

# Lecture 3

## *Chapter 4: Elements of Linear System Theory*

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# Elements of Linear System Theory

## 3.1 System Descriptions [4.1]

Let  $f(u)$  be a linear operator,  $u_1$  and  $u_2$  two independent variables, and  $\alpha_1$  and  $\alpha_2$  two real scalars, then

$$(3.1) \quad f(\alpha_1 u_1 + \alpha_2 u_2) = \alpha_1 f(u_1) + \alpha_2 f(u_2)$$

### 3.1.1 State Space Representation

$$(3.2) \quad \dot{x}(t) = Ax(t) + Bu(t)$$

$$(3.3) \quad y(t) = Cx(t) + Du(t)$$

or:

$$(3.4) \quad \begin{bmatrix} \dot{x} \\ y \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} x \\ u \end{bmatrix}, \quad G \stackrel{s}{=} \left[ \begin{array}{c|c} A & B \\ \hline C & D \end{array} \right]$$

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System (3.2)–(3.3) is *not* a unique description of the input-output behaviour of a linear system. Define new states  $q = Sx$ , i.e.  $x = S^{-1}q$ . Equivalent state-space realization (i.e., with same input-output behaviour):

$$(3.5) \quad A_q = SAS^{-1}, \quad B_q = SB, \quad C_q = CS^{-1}, \quad D_q = D$$

Dynamical system response  $x(t)$  for  $t \geq t_0$

$$(3.6) \quad x(t) = e^{A(t-t_0)}x(t_0) + \int_{t_0}^t e^{A(t-\tau)}Bu(\tau)d\tau$$

For a system with disturbances  $d$  and noise  $n$ :

$$(3.7) \quad \dot{x} = Ax + Bu + Ed$$

$$(3.8) \quad y = Cx + Du + Fd + n$$

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$$(3.9) \quad e^{At} = I + \sum_{k=1}^{\infty} \frac{(At)^k}{k!}$$

Let  $A_q = SAS^{-1} = \Lambda = \text{diag}\{\lambda_i\}$  be diagonal then

$$e^{At} = S^{-1}\{\text{diag}(e^{\lambda_i t})\}S$$

where  $e^{\lambda_i t}$  is the *mode* associated with eigenvalue  $\lambda_i(A)$ .

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## 3.1.2 Impulse response

The impulse response matrix is

$$(3.10) \quad g(t) = \begin{cases} 0 & t < 0 \\ Ce^{At}B + D\delta(t) & t \geq 0 \end{cases}$$

With initial state  $x(0) = 0$ , the dynamic response to an arbitrary input  $u(t)$  is

$$(3.11) \quad y(t) = g(t) * u(t) = \int_0^t g(t - \tau)u(\tau)d\tau$$

where  $*$  denotes the convolution operator.

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## 3.1.3 Transfer function representation (Laplace)

$$(3.12) \quad G(s) = \int_0^{\infty} g(t)e^{-st} dt$$

Laplace transforms of (3.2) and (3.3) become for  $x(0) = 0$

$$sX(s) = AX(s) + BU(s) \Rightarrow$$

$$(3.13) \quad \Rightarrow X(s) = (sI - A)^{-1}BU(s)$$

$$Y(s) = CX(s) + DU(s) \Rightarrow$$

$$(3.14) \quad \Rightarrow Y(s) = \underbrace{(C(sI - A)^{-1}B + D)}_{G(s)} u(s)$$

where  $G(s)$  is the transfer function matrix.

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Equivalently,

$$(3.15) G(s) = \frac{1}{\det(sI - A)} [C \operatorname{adj}(sI - A) B + D \det(sI - A)]$$

From Appendix A.2.1

$$(3.16) \quad \det(sI - A) = \prod_{i=1}^n \lambda_i(sI - A) = \prod_{i=1}^n (s - \lambda_i(A))$$

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## 3.1.4 State-space realizations [4.1.6]

Inverse system. For a square  $G(s)$  we have

$$(3.17) \quad G^{-1} \stackrel{s}{=} \left[ \begin{array}{c|c} A - BD^{-1}C & BD^{-1} \\ \hline -D^{-1}C & D^{-1} \end{array} \right]$$

If  $D = 0$ , set  $D = \varepsilon I$ . Be careful not to introduce RHP zeros with this modification.

**Improper systems cannot be represented in state space form.**



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**Realization of SISO transfer functions.**

$$(3.18) \quad G(s) = \frac{\beta_{n-1}s^{n-1} + \dots + \beta_1s + \beta_0}{s^n + a_{n-1}s^{n-1} + \dots + a_1s + a_0}$$

$y(s) = G(s)u(s)$  corresponds to

$$y^n(t) + a_{n-1}y^{n-1}(t) + \dots + a_1y'(t) + a_0y(t) = \beta_{n-1}u^{n-1}(t) + \dots + \beta_1u'(t) + \beta_0u(t)$$

(3.19)

where  $y^{n-1}(t)$  and  $u^{n-1}(t)$  represent  $n - 1$ 'th order derivatives, etc.

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Write this as

$$y^n = (-a_{n-1}y^{n-1} + \beta_{n-1}u^{n-1}) + \dots$$
$$\dots + (-a_1y' + \beta_1u') + \underbrace{(-a_0y + \beta_0u)}_{x'_n}$$
$$\underbrace{\hspace{10em}}_{x_{n-1}^2}$$

With the notation  $\dot{x} \equiv x'(t) = dx/dt$ , we get

$$\begin{aligned}\dot{x}_n &= -a_0x_1 + \beta_0u \\ \dot{x}_{n-1} &= -a_1x_1 + x_n + \beta_1u \\ &\vdots \\ \dot{x}_1 &= -a_{n-1}x_1 + x_2 + \beta_{n-1}u\end{aligned}$$

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corresponding to the realization (*observer canonical form*)

$$(3.20) \quad A = \begin{bmatrix} -a_{n-1} & 1 & 0 & \cdots & 0 & 0 \\ -a_{n-2} & 0 & 1 & & 0 & 0 \\ \vdots & \vdots & & \ddots & \vdots & \\ -a_2 & 0 & 0 & & 1 & 0 \\ -a_1 & 0 & 0 & \cdots & 0 & 1 \\ -a_0 & 0 & 0 & \cdots & 0 & 0 \end{bmatrix}, \quad B = \begin{bmatrix} \beta_{n-1} \\ \beta_{n-2} \\ \vdots \\ \beta_2 \\ \beta_1 \\ \beta_0 \end{bmatrix}$$

$$C = [1 \quad 0 \quad 0 \quad \cdots \quad 0 \quad 0]$$

**Example:** To obtain the state-space realization of  $G(s) = \frac{s-a}{s+a}$ , first bring out a constant term by division to get

$$G(s) = \frac{s-a}{s+a} = \frac{-2a}{s+a} + 1$$

Thus  $D = 1$ . Then (3.20) yields  $A = -a$ ,  $B = -2a$  and  $C = 1$ .

