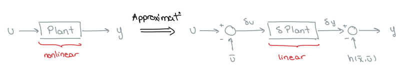
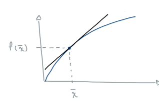
Modelling CS u: control input, y: plant output
State variable CS is in state variable form if
$\dot{x}_1 = f_1(t, x_1, \dots, x_n, u), \dots, \dot{x}_n = f_n(t, x_1,\dots,x_n,u)$
$y = h(t, x_1, \dots, x_n, u)$ is a collection of n 1st order ODEs.
Time-Invariant (TI) CS is TI if $f_i(\cdot)$ does not depend on t.
State space (SS) TI CS is in SS form if $\dot{x}=f(x,u), y=h(x,u)$ where $x(t)\in\mathbb{R}^n$ is called the state.
Single-input-single-output (SISO) CS is SISO if $u(t), y(t)\in\mathbb{R}$.
LTI CS in SS form is LTI if $\dot x = Ax+Bu$, $y=Cx+Du$
$A\in\mathbb{R}^{n\times n},\ B\in \mathbb{R}^{n\times m},\ C\in\mathbb{R}^{p\times n},\ D\in\mathbb{R}^{p\times m}$
where $x(t)\in\mathbb{R}^n,\ u(t)\in\mathbb{R}^m,\ y(t)\in\mathbb{R}^p$.
Input-Output (IO) LTI CS is in IO form if
$\frac{d^ny}{dt^n} + a_{n-1}\frac{d^{n-1}y}{dt^{n-1}} + \cdot\cdot + a_1\frac{dy}{dt} + a_0 y = b_m \frac{d^m u}{dt^m} + \cdot\cdot + b_1\frac{du}{dt} + b_0 u$
where $m\le n$ (causality)

IO to SS Model 1. Define x s.t. highest order derivative in $\dot x$
2. Write $\dot x = Ax+Bu = f(x,u)$ by isolating for components of $\dot x$
3. Write $y=Cx+Du=h(x,u)$ by setting measurement output y to component of x
Equilibria y_d (steady state) b/c if $y(0)=y_d$ at $t=0$, then $y(t)=y_d\ \forall t\ge 0$.
Equilibrium pair Consider the system $\dot x = f(x,u)$. The pair $(\bar x, \bar u)$ is an equilibrium pair if $f(\bar x,\bar u)=0$.
Equilibrium point $\bar x$ is an equilibrium point w/ control $u=\bar u$.
*If $u=\bar u$ and $x(0)=\bar x$ then $x(t)=\bar x\ \forall t\ge0$ (i.e. a system that starts at equilibrium remains at equilibrium).
Find Equilibrium Pair/Point 1. Set $f(x,u)=0$
2. Solve $f(x,u)=0$ to find $(x,u)=(\bar x,\bar u)$.
3. If specific $u=\bar u$, then find $x=\bar x$ by solving $f(x,\bar u)=0$.
Linearization of Nonlinear System Consider system $\dot x = f(x,u)$ w/ equ. pair $(\bar x,\bar u)$, then error coordinates around equ. pair
$\delta x = x-\bar x$, $\delta u = u-\bar u$, $\delta y = y - h(\bar x,\bar u)$ $\dot{\delta x} = \dot x - f(\bar x,\bar u)$ w/
$\dot{\delta x} = A\delta x + B\delta u$, $A = \frac{\partial f(\bar x,\bar u)}{\partial x}\in\mathbb{R}^{n_1\times n_1}$, $B=\frac{\partial f(\bar x,\bar u)}{\partial u}\in\mathbb{R}^{n_1}$,
$\delta y = C\delta x + D\delta u$, $C=\frac{\partial h}{\partial x}(\bar x,\bar u)\in\mathbb{R}^{1\times n_1}$, $D=\frac{\partial h(\bar x,\bar u)}{\partial u}\in\mathbb{R}$
*Only valid at equ. pairs.

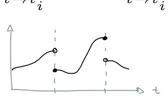


Linear Approx. Given a diff. fcn. $f:\mathbb{R}\to\mathbb{R}$, its linear approx. at $\bar x$ is $f_{\text{lin}} = f(\bar x)+f'(\bar x)(x-\bar x)$.
*Remainder Thm: $f(x)=f_{\text{lin}}+r(x)$ where $\lim_{x\to\bar x}\frac{r(x)}{x-\bar x}=0$.

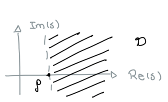


*Note: Can provide a good approx. near $\bar x$ but not globally.
*Gen. $f:\mathbb{R}^{n_1}\to\mathbb{R}^{n_2}$, $f(x)=f(\bar x)+\frac{\partial f}{\partial x}(\bar x)(x-\bar x)+R(x)$
*Jacobian: $\frac{\partial f}{\partial x}(\bar x) = \begin{bmatrix}\frac{\partial f}{\partial x_1}(\bar x) & \cdots & \frac{\partial f}{\partial x_{n_1}}(\bar x)\end{bmatrix}\in\mathbb{R}^{n_2\times n_1}$
Linearization Steps 1. Find equ. pair $(\bar x,\bar u)$
2. Derive A,B,C,D and then evaluate at $(\bar x,\bar u)$
3. Write $\dot{\delta x}=A\delta x+B\delta u$ and $\delta y = C\delta x + D\delta u$

