3.1: The $z$ transform

- Computer control requires analog-to-digital (A2D) and digital-to-analog (D2A) conversion.
- Everything shown in the dashed box below represents the internals of a computer or embedded system in a digital control scenario.

- The A2D comprises a sampler and quantizer; the D2A performs a zero-order-hold operation.
- The controller is designed in discrete time, where the $z$ transform is used as a design tool rather than the Laplace transform.

**DEFINITION:** The $z$ transform is defined for a signal $x(kT)$ to be

$$X(z) = x(0) + x(T)z^{-1} + x(2T)z^{-2} + x(3T)z^{-3} + \cdots$$

$$= \sum_{k=0}^{\infty} x(kT)z^{-k}$$

$$\left(\text{or, } X(z) = \sum_{k=0}^{\infty} x[k]z^{-k}.\right)$$
As with the Laplace-transform variable “$s$,” the $z$-transform variable “$z$” is a complex number.

I assume that you have studied the $z$ transform before, and so include only a short review in these notes.

**EXAMPLE**: Define the digital impulse (pulse) function to be

$$
\delta[k] = \begin{cases} 
1, & k = 0; \\
0, & \text{otherwise.} 
\end{cases}
$$

Note: This is very different from an analog impulse (e.g., $\delta[0]$ is defined), but plays a similar role.

so, let $x[k] = \delta[k]$

$$
X(z) = \sum_{k=0}^{\infty} x[k]z^{-k} = x[0] + x[1]z^{-1} + x[2]z^{-2} + \cdots = x[0] = 1.
$$

This sum converges regardless of the value of $z$, so ROC $= \{|z| > 0\}$.

**Selected important properties of the $z$ transform**

Some select properties of the $z$ transform may be useful.

**LINEARITY**: The $z$ transform is linear.

- If $x[k] \leftrightarrow X(z)$ and $v[k] \leftrightarrow V(z)$, then

  $$
  ax[k] + bv[k] \leftrightarrow aX(z) + bV(z).
  $$
RIGHT SHIFT IN TIME (DELAY): When shifting a signal in time,

\[ x[k - 1] \iff z^{-1}X(z) + x[-1] \]
\[ x[k - 2] \iff z^{-2}X(z) + x[-2] + z^{-1}x[-1] \]
\[ \vdots \]
\[ x[k - q] \iff z^{-q}X(z) + x[-q] + z^{-1}x[-q + 1] + \cdots + z^{-q+1}x[-1]. \]

CONVOLUTION: Discrete-time convolution is analogous to continuous-time convolution:

\[ (x * v)[k] = \sum_{i=0}^{k} x[i]v[k - i], \quad k \geq 0 \]

or,

\[ = \sum_{i=0}^{k} v[i]x[k - i], \quad k \geq 0. \]

- Then, assuming that \( x[k] \) and \( v[k] \) are zero for negative \( k \),

\[ (x * v)[k] \iff X(z)V(z). \]

INITIAL-VALUE THEOREM: If \( x[k] \iff X(z), x[0] = \lim_{z \to 1} X(z). \)

FINAL-VALUE THEOREM: If a finite final value exists (all poles of \( X(z) \) strictly inside the unit circle except perhaps for a single pole on the unit circle at \( z = 1 \)),

\[ \lim_{k \to \infty} x[k] = \lim_{z \to 1} (z - 1)X(z). \]

- The table on the next page lists \( z \) transforms for commonly encountered sampled time-domain signals:
**Example table showing $z$ transform of sampled time-domain signals:**