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**Example:** Ideal PID-controller

$$(3.21) \quad K(s) = K_c \left( 1 + \frac{1}{\tau_I s} + \tau_D s \right) = K_c \frac{\tau_I \tau_D s^2 + \tau_I s + 1}{\tau_I s}$$

⇒ Improper ⇒ no realization

Proper PID controller

$$(3.22) \quad K(s) = K_c \left( 1 + \frac{1}{\tau_I s} + \frac{\tau_D s}{1 + \epsilon \tau_D s} \right), \epsilon \leq 0.1$$

Four common realizations

$$(3.23) \quad D = K_c \frac{1 + \epsilon}{\epsilon}$$

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## 1. Diagonalized form (Jordan canonical form)

$$(3.24) \quad A = \begin{bmatrix} 0 & 0 \\ 0 & -\frac{1}{\epsilon\tau_D} \end{bmatrix}, \quad B = \begin{bmatrix} K_c/\tau_I \\ K_c/(\epsilon^2\tau_D) \end{bmatrix}, \quad C = [1 \quad -1]$$

## 2. Observability canonical form

$$(3.25) \quad A = \begin{bmatrix} 0 & 1 \\ 0 & -\frac{1}{\epsilon\tau_D} \end{bmatrix}, \quad B = \begin{bmatrix} \gamma_1 \\ \gamma_2 \end{bmatrix}, \quad C = [1 \quad 0]$$

$$(3.26) \quad \text{where } \gamma_1 = K_c\left(\frac{1}{\tau_I} - \frac{1}{\epsilon^2\tau_D}\right), \quad \gamma_2 = \frac{K_c}{\epsilon^3\tau_D^2}$$

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## 3. Controllability canonical form

$$(3.27) \quad A = \begin{bmatrix} 0 & 0 \\ 1 & -\frac{1}{\epsilon\tau_D} \end{bmatrix}, \quad B = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \quad C = [\gamma_1 \quad \gamma_2]$$

## 4. Observer canonical form in (3.20)

$$(3.28) \quad A = \begin{bmatrix} -\frac{1}{\epsilon\tau_D} & 1 \\ 0 & 0 \end{bmatrix}, \quad B = \begin{bmatrix} \beta_1 \\ \beta_0 \end{bmatrix}, \quad C = [1 \quad 0]$$

$$(3.29) \quad \text{where } \beta_0 = \frac{K_c}{\epsilon\tau_I\tau_D}, \quad \beta_1 = K_c \frac{\epsilon^2\tau_D - \tau_I}{\epsilon^2\tau_I\tau_D}$$

**Note:** Transfer function offers more immediate insight.

# Elements of Linear System Theory

## 3.2 State controllability and state observability

**State controllability.** The dynamical system  $\dot{x} = Ax + Bu$ , or equivalently the pair  $(A, B)$ , is said to be state controllable if, for any initial state  $x(0) = x_0$ , any time  $t_1 > 0$  and any final state  $x_1$ , there exists an input  $u(t)$  such that  $x(t_1) = x_1$ . Otherwise the system is said to be state uncontrollable.

1. The pair:  $(A, B)$  is state controllable if and only if the controllability matrix

$$(3.30) \quad C \triangleq [B \quad AB \quad A^2B \quad \dots \quad A^{n-1}B]$$

has rank  $n$  (full row rank). Here  $n$  is the number of states.

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2. From (3.6) one can verify that for  $x(t_1) = x_1$

$$(3.31) \quad u(t) = -B^T e^{A^T(t_1-t)} W_c(t_1)^{-1} (e^{At_1} x_0 - x_1)$$

where  $W_c(t)$  is the Gramian matrix at time  $t$ ,

$$(3.32) \quad W_c(t) \triangleq \int_0^t e^{A\tau} B B^T e^{A^T \tau} d\tau$$

Thus  $(A, B)$  is state controllable if and only if  $W_c(t)$  has full rank (and thus is positive definite) for any  $t > 0$ . For a stable system ( $A$  is stable) check only  $P \triangleq W_c(\infty)$ ,

$$(3.33) \quad P \triangleq \int_0^\infty e^{A\tau} B B^T e^{A^T \tau} d\tau$$



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$P$  may also be obtained as the solution to the Lyapunov equation

$$(3.34) \quad AP + PA^T = -BB^T$$

- Let  $p_i$  be the  $i$ 'th eigenvalue of  $A$  and  $q_i$  the corresponding left eigenvector,  $q_i^H A = p_i q_i^H$ . Then the system is state controllable if and only if  $q_i^H B \neq 0, \forall i$ .

**Example:**

$$A = \begin{bmatrix} -2 & -2 \\ 0 & -4 \end{bmatrix}, \quad B = \begin{bmatrix} 1 \\ 1 \end{bmatrix}, \quad C = [1 \quad 0], \quad D = 0$$

The transfer function

$$G(s) = C(sI - A)^{-1}B = \frac{1}{s + 4}$$

has only one state.

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1. The controllability matrix has two linearly dependent rows:

$$C = [B \quad AB] = \begin{bmatrix} 1 & -4 \\ 1 & -4 \end{bmatrix}.$$

2. The controllability Gramian is singular

$$P = \begin{bmatrix} 0.125 & 0.125 \\ 0.125 & 0.125 \end{bmatrix}$$

3.  $p_1 = -2$  and  $p_2 = -4$ ,  $q_1 = [0.707 \quad -0.707]^T$  and  $q_2 = [0 \quad 1]^T$ .

$$q_1^H B = 0, \quad q_2^H B = 1$$

the first mode (eigenvalue) is not state controllable.

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Controllability is a system-theoretic concept important for computation and realizations; but no practical insight:

1. It says nothing about how the states behave, e.g. it does not imply that one can *hold* (as  $t \rightarrow \infty$ ) the states at a given value.
2. Required inputs may be very large with sudden changes.
3. Some states may be of no practical importance.
4. Existence result which provides no “degree of controllability”.

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**Definition State observability.** The dynamical system  $\dot{x} = Ax + Bu, y = Cx + Du$  (or the pair  $(A, C)$ ) is said to be state observable if, for any time  $t_1 > 0$ , the initial state  $x(0) = x_0$  can be determined from the time history of the input  $u(t)$  and the output  $y(t)$  in the interval  $[0, t_1]$ . Otherwise the system, or  $(A, C)$ , is said to be state unobservable.

1.  $(A, C)$  is state observable if and only if the observability matrix

$$(3.35) \quad \mathcal{O} \triangleq \begin{bmatrix} C \\ CA \\ \vdots \\ CA^{n-1} \end{bmatrix}$$

has rank  $n$  (full column rank).

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2. For a stable system the observability Gramian

$$(3.36) \quad Q \triangleq \int_0^{\infty} e^{A^T \tau} C^T C e^{A \tau} d\tau$$

must have full rank  $n$  (and thus be positive definite).  $Q$  can also be found as the solution to the following Lyapunov equation

$$(3.37) \quad A^T Q + Q A = -C^T C$$

3. Let  $p_i$  be the  $i$ 'th eigenvalue of  $A$  and  $t_i$  the corresponding eigenvector,  $A t_i = p_i t_i$ . Then the system is state observable if and only if  $C t_i \neq 0, \forall i$ .

Observability is a system theoretical concept but may not give practical insight.