Laplace Transform Given a fcn $f:\mathbb{R}_+=[0,\infty)\to\mathbb{R}^n$, its Laplace transform is $F(s)=\mathcal{L}\{f(t)\}:=\int_{0^-}^\infty f(t)e^{-st}\,dt$, $s\in\mathbb{C}$.
*$\mathcal{L}: f(t)\mapsto F(s)$, $t\in\mathbb{R}_+$ (time dom.) & $s\in\mathbb{C}$ (Laplace dom.).
P.W. CTS: A fcn $f:\mathbb{R}_+\to\mathbb{R}^n$ is **p.w. cts** if on every finite interval of $\mathbb{R}$, $f(t)$ has at most a finite # of discontinuity points (t_i) and the limits $\lim_{t\to t_i^+}f(t)$, $\lim_{t\to t_i^-}f(t)$ are finite.



Exp. Order A function $f:\mathbb{R}_+\to\mathbb{R}^n$ is of **exp. order** if $\exists$ constants $K,\rho,T>0$ s.t. $\|f(t)\|\le Ke^{\rho t}$, $\forall t\ge T$.
Existence of LT Thm If $f(t)$ is p.w. cts and of exp. order w/ constants $K,\rho,T>0$, then $F(\cdot)$ exists and is defined $\forall s\in D := \{s\in\mathbb{C}: \text{Re}(s)>\rho\}$ and $F(\cdot)$ is analytic on D.
*Analytic fcn iff differentiable fcn.
*D: Region of convergence (ROC), open half plane.

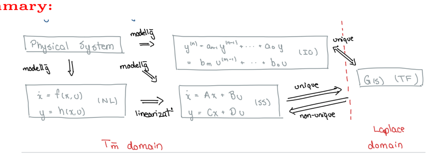


Unit Step $\mathbf{1}(t) := \begin{cases}1, & \text{if } t\ge0\\ 0, & \text{otherwise}\end{cases}$
Table of Common Laplace Transforms: $f(t)$ | $F(s)$
$\mathbf 1(t)\mapsto\frac1s$ $\quad t\mathbf1(t)\mapsto\frac{1}{s^2}$ $\quad t^k\,\mathbf1(t)\mapsto\frac{k!}{s^{k+1}}$ $\quad e^{at}\,\mathbf1(t)\mapsto\frac{1}{s-a}$
$t^ne^{at}\,\mathbf1(t)\mapsto\frac{n!}{(s-a)^{n+1}}$ $\quad \sin(at)\,\mathbf1(t)\mapsto\frac{a}{s^2+a^2}$
$\cos(at)\,\mathbf1(t)\mapsto\frac{s}{s^2+a^2}$ $\quad \frac{1}{2\omega^3}[\sin(\omega t)-\omega t\cos(\omega t)]\,\mathbf1(t)\mapsto\frac{1}{(s^2+\omega^2)^2}$
Prop. of Laplace Transform Linearity:
$\mathcal L\{cf(t)+g(t)\}=c\mathcal L\{f(t)\}+\mathcal L\{g(t)\}$, $c\sim$ constant.
Differentiation: If the Laplace transform of $f'(t)$ exists, then $\mathcal L\{f'(t)\}=s\mathcal L\{f(t)\}-f(0^-)$.
If the Laplace transform of $f^{(n)}(t):=\frac{d^nf}{dt^n}(t)$ exists, then $\mathcal L\{f^{(n)}(t)\}=s^n\mathcal L\{f(t)\}-\sum_{i=1}^n s^{n-i}f^{(i-1)}(0^-)$.
Integration: $\mathcal L\left\{\int_0^t f(\tau)d\tau\right\}=\frac1s\mathcal L\{f(t)\}$.
Convolution: Let $(f*g)(t):=\int_0^t f(\tau)g(t-\tau)d\tau=\int_0^t f(t-\tau)g(\tau)d\tau$, then $\mathcal L\{(f*g)(t)\}=\mathcal L\{f(t)\}\mathcal L\{g(t)\}$.
Time Delay: $\mathcal L\{f(t-T)\mathbf1(t-T)\}=e^{-Ts}\mathcal L\{f(t)\}, T\ge0$.
Multiplication by t: $\mathcal L\{tf(t)\}=-\frac{d}{ds}[\mathcal L\{f(t)\}]$.
Shift in s: $\mathcal L\{e^{at}f(t)\}=\mathcal L\{f(t)\}|_{s\to s-a}=F(s-a)$, where $F(s)=\mathcal L\{f(t)\}$ & a const.