<table>
<thead>
<tr>
<th>Time function</th>
<th>Laplace transform</th>
<th>$z$-Transform</th>
</tr>
</thead>
<tbody>
<tr>
<td>$e(t)$</td>
<td>$E(s)$</td>
<td>$E(z)$</td>
</tr>
<tr>
<td>$1(t)$</td>
<td>$\frac{1}{s}$</td>
<td>$\frac{z}{z-1}$</td>
</tr>
<tr>
<td>$t1(t)$</td>
<td>$\frac{1}{s^2}$</td>
<td>$\frac{Tz}{(z-1)^2}$</td>
</tr>
<tr>
<td>$\frac{t^2}{2}1(t)$</td>
<td>$\frac{1}{s^3}$</td>
<td>$\frac{T^2z(z+1)}{2(z-1)^3}$</td>
</tr>
<tr>
<td>$t^{k-1}1(t)$</td>
<td>$\frac{(k-1)!}{s^k}$</td>
<td>$\lim_{a\to0}(-1)^{k-1}\frac{\partial^k}{\partial a^k} \left[ \frac{z}{z-e^{-at}} \right]$</td>
</tr>
<tr>
<td>$e^{-at}1(t)$</td>
<td>$\frac{1}{s+a}$</td>
<td>$\frac{z}{z-e^{-at}}$</td>
</tr>
<tr>
<td>$te^{-at}1(t)$</td>
<td>$\frac{1}{(s+a)^2}$</td>
<td>$\frac{Tze^{-aT}}{(z-e^{-at})^2}$</td>
</tr>
<tr>
<td>$t^ke^{-at}1(t)$</td>
<td>$\frac{(k-1)!}{(s+a)^k}$</td>
<td>$(-1)^k\frac{\partial^k}{\partial a^k} \left[ \frac{z}{z-e^{-at}} \right]$</td>
</tr>
<tr>
<td>$1-e^{-at}$</td>
<td>$\frac{a}{s(s+a)}$</td>
<td>$\frac{z(1-e^{-at})}{(z-1)(z-e^{-at})}$</td>
</tr>
<tr>
<td>$t = \frac{1-e^{-aT}}{a}$</td>
<td>$\frac{a^2}{s^2(s+a)}$</td>
<td>$\frac{z}{z-1} - \frac{z}{z-e^{-at}} - \frac{aTe^{-aT}z}{z-e^{-at})^2}$</td>
</tr>
<tr>
<td>$1-(1+at)e^{-at}$</td>
<td>$\frac{b}{(s+a)(s+b)}$</td>
<td>$\frac{(e^{-aT}-e^{-bT})z}{(z-e^{-at})^2}$</td>
</tr>
<tr>
<td>$e^{-aT} - e^{-bT}$</td>
<td>$\frac{a}{s^2+a^2}$</td>
<td>$\frac{z\sin(at)}{z^2-2z\cos(at)+1}$</td>
</tr>
<tr>
<td>$\sin(at)$</td>
<td>$\frac{s}{s^2+a^2}$</td>
<td>$\frac{z(z\cos(at))}{z^2-2z\cos(at)+1}$</td>
</tr>
<tr>
<td>$\cos(at)$</td>
<td>$\frac{b}{(s+a)^2+b^2}$</td>
<td>$\frac{e^{-at}\sin(bt)}{z^2-2e^{-at}\cos(bt)+e^{-2at}}$</td>
</tr>
<tr>
<td>$e^{-at}\sin(bt)$</td>
<td>$\frac{s-a}{(s+a)^2+b^2}$</td>
<td>$\frac{z^2-2e^{-at}\cos(bt)+e^{-2at}}{z^2-2e^{-at}\cos(bt)+e^{-2at}}$</td>
</tr>
<tr>
<td>$e^{-at}\cos(bt)$</td>
<td>$\frac{s+a}{(s+a)^2+b^2}$</td>
<td>$\frac{z^2-2e^{-at}\cos(bt)+e^{-2at}}{z^2-2e^{-at}\cos(bt)+e^{-2at}}$</td>
</tr>
<tr>
<td>$1-e^{-at}(\cos(bt) + \frac{a}{b}\sin(bt))$</td>
<td>$\frac{a^2+b^2}{s[(s+a)^2+b^2]}$</td>
<td>$\frac{z(Az+B)}{(z-1)(z^2-2e^{-at}\cos(bt)+e^{-2at})}$</td>
</tr>
</tbody>
</table>

\[ A = 1 - e^{-aT} \left( \cos(bt) + \frac{a}{b}\sin(bt) \right) \]
\[ B = e^{-2aT} + e^{-aT} \left( \frac{a}{b}\sin(bt) - \cos(bt) \right) \]
3.2: Working with the $z$ transform

- The following shows some correspondence between the $z$-plane and some discrete-time unit-pulse-response signals.

**Unit step:**
- $h[k] = 1[k]$.
- $H(z) = \frac{z}{z-1}, \quad |z| > 1$.

**Exponential (geometric):**
- $h[k] = a^k 1[k], \quad |a| < 1$.
- $H(z) = \frac{z}{z-a}, \quad |z| > |a|$.

**General cosinusoid:**
- $h[k] = a^k \cos[\omega k] 1[k], \quad |a| < 1$.
- $H(z) = \frac{z(z - a \cos \omega)}{z^2 - 2a \cos \omega z + a^2}$
  for $|z| > |a|$.