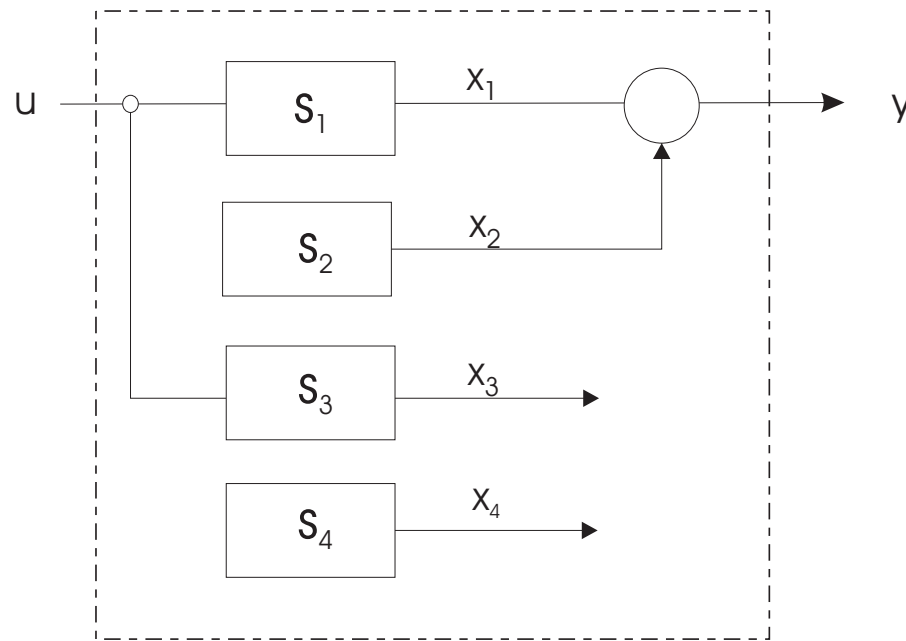
# Elements of Linear System Theory

## Kalman's decomposition

By performing an appropriate coordinate transformation, any system can be reduced to a decomposition indicating the states that are or aren't controllable and/or observable.

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \\ \dot{x}_4 \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} & 0 & 0 \\ 0 & A_{22} & 0 & 0 \\ A_{31} & A_{32} & A_{33} & A_{34} \\ 0 & A_{42} & 0 & A_{44} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} + \begin{bmatrix} B_1 \\ 0 \\ B_3 \\ 0 \end{bmatrix} u$$
$$y = [C_1 \quad C_2 \quad 0 \quad 0] \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix}$$

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A state-space realization  $(A, B, C, D)$  of  $G(s)$  is said to be a **minimal realization** of  $G(s)$  if  $A$  has the smallest possible dimension (i.e., the fewest number of states). The smallest dimension is called the **McMillan degree** of  $G(s)$ . A mode is hidden if it is not state controllable or observable and thus not appear in the minimal realization. The state-space realization is minimal if and only if  $(A, B)$  is controllable and  $(A, C)$  is observable.

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## 3.3 Stability [4.3]

### Definition

A system is **(internally) stable** if none of its components contains hidden unstable modes and the injection of bounded external signals at any place in the system results in bounded output signals measured anywhere in the system. “internal”, i.e. all the states must be stable not only inputs/outputs.

### Definition

**State stabilizable, state detectable and hidden unstable modes.** A system is state stabilizable if all unstable modes are state controllable. A system is state detectable if all unstable modes are state observable. A system with unstabilizable or undetectable modes is said to contain hidden unstable modes.



# Elements of Linear System Theory

## 3.4 Poles [4.4]

### Definition

**Poles.** The poles  $p_i$  of a system with state-space description (3.2)–(3.3) are the eigenvalues  $\lambda_i(A)$ ,  $i = 1, \dots, n$  of the matrix  $A$ . The pole or characteristic polynomial  $\phi(s)$  is defined as  $\phi(s) \triangleq \det(sI - A) = \prod_{i=1}^n (s - p_i)$ . Thus the poles are the roots of the characteristic equation

$$(3.38) \quad \phi(s) \triangleq \det(sI - A) = 0$$

# Elements of Linear System Theory

## 3.4.1 Poles and stability

**Theorem 1** *A linear dynamic system  $\dot{x} = Ax + Bu$  is stable if and only if all the poles are in the open left-half plane (LHP), that is,  $\text{Re}\{\lambda_i(A)\} < 0, \forall i$ . A matrix  $A$  with such a property is said to be “stable” or Hurwitz.*

## 3.4.2 Poles from transfer functions

**Theorem 2** *The pole polynomial  $\phi(s)$  corresponding to a minimal realization of a system with transfer function  $G(s)$ , is the least common denominator of all non-identically-zero minors of all orders of  $G(s)$ .*

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**Example:**

$$(3.39) \quad G(s) = \frac{1}{1.25(s+1)(s+2)} \begin{bmatrix} s-1 & s \\ -6 & s-2 \end{bmatrix}$$

The minors of order 1 are the four elements all have  $(s+1)(s+2)$  in the denominator.

Minor of order 2

$$(3.40) \quad \det G(s) = \frac{(s-1)(s-2) + 6s}{1.25^2(s+1)^2(s+2)^2} = \frac{1}{1.25^2(s+1)(s+2)}$$

Least common denominator of all the minors:

$$(3.41) \quad \phi(s) = (s+1)(s+2)$$

Minimal realization has two poles:  $s = -1$ ;  $s = -2$ .

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**Example:** Consider the  $2 \times 3$  system, with 3 inputs and 2 outputs,

$$G(s) = \frac{1}{(s+1)(s+2)(s-1)}^*$$

$$(3.42) \quad * \begin{bmatrix} (s-1)(s+2) & 0 & (s-1)^2 \\ -(s+1)(s+2) & (s-1)(s+1) & (s-1)(s+1) \end{bmatrix}$$

Minors of order 1:

$$(3.43) \quad \frac{1}{s+1}, \frac{s-1}{(s+1)(s+2)}, \frac{-1}{s-1}, \frac{1}{s+2}, \frac{1}{s+2}$$

Minor of order 2 corresponding to the deletion of column 2:

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$$M_2 = \frac{(s-1)(s+2)(s-1)(s+1) + (s+1)(s+2)(s-1)^2}{((s+1)(s+2)(s-1))^2} =$$
$$(3.44) \quad = \frac{2}{(s+1)(s+2)}$$

The other two minors of order two are

$$(3.45) \quad M_1 = \frac{-(s-1)}{(s+1)(s+2)^2}, \quad M_3 = \frac{1}{(s+1)(s+2)}$$

Least common denominator:

$$(3.46) \quad \phi(s) = (s+1)(s+2)^2(s-1)$$

The system therefore has four poles:  $s = -1$ ,  $s = 1$  and two at  $s = -2$ . Note MIMO-poles are essentially the poles of the elements. A procedure is needed to determine multiplicity.

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## 3.5 Zeros [4.5]

- SISO system: zeros  $z_i$  are the solutions to  $G(z_i) = 0$ .

In general, zeros are values of  $s$  at which  $G(s)$  loses rank.

### Example

$$\left[ Y = \frac{s + 2}{s^2 + 7s + 12} U \right]$$

Compute the response when

$$u(t) = e^{-2t}, y(0) = 0, \dot{y}(0) = -1$$

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$$\mathcal{L}\{u(t)\} = \frac{1}{s+2}$$

$$s^2Y - sy(0) - \dot{y}(0) + 7sY - 7y(0) + 12Y = 1$$

$$s^2Y + 7sY + 12Y + 1 = 1$$

$$\Rightarrow Y(s) = 0$$

Assumption:  $g(s)$  has a zero  $z$ ,  $g(z) = 0$ .