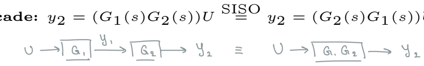
Trig. Id. $2\sin(2t)=2\sin(t)\cos(t)$, $\sin(a-b)=\sin(a)\cos(b)-\cos(a)\sin(b)$, $\cos(a-b)=\cos(a)\cos(b)+\sin(a)\sin(b)$
Complete the Square: $ax^2+bx+c=a(x+\frac{b}{2a})^2-\frac{b^2}{4a}+c$
LT Steps: 1. Write $f(t)$ as a sum and use linearity
*Trig. id. may be useful.
2. Use prop. of LT and common LT to find $F(s)$

Inverse Laplace Transform Given $F(s)$, its inverse LT is $f(t)=\mathcal L^{-1}\{F(s)\}:=\frac{1}{2\pi}\int_{c-j\infty}^{c+j\infty}F(s)e^{st}ds$
$=\lim_{w\to\infty}\frac{1}{2\pi}\int_{c-jw}^{c+jw}F(s)e^{st}ds$, $c\in\mathbb C$ is selected s.t. the line $L:=\{s\in\mathbb C: s=c+j\omega, \omega\in\mathbb R\}$ is inside the ROC of $F(s)$.
Zero: $z\in\mathbb C$ is a zero of $F(s)$ if $F(z)=0$.
Pole: $p\in\mathbb C$ is a pole of $F(s)$ if $\frac{1}{F(p)}=0$.
Cauchy's Residue THM If $F(s)$ is analytic (complex diff.) everywhere except at isolated poles $\{p_1,\dots,p_N\}$, then $\mathcal L^{-1}\{F(s)\}=\sum_{i=1}^N \text{Res}\left[F(s)e^{st}, s=p_i\right]\mathbf1(t)$,
*Res$[F(s)e^{st}, s=p_i]$: Residue of $F(s)e^{st}$ at $s=p_i$.
Residue Computation Let $G(s)$ be a complex analytic fcn w/ a pole at $s=p$, r be the multiplicity of the pole p. Then Res$[G(s), s=p]=\frac{1}{(r-1)!}\lim_{s\to p}\frac{d^{r-1}}{ds^{r-1}}[G(s)(s-p)^r]$.
Inv. LT Partial Frac.: 1. Factorize $F(s)$ into partial fractions.
2. Find coefficients and use LT table to find inverse LT.
*Complete the square.
Inv. LT Residue: 1. Find poles of $F(s)$ and their residues.
2. Use Cauchy's Residue THM to find inverse LT.
*Note: Complex Conjugate (CC) poles → CC residues (use Euler).
Transfer Function: Consider a CS in IO form. Assume zero initial conds. $y(0)=\dots=\frac{d^{(n-1)}y}{dt^{(n-1)}}(0)=0$ and $u(0)=\dots=\frac{d^{(m-1)}u}{dt^{(m-1)}}(0)=0$. Then the TF from u to y is
$G(s):=\frac{y(s)}{U(s)}=\frac{b_ms^m+\dots+b_0}{s^n+a_{n-1}s^{n-1}+\dots+a_0}$.
***0 Ini. Conds.:** $y_0(s)=G(s)u(s)$
***∅ Ini. Conds.:** $y_\emptyset(s)=y_0(s)+\frac{\text{poly. based on initial conds.}}{s^n+a_{n-1}s^{n-1}+\dots+a_0}$

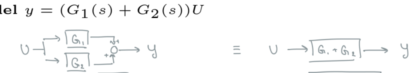
TF Steps (IO to TF): 1. Given IO form of CS, assume zero initial conds.
2. Find $G(s)$ by taking LT of IO form and forming $Y(s)/U(s)$.
*Careful: $Y(s)/U(s)=G(s)$ not $U(s)/Y(s)=G(s)$.
Impulse Response: Given CS modeled by TF $G(s)$, its IR is $g(t):=\mathcal L^{-1}\{G(s)\}$.
*$\mathcal L\{\delta(t)\}=1$, then if $u(t)=\delta(t)$, then $Y(s)=U(s)G(s)=G(s)$.
SS to TF: $G(s)=C(sI-A)^{-1}B+D$ s.t. $y(s)=G(s)U(s)$.
*Assume $x(0)=0\in\mathbb R^n$ (zero initial conds.).
***LTI:** $G(s)$ of an LTI system is always a rational fcn.
***Not Invertible:** Values of s s.t. $sI-A$ not invertible can correspond to poles of $G(s)$.
Inverse: 1. For $A\in\mathbb R^{n\times n}$, find $[\text{cof}(A)]_{(i,j)}=(-1)^{i+j}\det(A_{(i,j)})$.
*$A_{(i,j)}$: A w/ row i and col. j removed.
2. Assemble cof(A) and find $\det(A)=\sum_{j=1}^n a_{ij}[\text{cof}(A)]_{(i,j)}$ w/ fixed i or $\det(A)=\sum_{i=1}^n a_{ij}[\text{cof}(A)]_{(i,j)}$ w/ fixed j.
3. Find $A^{-1}=\frac{1}{\det(A)}\text{adj}(A)=\frac{1}{\det(A)}[\text{cof}(A)]^T$.
*2×2: $A^{-1}=\frac{1}{\det(A)}\begin{bmatrix}d&-b\\-c&a\end{bmatrix}$
TF (SS to TF): 1. Given SS form, assume zero initial conds.
2. Solve $G(s)=C(sI-A)^{-1}B+D$.
*If $C=[0\ \cdot\ 1_i\ \cdot\ 0]$ & $B=[0\ \cdot\ 1_j\ \cdot\ 0]$, then only need ith row & jth col. of adj$(sI-A)$ s.t. $G(s)=\frac{[\text{adj}(sI-A)]_{(i,j)}}{\det(sI-A)}+D$.
*Multiple i,j non-zero entries: Work it out using MM.
TF to SS: Consider $G(s)=\frac{b_ms^m+\dots+b_0}{s^n+a_{n-1}s^{n-1}+\dots+a_0}=\frac{N(s)}{D(s)}$, where $m<n$ (i.e. $G(s)$ is strictly proper). Then the SS form is
*$A=\begin{bmatrix}0&1&0&\cdots&0\\0&0&1&\cdots&0\\ \vdots&\vdots&\vdots&\ddots&\vdots\\0&0&0&0&1\\-a_0&-a_1&-a_2&\cdots&-a_{n-1}\end{bmatrix}$, $B=\begin{bmatrix}0\\ \vdots\\ 0\\1\end{bmatrix}$,
$C=[b_0\ \cdots\ b_m\ |\ 0\ \cdots\ 0]$, $D=0$.
***Unique:** State space of a TF is not unique.
Summary:



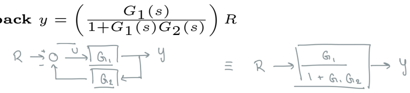
Block Diagram Types of Blocks:
Cascade: $y_2=(G_1(s)G_2(s))U\overset{\text{SISO}}{=}y_2=(G_2(s)G_1(s))U$



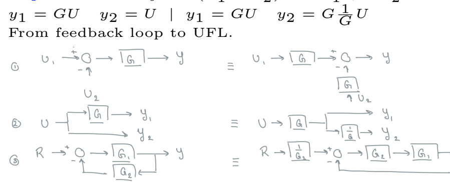
Parallel $y=(G_1(s)+G_2(s))U$



Feedback $y=\left(\frac{G_1(s)}{1+G_1(s)G_2(s)}\right)R$

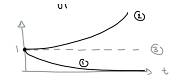


*SC: Unity Feedback Loop (UFL) if $G_2(s)=1$.
Manipulations: 1. $y=G(U_1-U_2)=GU_1+GU_2$
2. $y_1=GU\quad y_2=U$ | $y_1=GU\quad y_2=G\frac1G U$
3. From feedback loop to UFL.

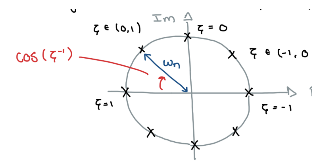


Find TF from Block Diagram: 1. Start from in → out, making simplifications using block diagram rules.
2. Simplify until you get the form $U(s)\to\boxed{G(s)}\to Y(s)$.
Time Response of Elementary Terms: $\mathbf1(t)\leftarrow$ pole @ 0
$t^n\mathbf1(t)\leftarrow$ pole @ 0 w/ mult. n | $e^{at}\mathbf1(t)\leftarrow$ pole @ a
$\sin(\omega t+\phi)\mathbf1(t)\leftarrow$ pole @ $\pm j\omega$ | $\cos(\omega t+\phi)\mathbf1(t)\leftarrow$ pole @ $\pm j\omega$

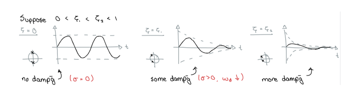
Real Pole: $y(s)=\frac{1}{s+a}$, real pole at $s=-a$, then $y(t)=e^{-at}\mathbf1(t)$
1. $a>0\implies\lim_{t\to\infty}y(t)=0$ | 2. $a<0\implies\lim_{t\to\infty}y(t)=\infty$.
3. $a=0\implies y(t)=\mathbf1(t)$ is constant.



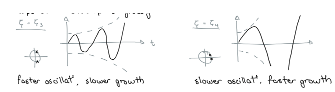
Time Constant: $\tau=\frac1a$ of the pole $s=-a$ for $a>0$
Pair of Comp. Conj. Poles:
$y(s)=\frac{\omega_n^2}{s^2+2\zeta\omega_ns+\omega_n^2}=\frac{\omega_n^2}{(s+\sigma)^2+\omega_d^2}$, $|\zeta|<1$, then
$y(t)=\frac{\omega_n}{\sqrt{1-\zeta^2}}e^{-\sigma t}\sin(\omega_dt)\mathbf1(t)$
*Poles: $s_{1,2}=-\zeta\omega_n\pm j\omega_n\sqrt{1-\zeta^2}=-\sigma\pm j\omega_d$
*$\zeta=\frac{\sigma}{\omega_n}$: Damping ratio (or damping coefficient)
*$\sigma=\zeta\omega_n$: Decay/growth rate | ω_d: Freq. of oscillation
*$\omega_n=\sqrt{\sigma^2+\omega_d^2}\left[\frac{\text{radians}}{\text{seconds}}\right]$: Undamped natural freq.
*$\omega_d=\omega_n\sqrt{1-\zeta^2}\left[\frac{\text{radians}}{\text{seconds}}\right]$: Damped natural freq.
*$|s_{1,2}|^2=\omega_n^2$: Mag. of poles is ω_n.
*$\cos^{-1}(\zeta)$: Angle of s_1 on complex plane CW from -ve Re axis.



