable if there exists an unconstrained control law \(u(t)\) that can drive the system from an initial state \(x(t_0)\) to an arbitrarily final state \(x(t_f)\) in a finite time interval \(t_0 \leq t \leq t_f\). Qualitatively, the system is state controllable if each state variable can be influenced by the input.

Signal flow graph of an incomplete state controllable system
Controllability condition

★ System:
\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) \\
y(t) &= Cx(t)
\end{align*}
\]

★ Controllability matrix
\[
C = \begin{bmatrix} B & AB & A^2B & \ldots & A^{n-1}B \end{bmatrix}
\]

★ The necessary and sufficient condition for the controllability is:
\[
\text{rank}(C) = n
\]

★ Note: we use the term “controllable” instead of “complete state controllable” for short.
Consider a system
\[ \begin{cases} \dot{x}(t) = Ax(t) + Bu(t) \\ y(t) =Cx(t) \end{cases} \]
where:
\[ A = \begin{bmatrix} 0 & 1 \\ -2 & -3 \end{bmatrix}, \quad B = \begin{bmatrix} 5 \\ 2 \end{bmatrix}, \quad C = \begin{bmatrix} 1 & 3 \end{bmatrix} \]

Evaluate the controllability of the system.

**Solution:** Controllability matrix:
\[ \mathcal{C} = \begin{bmatrix} B & AB \end{bmatrix} \Rightarrow \mathcal{C} = \begin{bmatrix} 5 & 2 \\ 2 & -16 \end{bmatrix} \]

Because:
\[ \det(\mathcal{C}) = -84 \Rightarrow \text{rank}(\mathcal{C}) = 2 \]

⇒ The system is controllable
Consider a system described by the state equations:

\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) \\
y(t) &= Cx(t)
\end{align*}
\]

The state feedback controller:

\[
u(t) = r(t) - Kx(t)
\]

The state equations of the closed-loop system:

\[
\begin{align*}
\dot{x}(t) &= [A - BK]x(t) + Br(t) \\
y(t) &= Cx(t)
\end{align*}
\]
Pole placement method

If the system is controllable, then it is possible to determine the feedback gain $K$ so that the closed-loop system has the poles at any location.

★ **Step 1**: Write the characteristic equation of the closed-loop system

$$\det[sI - A + BK] = 0$$  \hfill (1)

★ **Step 2**: Write the desired characteristic equation:

$$\prod_{i=1}^{n} (s - p_i) = 0$$  \hfill (2)

$p_i, \ (i = 1, n)$ are the desired poles

★ **Step 3**: Balance the coefficients of the equations (1) and (2), we can find the state feedback gain $K$. 
Pole placement method – Example

**Problem**: Given a system described by the state-state equation:

\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) \\
y(t) &= Cx(t)
\end{align*}
\]

\[
A = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -4 & -7 & -3 \end{bmatrix}, \quad B = \begin{bmatrix} 0 \\ 3 \\ 1 \end{bmatrix}, \quad C = \begin{bmatrix} 0 & 0 & 1 \end{bmatrix}
\]

**Determine the state feedback controller** \( u(t) = r(t) - Kx(t) \) so that the closed-loop system has complex poles with \( \xi = 0.6; \omega_n = 10 \) and the third pole at \(-20\).
Solution

The characteristic equation of the closed-loop system:
\[
\det[sI - A + BK] = 0
\]

\[
\Rightarrow \det(s \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} - \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -4 & -7 & -3 \end{bmatrix} + \begin{bmatrix} 0 \\ 3 \\ 1 \end{bmatrix} \begin{bmatrix} k_1 \\ k_2 \\ k_3 \end{bmatrix}) = 0
\]

\[
\Rightarrow s^3 + (3 + 3k_2 + k_3)s^2 + (7 + 3k_1 + 10k_2 - 21k_3)s + (4 + 10k_1 - 12k_3) = 0 \quad (1)
\]

The desired characteristic equation:

\[
(s + 20)(s^2 + 2\xi\omega_n s + \omega_n^2) = 0
\]

\[
\Rightarrow s^3 + 32s^2 + 340s + 2000 = 0 \quad (2)
\]
Balance the coefficients of the equations (1) and (2), we have:

\[
\begin{align*}
3 + 3k_2 + k_3 &= 32 \\
7 + 3k_1 + 10k_2 - 21k_3 &= 340 \\
4 + 10k_1 - 12k_2 &= 2000
\end{align*}
\]

Solve the above set of equations, we have:

\[
\begin{align*}
k_1 &= 220,578 \\
k_2 &= 3,839 \\
k_3 &= 17,482
\end{align*}
\]

Conclusion: \( K = [220,578 \ 3,839 \ 17,482] \)
Design of state estimators
The concept of state estimation

To be able to implement state feedback control system, it is required to measure all the states of the system.

However, in some applications, we can only measure the output, but cannot measure the states of the system.

The problem is to estimate the states of the system from the output measurement.

⇒ State estimator (or state observer)
Observability

Consider a system:

\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) \\
y(t) &= Cx(t)
\end{align*}
\]

The system is complete state observable if given the control law \( u(t) \) and the output signal \( y(t) \) in a finite time interval \( t_0 \leq t \leq t_f \), it is possible to determine the initial states \( x(t_0) \).