Then for input  $u(t) = u_0 e^{zt}$  the output is  $y(t) \equiv 0$ ,  $t > 0$ .  
(with appropriate initial conditions)

# Elements of Linear System Theory

## 3.5.1 Zeros from state-space realizations [4.5.1]

The state-space equations of a system can be written as:

$$(3.47) \quad P(s) \begin{bmatrix} x \\ u \end{bmatrix} = \begin{bmatrix} 0 \\ y \end{bmatrix}, \quad P(s) = \begin{bmatrix} sI - A & -B \\ C & D \end{bmatrix}$$

The zeros are then the values  $s = z$  for which the polynomial system matrix  $P(s)$  loses rank, resulting in zero output for some non-zero input

$$\begin{bmatrix} zI - A & -B \\ C & D \end{bmatrix} \begin{bmatrix} x_z \\ u_z \end{bmatrix} = 0$$



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The zeros are the solutions of

$$\det \begin{bmatrix} zI - A & -B \\ C & D \end{bmatrix} = 0$$

**MATLAB**

```
zero = tzero(A,B,C,D)
```

# Elements of Linear System Theory

## 3.5.2 Zeros from transfer functions [4.5.2]

**Definition Zeros.**  $z_i$  is a zero of  $G(s)$  if the rank of  $G(z_i)$  is less than the normal rank of  $G(s)$ . The zero polynomial is defined as  $z(s) = \prod_{i=1}^{n_z} (s - z_i)$  where  $n_z$  is the number of finite zeros of  $G(s)$ .

**Theorem** The zero polynomial  $z(s)$ , corresponding to a minimal realization of the system, is the greatest common divisor of all the numerators of all order- $r$  minors of  $G(s)$ , where  $r$  is the normal rank of  $G(s)$ , provided that these minors have been adjusted in such a way as to have the pole polynomial  $\phi(s)$  as their denominators.

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## Example

$$(3.48) \quad G(s) = \frac{1}{s+2} \begin{bmatrix} s-1 & 4 \\ 4.5 & 2(s-1) \end{bmatrix}$$

The normal rank of  $G(s)$  is 2.

Minor of order 2:  $\det G(s) = \frac{2(s-1)^2 - 18}{(s+2)^2} = 2 \frac{s-4}{s+2}$ .

Pole polynomial:  $\phi(s) = s + 2$ .

Zero polynomial:  $z(s) = s - 4$ .

**Note** Multivariable zeros have no relationship with the zeros of the transfer function elements.

# Elements of Linear System Theory

## Example

$$(3.49) \quad G(s) = \frac{1}{1.25(s+1)(s+2)} \begin{bmatrix} s-1 & s \\ -6 & s-2 \end{bmatrix}$$

Minor of order 2 is the determinant

$$(3.50) \quad \det G(s) = \frac{(s-1)(s-2) + 6s}{1.25^2(s+1)^2(s+2)^2} = \frac{1}{1.25^2(s+1)(s+2)}$$

$$\phi(s) = 1.25^2(s+1)(s+2)$$

Zero polynomial = numerator of (3.50)

⇒ no multivariable zeros.

# Elements of Linear System Theory

## Example

$$(3.51) \quad G(s) = \begin{bmatrix} \frac{s-1}{s+1} & \frac{s-2}{s+2} \end{bmatrix}$$

- The normal rank of  $G(s)$  is 1
- no value of  $s$  for which  $G(s) = 0 \Rightarrow G(s)$  has no zeros.

# Elements of Linear System Theory

## 3.6 More on poles and zeros[4.6]

### 3.6.1 Directions of poles and zeros

Let  $G(s) = C(sI - A)^{-1}B + D$ .

**Zero directions.** Let  $G(s)$  have a zero at  $s = z$ . Then  $G(s)$  loses rank at  $s = z$ , and there exist non-zero vectors  $u_z$  and  $y_z$  such that

$$(3.52) \quad G(z)u_z = 0, \quad y_z^H G(z) = 0$$

$u_z$  = input zero direction

$y_z$  = output zero direction

$y_z$  gives information about which output (or combination of outputs) may be difficult to control.

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**Pole directions.** Let  $G(s)$  have a pole at  $s = p$ . Then  $G(p)$  is infinite, and we may write

$$(3.53) \quad G(p)u_p = \infty, \quad y_p^H G(p) = \infty$$

$u_p$  = input pole direction

$y_p$  = output pole direction.

**SVD:**

$$G(z/p) = U\Sigma V^H$$

$u_z$  = last column in  $V$ ,  $y_z$  = last column of  $U$   
(corresponding to the zero singular value of  $G(z)$ )

$u_p$  = first column in  $V$ ,  $y_p$  = first column of  $U$   
(corresponding to the infinite singular value of  $G(p)$ )

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## Example

Plant in (3.48) has a RHP-zero at  $z = 4$  and a LHP-pole at  $p = -2$ .

$$\begin{aligned} G(z) &= G(4) = \frac{1}{6} \begin{bmatrix} 3 & 4 \\ 4.5 & 6 \end{bmatrix} \\ &= \frac{1}{6} \begin{bmatrix} 0.55 & -0.83 \\ 0.83 & 0.55 \end{bmatrix} \begin{bmatrix} 9.01 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} 0.6 & -0.8 \\ 0.8 & 0.6 \end{bmatrix}^H \end{aligned}$$

$$(3.54) \quad u_z = \begin{bmatrix} -0.80 \\ 0.60 \end{bmatrix} \quad y_z = \begin{bmatrix} -0.83 \\ 0.55 \end{bmatrix}$$

For pole directions consider

$$(3.55) \quad G(p + \epsilon) = G(-2 + \epsilon) = \frac{1}{\epsilon^2} \begin{bmatrix} -3 + \epsilon & 4 \\ 4.5 & 2(-3 + \epsilon) \end{bmatrix}$$



# Elements of Linear System Theory

The SVD as  $\epsilon \rightarrow 0$  yields

$$G(-2 + \epsilon) = \frac{1}{\epsilon^2} \begin{bmatrix} -0.55 & -0.83 \\ 0.83 & -0.55 \end{bmatrix} \begin{bmatrix} 9.01 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} 0.6 & -0.8 \\ -0.8 & -0.6 \end{bmatrix}^H$$

$$(3.56) \quad u_p = \begin{bmatrix} 0.60 \\ -0.80 \end{bmatrix} \quad y_p = \begin{bmatrix} -0.55 \\ 0.83 \end{bmatrix}$$

**Note** Locations of poles and zeros are independent of input and output scalings, their directions are *not*.

# Elements of Linear System Theory

## 3.6.2 Remarks on poles and zeros [4.6.2]

1. For square systems the poles and zeros of  $G(s)$  are “essentially” the poles and zeros of  $\det G(s)$ . This fails when zero and pole in different parts of the system cancel when forming  $\det G(s)$ .

$$(3.57) \quad G(s) = \begin{bmatrix} (s+2)/(s+1) & 0 \\ 0 & (s+1)/(s+2) \end{bmatrix}$$

$\det G(s) = 1$ , although the system obviously has poles at  $-1$  and  $-2$  and (multivariable) zeros at  $-1$  and  $-2$ .

2. System (3.57) has poles and zeros at the same locations (at  $-1$  and  $-2$ ). Their directions are different. They do not cancel or otherwise interact.