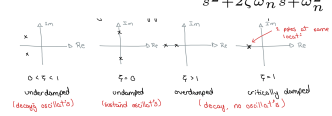
Damping Ratio Effect: $0<\zeta_1<\zeta_2<1$, then



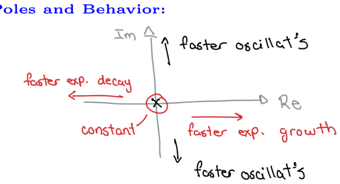
$-1<\zeta_4<\zeta_3<0$, then $\sigma=\zeta\omega_n<0$, (exp. envelop ↑)



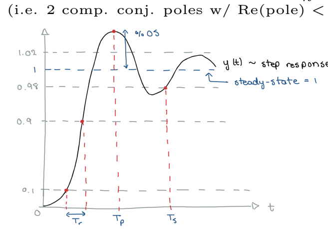
Class. of 2nd Order Sys.: $y(s)=\frac{\omega_n^2}{s^2+2\zeta\omega_ns+\omega_n^2}$, w/ $|\zeta|<1$



Loc. of Poles and Behavior:



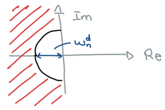
Control Spec. of 2nd Order Sys.: Step Response: Given a TF $G(s)$, its SR is $y(t)$ resulting from applying the input $u(t)=\mathbf1(t)$, i.e. $\mathcal L^{-1}\left\{G(s)\frac1s\right\}$.
Control Spec. A control spec. is a criterion specifiying how we would like a CS to behave.
2nd Order Sys. Metrics: $G(s)=\frac{\omega_n^2}{s^2+2\zeta\omega_ns+\omega_n^2}$ w/ $U(s)=\frac1s$
*$0<\zeta<1$ (i.e. 2 comp. conj. poles w/ Re(pole) < 0).



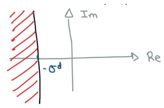
Rise Time (RT): T_r is the time it takes $y(t)$ to go from 10% to 90% of its steady-state value.
RT: 1. Find $t_1>0$ s.t. $y(t_1)=0.1$, $t_2>0$ s.t. $y(t_2)=0.9$.
3. Compute $T_r=t_2-t_1$. $\boxed{T_r\approx\frac{1.8}{\omega_n}}$.
Settling Time (ST): T_s is the time required to reach and stay w/in 2% of the steady-state value.
ST: 1. Find when it's first that $|y(t)-1|\le0.02$. $\boxed{T_s\approx\frac{4}{\zeta\omega_n}}$.
Peak Time: T_p is time req'd to reach the max (peak) value.
Peak Time: 1. Find the first time when $\dot y(t)=0$.
*$\boxed{T_p=\frac{\pi}{\omega_d}=\frac{\pi}{\omega_n\sqrt{1-\zeta^2}}}$
% Overshoot: %OS $=\frac{[\text{peak value}]-[\text{steady-state value}]}{[\text{steady-state value}]}\times100\%$
*% OS = OS × 100%.
*$\boxed{\%\text{OS}=\exp\left(-\frac{\pi\zeta}{\sqrt{1-\zeta^2}}\right)\iff\zeta=\frac{-\ln(\text{OS})}{\sqrt{\pi^2+(\ln(\text{OS}))^2}}}$.

Transient Performance Sat.: Given performance spec. $T_r \le T_r^d$, $T_s \le T_s^d$, $\text{OS} \le \text{OS}^d$, find loc. of poles of $G(s)$.
*Admissible region for the poles of $G(s)$ s.t. the step response meets all three spec. is the intersection of the above three regions.

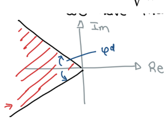
Rise Time: $T_r \approx \frac{1.8}{\omega_n} \le T_r^d \overset{\text{app.}}{\iff} \omega_n \ge \frac{1.8}{T_r^d} \equiv \omega_n^d$



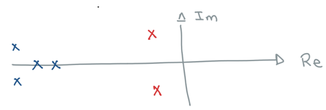
Settling Time: $T_s \approx \frac{4}{\zeta\omega_n} = \frac{4}{\sigma} \le T_s^d \overset{\text{app.}}{\iff} \sigma \ge \frac{4}{T_s^d} \equiv \sigma^d$



OS: $\exp\left(\frac{-\pi\zeta}{\sqrt{1\text{-}\zeta^2}}\right) \le \text{OS}^d \overset{\text{app.}}{\iff} \zeta \ge \frac{-\ln(\text{OS}^d)}{\sqrt{\pi^2+(\ln(\text{OS}^d))^2}} \equiv \zeta^d$



Add. Poles & Zeros: The analysis remains approx. correct under the following assumptions:
1. Any add. poles of $G(s)$ have much more -ve real part (5-10 times) than the real part of the dom. complex conjugate poles.



*dominant poles, additional poles.
2. Real part of zeros are -ve & very diff. from the real part of the two dom. poles.