Qualitatively, the system is state observable if all state variable \( x(t) \) influences the output \( y(t) \).

Signal flow graph of an incomplete state observable system
Observability condition

★ System

\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) \\
y(t) &= Cx(t)
\end{align*}
\]

It is necessary to estimate the state \( \hat{x}(t) \) from mathematical model of the system and the input-output data.

★ Observability matrix:

\[
\Phi = \begin{bmatrix} C \\ CA \\ CA^2 \\ \vdots \\ CA^{n-1} \end{bmatrix}
\]

★ The necessary and sufficient condition for the observability is:

\[ \text{rank}(\Phi) = n \]
Observability – Example

Consider the system
\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) \\
y(t) &= Cx(t)
\end{align*}
\]

where:
\[
A = \begin{bmatrix} 0 & 1 \\ -2 & -3 \end{bmatrix}, \quad B = \begin{bmatrix} 1 \\ 2 \end{bmatrix}, \quad C = [1 \ 3]
\]

Evaluate the observability of the system.

Solution: Observability matrix:
\[
\mathcal{O} = \begin{bmatrix} C \\ CA \end{bmatrix} \quad \Rightarrow \quad \mathcal{O} = \begin{bmatrix} 1 & 3 \\ -6 & -8 \end{bmatrix}
\]

Because \( \det(\mathcal{O}) = 10 \) \( \Rightarrow \) \( \text{rank}(\mathcal{O}) = 2 \)

\( \Rightarrow \) The system is observable
State estimator:

\[
\begin{align*}
\dot{x}(t) &= Ax(t) + Bu(t) + L(y(t) - \hat{y}(t)) \\
\hat{y}(t) &= C\hat{x}(t)
\end{align*}
\]

where:

\[
L = [l_1 \ l_2 \ \ldots \ l_n]^T
\]
Design of state estimators

★ Requirements:

- The state estimator must be stable, estimation error should approach to zero.
- Dynamic response of the state estimator should be fast enough in comparison with the dynamic response of the control loop.

★ It is required to chose $L$ satisfying:

- All the roots of the equation $\det(sI - A + LC) = 0$ locates in the half-left s-plane.
- The roots of the equation $\det(sI - A + LC) = 0$ are further from the imaginary axis than the roots of the equation $\det(sI - A + BK) = 0$

★ Depending on the design of $L$, we have different state estimator:

- Luenberger state observer
- Kalman filter
Procedure for designing the Luenberger state observer

- **Step 1**: Write the characteristic equation of the state observer

\[
\det[sI - A + LC] = 0 \tag{1}
\]

- **Step 1**: Write the desired characteristic equation:

\[
\prod_{i=1}^{n} (s - p_i) = 0 \tag{2}
\]

\(p_i, \ (i = 1, n)\) are the desired poles of the state estimator

- **Step 3**: Balance the coefficients of the characteristic equations (1) and (2), we can find the gain \(L\).
**Problem**: Given a system described by the state equation:

\[
\begin{cases}
\dot{x}(t) = Ax(t) + Bu(t) \\
y(t) = Cx(t)
\end{cases}
\]

where

\[
A = \begin{bmatrix}
0 & 1 & 0 \\
0 & 0 & 1 \\
-4 & -7 & -3
\end{bmatrix}, \quad
B = \begin{bmatrix}
0 \\
3 \\
1
\end{bmatrix}, \quad
C = \begin{bmatrix}
1 & 0 & 0
\end{bmatrix}
\]

Assuming that the states of the system cannot be directly measured. Design the Luenberger state estimator so that the poles of the state estimator lying at \(-20, -20\) and \(-50\).
**Solution**

The characteristic equation of the Luenberger state estimator:

\[
\det [sI - A + LC] = 0
\]

\[
\Rightarrow \det \left( \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} - \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -4 & -7 & -3 \end{bmatrix} + \begin{bmatrix} l_1 \\ l_2 \\ l_3 \end{bmatrix} \right) = 0
\]

\[
\Rightarrow s^3 + (l_1 + 3)s^2 + (3l_1 + l_2 + 7)s + (7l_1 + 5l_2 + l_3 + 4) = 0 \quad (1)
\]

The desired characteristic equation:

\[
(s + 20)^2 (s + 50) = 0
\]

\[
\Rightarrow s^3 + 90s^2 + 2400s + 20000 = 0 \quad (2)
\]
Design of state estimators – Example (cont’)

**Balancing the coefficients of the equ. (1) and (2) leads to:**

\[
\begin{align*}
    l_1 + 3 &= 90 \\
    3l_1 + l_2 + 7 &= 2400 \\
    7l_1 + 3l_2 + l_3 + 4 &= 20000
\end{align*}
\]

**Solve the above set of equations, we have:**

\[
\begin{align*}
    l_1 &= 87 \\
    l_2 &= 2132 \\
    l_3 &= 12991
\end{align*}
\]

**Conclusion**

\[ L = \begin{bmatrix} 87 & 2132 & 12991 \end{bmatrix}^T \]