# Elements of Linear System Theory

3. There are no zeros if the outputs contain direct information about all the states; that is, if from  $y$  we can directly obtain  $x$  (e.g.  $C = I$  and  $D = 0$ );
4. Zeros usually appear when there are fewer inputs or outputs than states
5. **Moving poles.** (a) feedback control ( $G(I + KG)^{-1}$ ) moves the poles, (b) series compensation ( $GK$ , feedforward control) can cancel poles in  $G$  by placing zeros in  $K$  (but not move them), and (c) parallel compensation ( $G + K$ ) cannot affect the poles in  $G$ .

# Elements of Linear System Theory

6. **Moving zeros.** (a) With feedback, the zeros of  $G(I + KG)^{-1}$  are the zeros of  $G$  plus the poles of  $K$ , i.e. the zeros are unaffected by feedback. (b) Series compensation can counter the effect of zeros in  $G$  by placing poles in  $K$  to cancel them, but cancellations are not possible for RHP-zeros due to internal stability (see Section 3.7). (c) The only way to move zeros is by parallel compensation,  $y = (G + K)u$ , which, if  $y$  is a physical output, can only be accomplished by adding an extra input (actuator).

# Elements of Linear System Theory

**Example. Effect of feedback on poles and zeros.**

SISO plant  $G(s) = z(s)/\phi(s)$  and  $K(s) = k$ .

$$(3.58) \quad T(s) = \frac{L(s)}{1 + L(s)} = \frac{kG(s)}{1 + kG(s)} = \frac{kz(s)}{\phi(s) + kz(s)} = k \frac{z_{cl}(s)}{\phi_{cl}(s)}$$

**Note the following:**

1. Zero polynomial:  $z_{cl}(s) = z(s)$

$\Rightarrow$  zero locations are unchanged.

2. Pole locations are changed by feedback.

For example,

$$(3.59) \quad k \rightarrow 0 \quad \Rightarrow \quad \phi_{cl}(s) \rightarrow \phi(s)$$

$$(3.60) \quad k \rightarrow \infty \quad \Rightarrow \quad \phi_{cl}(s) \rightarrow z(s) \cdot \tilde{z}(s)$$

where roots of  $\tilde{z}(s)$  move with  $k$  to infinity (complex pattern)  
(cf. root locus)

# Elements of Linear System Theory

## 3.7 Internal stability of feedback systems [4.7]

**Note:** Checking the pole of  $S$  or  $T$  is not sufficient to determine internal stability

**Example** (Figure 1). In forming  $L = GK$  we cancel the term  $(s - 1)$  (a RHP pole-zero cancellation) to obtain

$$(3.61) \quad L = GK = \frac{k}{s}, \text{ and } S = (I + L)^{-1} = \frac{s}{s + k}$$

$S(s)$  is stable, i.e. transfer function from  $d_y$  to  $y$  is stable.

However, the transfer function from  $d_y$  to  $u$  is unstable:

$$(3.62) \quad u = -K(I + GK)^{-1}d_y = -\frac{k(s + 1)}{(s - 1)(s + k)}d_y$$

# Elements of Linear System Theory

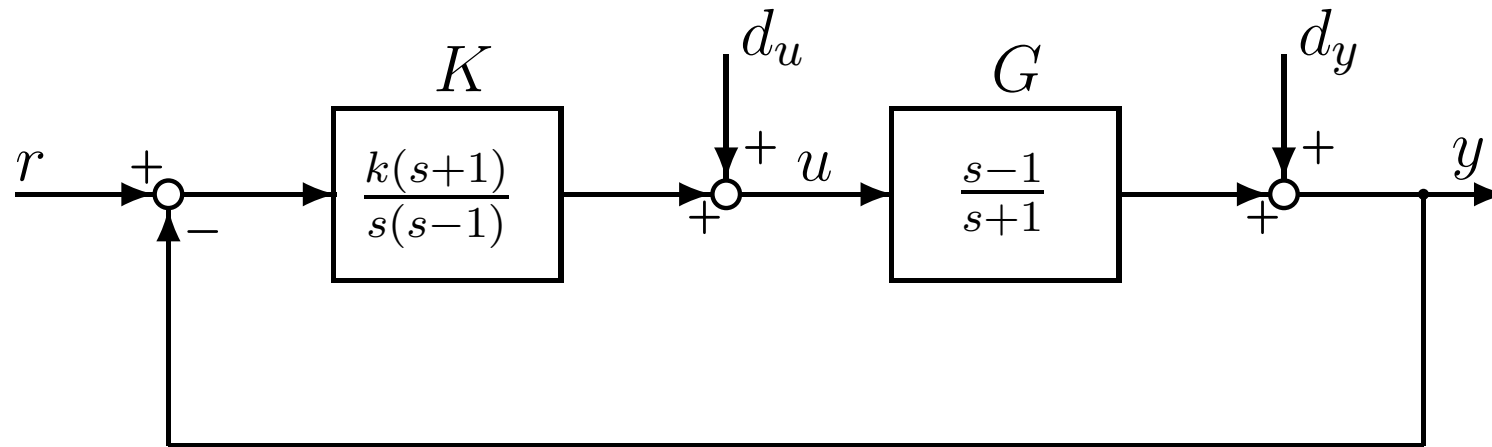


Figure 1: Internally unstable system

# Elements of Linear System Theory

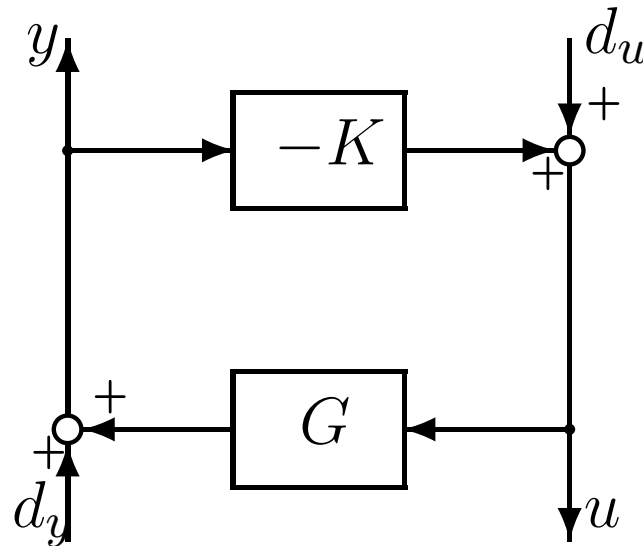


Figure 2: Block diagram used to check internal stability of feedback system

For *internal* stability consider

$$(3.63) \quad u = (I + KG)^{-1}d_u - K(I + GK)^{-1}d_y$$

$$(3.64) \quad y = G(I + KG)^{-1}d_u + (I + GK)^{-1}d_y$$



# Elements of Linear System Theory

**Theorem 4.4** The feedback system in Figure 2 is **internally stable** if and only if all four closed-loop transfer matrices in (3.63) and (3.64) are stable.

**Theorem 4.5** Assume there are no RHP pole-zero cancellations between  $G(s)$  and  $K(s)$ . Then the feedback system in Figure 2 is internally stable if and only if one of the four closed-loop transfer function matrices in (3.63) and (3.64) is stable.