Internal Stablity: $\dot{x} = Ax$ is
1. **Stable** if $\forall x(0) \in \mathbb{R}^n$, the soln. $x(t)$ is bdd; that is, $\exists M > 0$ s.t. $\|x(t)\| \le M \ \forall t \ge 0$.
2. **Asymp. Stable** if it's stable & $\forall x(0) \in \mathbb{R}^n$, the soln. $x(t)$ converges to the origin; that is, $\lim_{t\to\infty} x(t) = 0$.
3. **Unstable** if it's not stable; that is, $\exists x(0) \in \mathbb{R}^n$ s.t. $x(t)$ is not bdd.

Asymptotic Stablity Thm. $\dot{x} = Ax$ is A.S. iff $\text{eig}(A) \subseteq \mathbb{C}^- \equiv \{s \in \mathbb{C} \mid \text{Re}(s) < 0\}$, i.e. open left half plane (OLHP).
Instability Thm. If $\exists$ an eigenvalue λ of A w/ $\text{Re}(\lambda) > 0$, then $\dot{x} = Ax$ is unstable.
Fact: Zeros of $s^2 + a_1 s + a_0$ are in $\mathbb{C}^-$ iff $a_1, a_0 > 0$.
Internal Stability 1. Linearize around $(\bar{x}, \bar{u})$ w/ $\bar{u} = 0$.
2. Find A and determine $\text{eig}(A) = \lambda$ s.t. $\det(sI - A) = 0$.
3. Check if $\text{eig}(A) \subseteq \mathbb{C}^- \ \mid \ \text{Re}(\text{eig}(A)) > 0$.
BIBO Stability: An LTI system w/ 0 i.c. is Bounded Input Bounded Output (BIBO) stable if for any bdd input $u(t)$, the output $y(t)$ is also bdd.
BIBO Unstable: An LTI system w/ 0 i.c. is BIBO unstable if it's not BIBO stable; that is, $\exists$ a bdd $u(t)$ s.t. $y(t)$ is not bdd.
BIBO Stable Thm. A system $y(s) = G(s)U(s)$ is BIBO stable iff $\text{poles}(G(s)) \subseteq \mathbb{C}^-$.
Lemma: If p is a pole of $G(s)$, then p is an $\text{eig}(A)$. I.e. $\text{poles}(G(s)) := \{p \in \mathbb{C} \mid p \text{ is a pole of } G(s)\} \subseteq \text{eig}(A)$.
***Pole-0 Cancellation:** $\text{eige}(A)$ need not be a pole of $G(s)$.
Thm. If $\text{eig}(A) \subseteq \mathbb{C}^-$, then $\forall B, C, D$ the TF $G(s)$ is BIBO stable. That is, internal asymptotic stability $\Rightarrow$ BIBO stability.
BIBO Stability 1. Find $G(s)$ from SS form and determine poles.
2. Check if $\text{poles}(G(s)) \subseteq \mathbb{C}^-$.

Routh-Hurwitz: Consider $a(s) = s^n + a_{n-1}s^{n-1} + \cdots + a_0$.

*s^n	1	a_{n-2}	a_{n-4}	a_{n-6}	$\cdots$	0
*s^{n-1}	a_{n-1}	a_{n-3}	a_{n-5}	a_{n-7}	$\cdots$	0
*s^{n-2}	b_1	b_2	b_3	$\cdots$		
*s^{n-3}	c_1	c_2	$\cdots$			
$\vdots$						
*s	$*$	0				
*1	$*$	0				

$b_1 = -\frac{1}{a_{n-1}} \det\begin{bmatrix} 1 & a_{n-2} \\ a_{n-1} & a_{n-3} \end{bmatrix}$ $b_2 = -\frac{1}{a_{n-1}} \det\begin{bmatrix} 1 & a_{n-4} \\ a_{n-1} & a_{n-5} \end{bmatrix}$

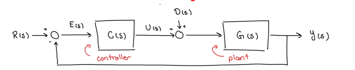
$b_3 = -\frac{1}{a_{n-1}} \det\begin{bmatrix} 1 & a_{n-6} \\ a_{n-1} & a_{n-7} \end{bmatrix}$ $c_1 = -\frac{1}{b_1} \det\begin{bmatrix} a_{n-1} & a_{n-3} \\ b_1 & b_2 \end{bmatrix}$

$c_2 = -\frac{1}{b_1} \det\begin{bmatrix} a_{n-1} & a_{n-5} \\ b_1 & b_3 \end{bmatrix}$

Routh-Hurwitz Stability Criterion: The roots of $a(s)$ are in $\mathbb{C}^-$ iff the 1st col of Routh array has no sign changes. The # of sign changes is equal to the # of roots of $a(s) \in \mathbb{C}^+ := \{s \in \mathbb{C} : \text{Re}(s) > 0\}$.
*If 1st element of a row is 0, Rooth array cannot be completed.
FVT v1: Suppose $Y(s) = \mathcal{L}\{y(t)\}$ is a proper rational fcn. If $y(\infty) := \lim_{t\to\infty} y(t)$ exists and is finite, then $y(\infty) = \lim_{s\to 0} sY(s)$.
FVT v2: Suppose $Y(s) = \mathcal{L}\{y(t)\}$ is a proper rational fcn. Moreover, suppose either:
1. $\text{poles}(Y(s)) \subseteq \mathbb{C}^-$
2. $Y(s)$ has only one pole at $s = 0$ and all other poles are in $\mathbb{C}^-$.
Then $y(\infty) := \lim_{t\to\infty} y(t)$ exists and is finite and satisfies $y(\infty) = \lim_{s\to 0} sY(s)$.
FVT 1. Does $y(\infty)$ exist? Check if pole at $s = 0$, then compute Rooth Array to see if poles are in $\mathbb{C}^-$.
2. Compute $\lim_{s\to 0} sY(s)$ if it exists.