# Elements of Linear System Theory

## Implications of the internal stability requirement

- 1. If  $G(s)$  has a RHP-zero at  $z$ , then  $L = GK$ ,  $T = GK(I + GK)^{-1}$ ,  $SG = (I + GK)^{-1}G$ ,  $L_I = KG$  and  $T_I = KG(I + KG)^{-1}$  will each have a RHP-zero at  $z$ .*
- 2. If  $G(s)$  has a RHP-pole at  $p$ , then  $L = GK$  and  $L_I = KG$  also have a RHP-pole at  $p$ , while  $S = (I + GK)^{-1}$ ,  $KS = K(I + GK)^{-1}$  and  $S_I = (I + KG)^{-1}$  have a RHP-zero at  $p$ .*

# Elements of Linear System Theory

**Exercise: Interpolation constraints.** Prove for SISO feedback systems when the plant  $G(s)$  has a RHP-zero  $z$  or a RHP-pole  $p$ :

$$(3.65) \quad G(z) = 0 \quad \Rightarrow \quad L(z) = 0 \quad \Leftrightarrow \quad T(z) = 0, S(z) = 1$$

$$(3.66) \quad G^{-1}(p) = 0 \quad \Rightarrow \quad L(p) = \infty \quad \Leftrightarrow \quad T(p) = 1, S(p) = 0$$

**Remark** “Perfect control” implies  $S \approx 0$  and  $T \approx 1$ .

RHP-zero  $\Rightarrow$  perfect control impossible.

RHP-pole  $\Rightarrow$  perfect control possible.

RHP-poles cause problems when tight (high gain) control is *not* possible.

# Elements of Linear System Theory

## 3.8 Stabilizing controllers [4.8]

### Stable plants

**Lemma** For a stable plant  $G(s)$  the negative feedback system in Figure 2 is internally stable if and only if

$Q = K(I + GK)^{-1}$  is stable.

*Proof:* The four transfer functions in (3.63) and (3.64) are

$$(3.67) \quad K(I + GK)^{-1} = Q$$

$$(3.68) \quad (I + GK)^{-1} = I - GQ$$

$$(3.69) \quad (I + KG)^{-1} = I - QG$$

$$(3.70) \quad G(I + KG)^{-1} = G(I - QG)$$

which are clearly all stable if and only if  $G$  and  $Q$  are stable.

# Elements of Linear System Theory

**Consequences:** *All stabilizing negative feedback controllers for the stable plant  $G(s)$  are given by*

$$(3.71) \quad K = (I - QG)^{-1}Q = Q(I - GQ)^{-1}$$

where the “parameter”  $Q$  is *any stable transfer function matrix*.  
(Identical to the internal model control (IMC) parameterization of stabilizing controllers.)

# Elements of Linear System Theory

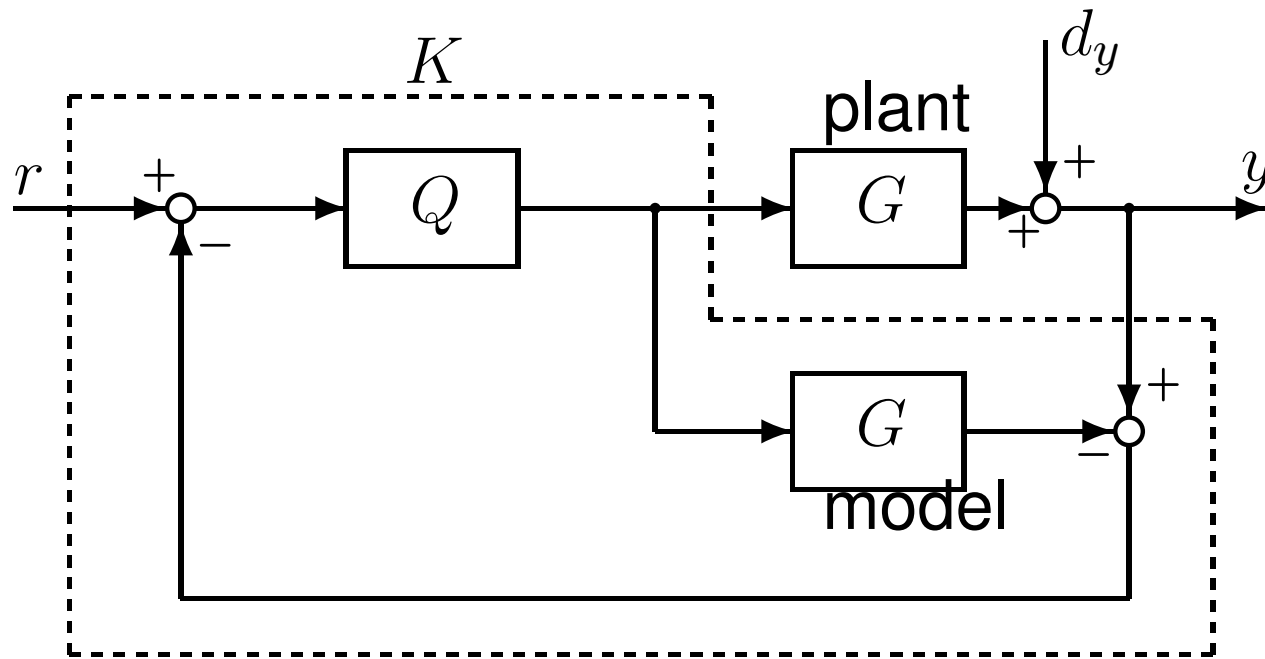


Figure 3: The internal model control (IMC) structure

# Elements of Linear System Theory

## 3.9 Stability analysis in the frequency domain

Generalization of Nyquist's stability test for SISO systems.

### 3.9.1 Open and closed-loop characteristic polynomials

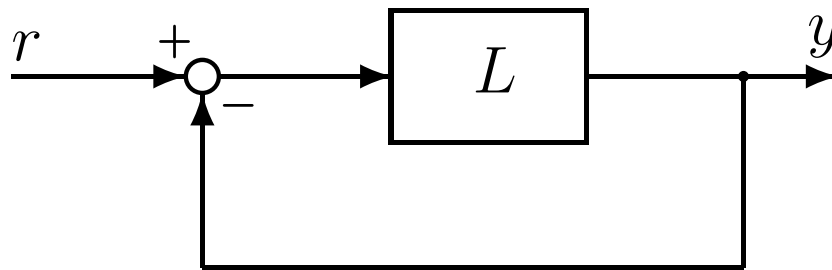


Figure 4: Negative feedback system

# Elements of Linear System Theory

Open Loop:

$$(3.72) \quad L(s) = C_{ol}(sI - A_{ol})^{-1}B_{ol} + D_{ol}$$

Poles of  $L(s)$  are the roots of the *open-loop* characteristic polynomial

$$(3.73) \quad \phi_{ol}(s) = \det(sI - A_{ol})$$

Assume no RHP pole-zero cancellations between  $G(s)$  and  $K(s)$ . Then from Theorem 4.5 internal stability of the *closed-loop* system is equivalent to the stability of  $S(s) = (I + L(s))^{-1}$ .

The realization of  $S(s)$  can be derived as follow:

$$(3.74) \quad \dot{x} = A_{ol}x + B_{ol}(r - y)$$

$$(3.75) \quad -e = r - y = r - C_{ol}x - D_{ol}(r - y)$$



# Elements of Linear System Theory

or

$$(3.76) \quad r - y = (I + D_{ol})^{-1}(r - C_{ol}x)$$

and

$$(3.77) \quad \dot{x} = (A_{ol} - B_{ol}(I + D_{ol})^{-1}C_{ol})x + B_{ol}(I + D_{ol})^{-1}r$$

Therefore the state matrix of  $S(s)$  is:

$$(3.78) \quad A_{cl} = A_{ol} - B_{ol}(I + D_{ol})^{-1}C_{ol}$$

And the closed-loop characteristic polynomial is

$$(3.79) \quad \phi_{cl}(s) \triangleq \det(sI - A_{cl}) = \det(sI - A_{ol} + B_{ol}(I + D_{ol})^{-1}C_{ol})$$

# Elements of Linear System Theory

## Relationship between characteristic polynomials

From (3.72) we get

$$(3.80) \quad \det(I + L(s)) = \det(I + C_{ol}(sI - A_{ol})^{-1}B_{ol} + D_{ol})$$

Schur's formula yields (with

$$A_{11} = I + D_{ol}, A_{12} = -C_{ol}, A_{22} = sI - A_{ol}, A_{21} = B_{ol})$$

$$(3.81) \quad \det(I + L(s)) = \frac{\phi_{cl}(s)}{\phi_{ol}(s)} \cdot c$$

where  $c = \det(I + D_{ol})$  is a constant (cf. SISO result from RSI).

# Elements of Linear System Theory

Side calculation:

$$\begin{aligned} & \det \begin{bmatrix} I + D_{ol} & -C_{ol} \\ B_{ol} & sI - A_{ol} \end{bmatrix} \\ &= \det [I + D_{ol}] \det \left[ sI - A_{ol} + B_{ol} (I + D_{ol})^{-1} C_{ol} \right] \\ &= \det [sI - A_{ol}] \det \left[ I + D_{ol} + C_{ol} (sI - A_{ol})^{-1} B_{ol} \right] \end{aligned}$$

**Schur's formula:**

$$\begin{aligned} \det \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} &= \det(A_{11}) \cdot \det(A_{22} - A_{21} A_{11}^{-1} A_{12}) \\ &= \det(A_{22}) \cdot \det(A_{11} - A_{12} A_{22}^{-1} A_{21}) \end{aligned}$$

# Elements of Linear System Theory

## 3.9.2 MIMO Nyquist stability criteria

**Theorem: Generalized (MIMO) Nyquist theorem.** Let  $P_{ol}$  denote the number of open-loop unstable poles in  $L(s)$ . The closed-loop system with loop transfer function  $L(s)$  and negative feedback is stable if and only if the Nyquist plot of  $\det(I + L(s))$

- i) makes  $P_{ol}$  anti-clockwise encirclements of the origin, and
- ii) does not pass through the origin.

### Note

By “Nyquist plot of  $\det(I + L(s))$ ” we mean “the image of  $\det(I + L(s))$  as  $s$  goes clockwise around the Nyquist  $D$ -contour”.

# Elements of Linear System Theory

## 3.9.4 Small gain theorem [4.9.4]

$$(3.82) \quad \rho(L(j\omega)) \triangleq \max_i |\lambda_i(L(j\omega))|$$

**Theorem: Spectral radius stability condition.** Consider a system with a stable loop transfer function  $L(s)$ . Then the closed-loop system is stable if

$$(3.83) \quad \rho(L(j\omega)) \triangleq \max_i |\lambda_i(L(j\omega))| < 1 \quad \forall \omega$$

# Elements of Linear System Theory

*Proof:* Assume the system is unstable. Therefore  $\det(I + L(s))$  encircles the origin, and there is an eigenvalue,  $\lambda_i(L(j\omega))$  which is larger than 1 at some frequency. If  $\det(I + L(s))$  does encircle the origin, then there must exist a gain  $\epsilon \in (0, 1]$  and a frequency  $\omega'$  such that

$$(3.84) \quad \det(I + \epsilon L(j\omega')) = 0$$

or

$$(3.85) \quad \prod_i \lambda_i(I + \epsilon L(j\omega')) = 0$$

$$(3.86) \quad = 1 + \epsilon \lambda_i(L(j\omega')) = 0 \quad \text{for some } i$$

$$(3.87) \quad = \lambda_i(L(j\omega')) = -\frac{1}{\epsilon} \quad \text{for some } i$$

$$(3.88) \quad \Rightarrow |\lambda_i(L(j\omega'))| \geq 1 \quad \text{for some } i$$

$$(3.89) \quad = \rho(L(j\omega')) \geq 1$$

# Elements of Linear System Theory

**Interpretation:** If the system gain is less than 1 in all directions (all eigenvalues) and for all frequencies ( $\forall \omega$ ), then all signal deviations will eventually die out, and the system is stable.

Spectral radius theorem is conservative because phase information is not considered.

**Small Gain Theorem.** Consider a system with a stable loop transfer function  $L(s)$ . Then the closed-loop system is stable if

$$(3.90) \quad \|L(j\omega)\| < 1 \quad \forall \omega$$

where  $\|L\|$  denotes any matrix norm satisfying  $\|AB\| \leq \|A\| \cdot \|B\|$ , for example the singular value  $\bar{\sigma}(L)$ .

**Note** The small gain theorem is generally more conservative than the spectral radius condition in (3.83).

# Elements of Linear System Theory

## 3.10 System norms [4.10]



Figure 5: System  $G$

Figure 5: System with stable transfer function matrix  $G(s)$  and impulse response matrix  $g(t)$ .

Question: given information about the allowed input signals  $w(t)$ , how large can the outputs  $z(t)$  become?

We use the 2-norm,

$$(3.91) \quad \|z(t)\|_2 = \sqrt{\sum_i \int_{-\infty}^{\infty} |z_i(\tau)|^2 d\tau}$$



# Elements of Linear System Theory

and consider three inputs:

1.  $w(t)$  is a series of unit impulses.
2.  $w(t)$  is any signal satisfying  $\|w(t)\|_2 = 1$ .
3.  $w(t)$  is any signal satisfying  $\|w(t)\|_2 = 1$ , but  $w(t) = 0$  for  $t \geq 0$ , and we only measure  $z(t)$  for  $t \geq 0$ .

The relevant system norms in the three cases are the  $\mathcal{H}_2$ ,  $\mathcal{H}_\infty$ , and Hankel norms, respectively.

# Elements of Linear System Theory

## 3.10.1 $\mathcal{H}_2$ norm [4.10.1]

$G(s)$  strictly proper.

For the  $\mathcal{H}_2$  norm we use the Frobenius norm spatially (for the matrix) and integrate over frequency, i.e.,

$$(3.92) \quad \|G(s)\|_2 \triangleq \sqrt{\frac{1}{2\pi} \int_{-\infty}^{\infty} \underbrace{\text{tr}(G(j\omega)^H G(j\omega))}_{\|G(j\omega)\|_F^2 = \sum_{ij} |G_{ij}(j\omega)|^2} d\omega}$$

$G(s)$  must be strictly proper, otherwise the  $\mathcal{H}_2$  norm is infinite. By Parseval's theorem, (3.92) is equal to the  $\mathcal{H}_2$  norm of the impulse response

$$(3.93) \quad \|G(s)\|_2 = \|g(t)\|_2 \triangleq \sqrt{\int_0^{\infty} \underbrace{\text{tr}(g^T(\tau)g(\tau))}_{\|g(\tau)\|_F^2 = \sum_{ij} |g_{ij}(\tau)|^2} d\tau}$$

# Elements of Linear System Theory

- Note that  $G(s)$  and  $g(t)$  are dynamic *systems* while  $G(j\omega)$  and  $g(\tau)$  are constant *matrices* (for given value of  $\omega$  or  $\tau$ ).
- We can change the order of integration and summation in (3.93) to get

$$(3.94) \quad \|G(s)\|_2 = \|g(t)\|_2 = \sqrt{\sum_{ij} \int_0^{\infty} |g_{ij}(\tau)|^2 d\tau}$$

where  $g_{ij}(t)$  is the  $ij$ 'th element of the impulse response matrix,  $g(t)$ . Thus  $\mathcal{H}_2$  norm can be interpreted as the 2-norm output resulting from applying unit impulses  $\delta_j(t)$  to each input, one after another (allowing the output to settle to zero before applying an impulse to the next input). Thus  $\|G(s)\|_2^2 = \sqrt{\sum_{i=1}^m \|z_i(t)\|_2^2}$  where  $z_i(t)$  is the output vector resulting from applying a unit impulse  $\delta_i(t)$  to the  $i$ 'th input.