MIDTERM CUTOFF

Standard Feedback Control Loop



$R(s)$: Ref., $E(s) = R(s) - y(s)$: Err., $C(s)$: Controller, $U(s)$: Control input, $D(s)$: Dist., $G(s)$: Plant, $y(s)$: Plant output.
***Assume:** $R(s)$ and $D(s)$ are strictly proper rational fcns w/ a fixed set of poles but arbitrary zeros & gain.
*$\mathcal{R}, \mathcal{D}$: Classes of ref. and dist. satisfying the above assumption.
Basic Control Prob.: Design $C(s)$ s.t. 3 spec. are met:
1. **Stability:** $\forall$ bdd $r(t), d(t)$, we have $u(t), e(t)$ bdd.
2. **Asymptotic Tracking:** When $d(t) = 0 \ \forall t \ge 0$, then $\forall r(t) \in \mathcal{R}$, $\lim_{t\to\infty} e(t) = \lim_{t\to\infty} r(t) - y(t) = 0$.
3. **Disturbance Rejection:** When $r(t) = 0 \ \forall t \ge 0$, then $\forall d(t) \in \mathcal{D}$, $\lim_{t\to\infty} y(t) = 0$.
Open-Loop Control: 1. Design $u(t)$ s.t. $y(t)$ tracks ref. $y_r \in \mathbb{R}$, i.e. $\lim_{t\to\infty} y(t) = y_r$.
2. Set $u(t) = \gamma y_r \mathbf{1}(t)$ w/ $\gamma \in \mathbb{R}$ (const. scaling factor)
3. Apply FVT to find γ s.t. $\lim_{t\to\infty} y(t) = y_r$. 4. Determine $\lim_{t\to\infty} e(t) = \lim_{t\to\infty} y_r - y(t)$
Limitations: 1. Req. perfect knowledge of plant paramters.
2. Not robust against parameter var./(unknown) dist.
3. Does not allow us to speed up convergence.
Feedback Control: 1. Design $u(t)$ s.t. $y(t)$ tracks ref. $y_r \in \mathbb{R}$, i.e. $\lim_{t\to\infty} y(t) = y_r$.
2. Set $u(t) = Ke(t) = K(y_r - y(t))$ w/ $K > 0$ (const. gain).
3. Use block mani. to find $y(s)$ in terms of input and $G(s)$.
4. Apply FVT to find K s.t. $\lim_{t\to\infty} y(t) = y_r$.
5. Determine $\lim_{t\to\infty} e(t) = \lim_{t\to\infty} y_r - y(t)$
Advantages: 1. Doesn't req. perfect knowledge of plant param.
2. Robust against param. var./dist. by $\uparrow K$.
3. Allows us to speed up the rate of convergence by $\uparrow K$.
Disadvantages: 1. Feedback can introduce instability.
2. High-gain amplifies noise.
3. Asymptotic tracking doesn't occur.
Integral Control: 1. Design $u(t)$ s.t. $y(t)$ tracks ref. $y_r \in \mathbb{R}$, i.e. $\lim_{t\to\infty} y(t) = y_r$.
2. Set $u(t) = \mathcal{L}^{-1}\{C(s)E(s)\} = Ke(t) + KT_I \int_0^t e(\tau)d\tau$ (prop. int. (PI) controller) w/ $K, T_I > 0$ (const. gains).
*$C(s) = K\left(1 + \frac{T_I}{s}\right)$
3. Use block mani. to find $y(s)$ in terms of input and $G(s)$.
4. Apply FVT to find $\lim_{t\to\infty} y(t) = y_r$ as desired.
BIBO Stability of Closed-Loop System: Gang of 4 TF:

$$\begin{bmatrix} E(s) \\ U(s) \end{bmatrix} = \begin{bmatrix} \frac{1}{1+C(s)G(s)} & \frac{-G(s)}{1+C(s)G(s)} \\ \frac{C(s)}{1+C(s)G(s)} & \frac{-C(s)G(s)}{1+C(s)G(s)} \end{bmatrix} \begin{bmatrix} R(s) \\ D(s) \end{bmatrix}$$