# Elements of Linear System Theory

**Numerical computations of the  $\mathcal{H}_2$  norm.**

Consider  $G(s) = C(sI - A)^{-1}B$ . Then

$$(3.95) \quad \|G(s)\|_2 = \sqrt{\text{tr}(B^T Q B)} \quad \text{or} \quad \|G(s)\|_2 = \sqrt{\text{tr}(C P C^T)}$$

where  $Q$  = observability Gramian  
and  $P$  = controllability Gramian

# Elements of Linear System Theory

## 3.10.2 $\mathcal{H}_\infty$ norm [4.10.2]

$G(s)$  proper.

For the  $\mathcal{H}_\infty$  norm we use the singular value (induced 2-norm) spatially (for the matrix) and pick out the peak value as a function of frequency

$$(3.96) \quad \|G(s)\|_\infty \triangleq \max_{\omega} \bar{\sigma}(G(j\omega))$$

The  $\mathcal{H}_\infty$  norm is the peak of the transfer function “magnitude”.

# Elements of Linear System Theory

**Time domain performance interpretations of the  $\mathcal{H}_\infty$  norm.**

- Worst-case steady-state gain for sinusoidal inputs at any frequency.
- Induced (worst-case) 2-norm in the time domain:

$$(3.97) \quad \|G(s)\|_\infty = \max_{w(t) \neq 0} \frac{\|z(t)\|_2}{\|w(t)\|_2} = \max_{\|w(t)\|_2=1} \|z(t)\|_2$$

(In essence, (3.97) arises because the worst input signal  $w(t)$  is a sinusoid with frequency  $\omega^*$  and a direction which gives  $\bar{\sigma}(G(j\omega^*))$  as the maximum gain.)

# Elements of Linear System Theory

Numerical computation of the  $\mathcal{H}_\infty$  norm. Consider

$$G(s) = C(sI - A)^{-1}B + D$$

$\mathcal{H}_\infty$  norm is the smallest value of  $\gamma$  such that the Hamiltonian matrix  $H$  has no eigenvalues on the imaginary axis, where

$$(3.98)H = \begin{bmatrix} A + BR^{-1}D^T C & BR^{-1}B^T \\ -C^T(I + DR^{-1}D^T)C & -(A + BR^{-1}D^T C)^T \end{bmatrix}$$

and  $R = \gamma^2 I - D^T D$

# Elements of Linear System Theory

## 3.10.3 Difference between the $\mathcal{H}_2$ and $\mathcal{H}_\infty$ norms

Frobenius norm in terms of singular values

$$(3.99) \quad \|G(s)\|_2 = \sqrt{\frac{1}{2\pi} \int_{-\infty}^{\infty} \sum_i \sigma_i^2(G(j\omega)) d\omega}$$

Thus when optimizing performance in terms of the different norms:

- $\mathcal{H}_\infty$ : “push down peak of largest singular value”.
- $\mathcal{H}_2$ : “push down whole thing” (all singular values over all frequencies).



# Elements of Linear System Theory

## Example

$$(3.100) \quad G(s) = \frac{1}{s + a}$$

$\mathcal{H}_2$  norm:

$$\begin{aligned} \|G(s)\|_2 &= \left( \frac{1}{2\pi} \int_{-\infty}^{\infty} \underbrace{|G(j\omega)|^2}_{\frac{1}{\omega^2 + a^2}} d\omega \right)^{\frac{1}{2}} \\ &= \left( \frac{1}{2\pi a} \left[ \tan^{-1} \left( \frac{\omega}{a} \right) \right]_{-\infty}^{\infty} \right)^{\frac{1}{2}} = \sqrt{\frac{1}{2a}} \end{aligned}$$

Alternatively: Consider the impulse response

$$(3.101) \quad g(t) = \mathcal{L}^{-1} \left( \frac{1}{s + a} \right) = e^{-at}, t \geq 0$$

# Elements of Linear System Theory

to get

$$(3.102) \quad \|g(t)\|_2 = \sqrt{\int_0^{\infty} (e^{-at})^2 dt} = \sqrt{\frac{1}{2a}}$$

as expected from Parseval's theorem.

$\mathcal{H}_\infty$  norm:

$$(3.103) \quad \|G(s)\|_\infty = \max_{\omega} |G(j\omega)| = \max_{\omega} \frac{1}{(\omega^2 + a^2)^{\frac{1}{2}}} = \frac{1}{a}$$

# Elements of Linear System Theory

## Example

There is no general relationship between the  $\mathcal{H}_2$  and  $\mathcal{H}_\infty$  norms.

$$(3.104) \quad f_1(s) = \frac{1}{\epsilon s + 1}, \quad f_2(s) = \frac{\epsilon s}{s^2 + \epsilon s + 1}$$

$$(3.105) \quad \begin{aligned} \|f_1\|_\infty &= 1 & \|f_1\|_2 &= \infty \\ \|f_2\|_\infty &= 1 & \|f_2\|_2 &= 0 \end{aligned}$$

**Why is the  $\mathcal{H}_\infty$  norm so popular?** In robust control convenient for representing unstructured model uncertainty, and because it satisfies the multiplicative property:

$$(3.106) \quad \|A(s)B(s)\|_\infty \leq \|A(s)\|_\infty \cdot \|B(s)\|_\infty$$

**What is wrong with the  $\mathcal{H}_2$  norm?** It is *not* an induced norm and does *not* satisfy the multiplicative property.

# Elements of Linear System Theory

## Example

Consider again  $G(s) = 1/(s + a)$  in (3.100), for which

$$\|G(s)\|_2 = \sqrt{1/2a}.$$

$$\begin{aligned} \|G(s)G(s)\|_2 &= \sqrt{\int_0^\infty \underbrace{\left| \mathcal{L}^{-1}\left[\left(\frac{1}{s+a}\right)^2\right] \right|^2}_{te^{-at}}} \\ &= \sqrt{\frac{1}{a} \frac{1}{2a}} = \sqrt{\frac{1}{a}} \|G(s)\|_2^2 \end{aligned}$$

(3.107)

# Elements of Linear System Theory

for  $a < 1$ ,

$$(3.108) \quad \|G(s)G(s)\|_2 > \|G(s)\|_2 \cdot \|G(s)\|_2$$

which does not satisfy the multiplicative property.

$\mathcal{H}_\infty$  norm does satisfy the multiplicative property

$$\|G(s)G(s)\|_\infty = \frac{1}{a^2} = \|G(s)\|_\infty \cdot \|G(s)\|_\infty$$