BIBO Stable of CLS: The std. feedback control loop (CLS) is BIBO stable if all the Gang of 4 TFs are BIBO stable.
CLS is BIBO Stable THM: The CLS is BIBO stable iff
1. Poles of $\frac{1}{1+C(s)G(s)} \subseteq \mathbb{C}^-$
2. $C(s)G(s)$ has no pole-zero cancel. in $\bar{\mathbb{C}}^+ = \{s \in \mathbb{C} : \text{Re}(s) \ge 0\}$.
Practical Considerations:
1. Don't cancel an unstable 0 of $G(s)$ w/ an unstable pole in $C(s)$.
2. Don't cancel an unstable pole of $G(s)$ w/ an unstable 0 in $C(s)$.
Asymp. Tracking of Poly. Suppose $d(t) = 0$ & want to track a <u>poly. ref. signal</u> of the form: $r(t) = \sum_{i=0}^{k-1} c_i t^i 1(t)$, that is: $R(s) = \frac{N_R(s)}{s^k}$, w/ $N_R(0) \ne 0$ and $\deg(N_R(s)) \le k - 1$.
***GOAL:** Design $C(s)$ to achieve $\lim_{t\to\infty} e(t) = 0$.
Prop: Suppose $C(s)$ is designed so that:
1. $\frac{1}{1+C(s)G(s)}$ is BIBO stable
2. $C(s)G(s) = \frac{C'(s)G'(s)}{s^k}$ with $C'(0)G'(0) \ne 0$.
Then $\frac{1}{s^k + C'(s)G'(s)}$ is BIBO stable.
Asymp. Tracking of Poly. Thm Suppose $C(s)$ satisfies CLS is BIBO stable THM and $d(t) = 0 \ \forall t \ge 0$. For any poly. ref. signal $r(t) = \sum_{i=0}^{k-1} c_i t^i 1(t)$, the following hold:
a. If $C(s)G(s)$ has k or more poles at $s = 0$, then $\lim_{t\to\infty} e(t) = 0$.
b. If $C(s)G(s)$ has $k - 1$ poles at $s = 0$, then:

$$\lim_{t\to\infty} e(t) = \begin{cases} \frac{N_R(0)}{1+C'(0)G'(0)}, & \text{if } k = 1 \\ \frac{N_R(0)}{C'(0)G'(0)}, & \text{if } k \ge 2 \end{cases}$$

c. If $C(s)G(s)$ has k-2 or fewer poles at $s = 0$, then $\lim_{t\to\infty} |e(t)| = \infty$.
Type k: The TF $C(s)G(s)$ is of type k if it has k poles at $s = 0$.
Dist. Rejection: Suppose $r(t) = 0 \ \forall t \ge 0$ and $d(t)$ is a poly. dist. signal of the form: $d(t) = \sum_{i=0}^{k-1} c_i t^i 1(t)$, that is: $D(s) = \frac{N_D(s)}{s^k}$, with $N_D(0) \ne 0$ and $\deg(N_D(s)) \le k - 1$.
***GOAL:** Design $C(s)$ to achieve $\lim_{t\to\infty} e(t) = 0$.
Dist. Rejection Thm: Suppose $C(s)$ satisfies CLS is BIBO stable THM and $r(t) = 0 \ \forall t \ge 0$. For any poly. dist. signal $d(t) = \sum_{i=0}^{k-1} c_i t^i 1(t)$, the following hold:
a. If $C(s)$ has k or more poles at $s = 0$, then $\lim_{t\to\infty} e(t) = 0$.
b. If $C(s)$ has $k - 1$ poles at $s = 0$, then $\lim_{t\to\infty} e(t) \ne 0$ exists.
c. If $C(s)$ has k-2 or fewer poles at $s = 0$, then $\lim_{t\to\infty} |e(t)| = \infty$.
General Thm (Internal Model Principle): Suppose $R(s)$ and $D(s)$ are strictly proper rational fns w/ poles in $\overline{\mathbb{C}^+}$. $C(s)$ solves the Basic Control Problem iff:
1) $C(s)$ makes the CLS BIBO stable;
2) $C(s)G(s)$ has the poles$(R(s))$ w/ at least same multiplicities;
3) $C(s)$ has the poles$(D(s))$ w/ at least same multiplicities.
Corollary: If $G(s)$ has zeros that are also poles of $R(s)$ or $D(s)$, then the Basic Control Problem is unsolvable.
Internal Model: The IMP states if $G(s)$ does not contain the poles of $R(s)$ and $D(s)$, then $C(s)$ must contain these poles. Since these poles enable $C(s)$ to reproduce $r(t)$ and $d(t)$, we say $C(s)$ must contain an <u>internal model</u> of $r(t)$ and $d(t)$.
Proposition: Suppose $G(s)$ is BIBO stable. Let $Y(s) = G(s)U(s)$, where $Y(s) = \mathcal{L}\{y(t)\}$ and $U(s) = \mathcal{L}\{u(t)\}$. If $\lim_{t\to\infty} u(t) = 0$, then $\lim_{t\to\infty} y(t) = 0$.
*Decaying input $\implies$ decaying output so don't worry in IMP.
General Controller Design Procedure: Given $R(s) = \mathcal{L}\{r(t)\}$ and $D(s) = \mathcal{L}\{d(t)\}$:
1. **Feasibility:** Verify no zero of $G(s)$ is an unstable pole of $R(s)$ or $D(s)$.
2. **Internal Model:** Let $p_1, \ldots, p_k$ denote the unstable poles of $R(s)$ or $D(s)$ not in $G(s)$, accounting for multiplicities. Construct:

$$C(s) = C'(s) \cdot \frac{1}{(s - p_1)\ldots(s - p_k)}$$

3. **Stability:** Design $C'(s)$ so that the CLS is BIBO stable.
4. **Performance:** Tune controller parameters to achieve the desired performance specifications.