- The radius to the two poles is $a$; the angle to the poles is $\omega$.
- The zero (not at the origin) has the same real part as the two poles.
  - If $\omega = 0$, $H(z) = \frac{z}{z-a} \ldots$ geometric!
  - If $\omega = 0$, $a = 1$, $H(z) = \frac{z}{z-1} \ldots$ step!
- Pole radius $a$ is the geometric factor, determines settling time.
  1. $|a| = 0$, finite-duration response. *e.g.*, $\delta[k - N] \iff z^{-N}$.
  2. $|a| > 1$, growing signal that will not decay.
  3. $|a| = 1$, signal with constant amplitude; either step or cosine.
  4. $|a| < 1$, decaying signal. Small $a = \text{fast decay (see below)}$. 
Pole angle $\omega$ determines number of samples per oscillation.

- That is, if we require $\cos(\omega k) = \cos(\omega (k + N))$, then

$$N = \frac{2\pi}{\omega} \text{ rad} = \frac{360}{\omega} \text{ deg}.$$  

- Solid: cst. damping ratio $\zeta$.
- Dashed: constant natural frequency $\omega_n$.

Plot to right shows discrete-time unit-pulse responses versus pole locations.
Correspondence with continuous-time signals

- Let \( g(t) = e^{-at} \cos(bt)1(t) \).
- Suppose
  \[
  \begin{align*}
  a &= \frac{0.3567}{T} \\
  b &= \frac{\pi/4}{T} 
  \end{align*}
  \]
  \( T \) = sampling period.
- Then,
  \[
  g[k] = g(kT) = (e^{-0.3567})^k \cos\left(\frac{\pi k}{4}\right)1[k] \\
  = 0.7^k \cos\left(\frac{\pi k}{4}\right)1[k].
  \]
  (This is the cosinusoid example used in the earlier example).
- \( G(s) \) has poles at \( s_{1,2} = -a \pm jb \) and \( -a - jb \).
- \( G(z) \) has poles at radius \( e^{-aT} \) angle \( \omega = \pm bT \) or at \( e^{-aT} \pm jbT \).
  - So, \( z_{1,2} = e^{s_{1,2}T} \) and \( e^{s_{2}T} \).
- In general, poles convert between the \( s \)-plane and \( z \)-plane via \( z = e^{sT} \).

**EXAMPLE:** Some corresponding pole locations:

![Diagram showing pole locations in s-plane and z-plane.](image)
- \( j \omega \)-axis maps to unit circle.

- Constant damping ratio \( \zeta \) maps to strange spiral.

- When considering system response to a step input for controls purposes, the following diagrams may be helpful:

  ![Diagram](image)

  - Damping \( \zeta \)
  - Frequency \( \omega_n \)
  - Settling time

- Higher-order systems:
  
  - Pole moving toward \( z = 1 \), system slows down.
  - Zero moving toward \( z = 1 \), overshoot.
  - Pole and zero moving close to each other cancel.
3.3: Discrete-time state-space form

Discrete-time systems can also be represented in state-space form.

\[ x[k + 1] = A_dx[k] + B_d u[k] \]
\[ y[k] = C_dx[k] + D_d u[k]. \]

The subscript "d" is used here to emphasize that, in general, the "A", "B", "C" and "D" matrices are different for discrete-time and continuous-time systems, even if the underlying plant is the same.

I will usually drop the "d" and expect you to interpret the system from its context.

Formulating from transfer functions

Discrete-time dynamics are represented as difference equations. e.g.,
\[ y[k + 3] + a_1 y[k + 2] + a_2 y[k + 1] + a_3 y[k] = b_1 u[k + 2] + b_2 u[k + 1] + b_3 u[k] \]
\[ y[k] + a_1 y[k - 1] + a_2 y[k - 2] + a_3 y[k - 3] = b_1 u[k - 1] + b_2 u[k - 2] + b_3 u[k - 3]. \]

This particular example has transfer function
\[ G(z) = \frac{b_1 z^2 + b_2 z + b_3}{z^3 + a_1 z^2 + a_2 z + a_3} = \frac{Y(z)}{U(z)}. \]

This transfer function may be converted to state-space in a very similar way to continuous-time systems.

First, consider the poles:

\[ G_p(z) = \frac{1}{z^3 + a_1 z^2 + a_2 z + a_3} = \frac{V(z)}{U(z)} \]
\[ \Rightarrow v[k + 3] + a_1 v[k + 2] + a_2 v[k + 1] + a_3 v[k] = u[k]. \]
Choose current and advanced versions of $v[k]$ as state.

$$x[k] = \begin{bmatrix} v[k + 2] & v[k + 1] & v[k] \end{bmatrix}^T.$$  

Then

$$x[k + 1] = \begin{bmatrix} v[k + 3] \\ v[k + 2] \\ v[k + 1] \end{bmatrix} = \begin{bmatrix} -a_1 & -a_2 & -a_3 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} v[k + 2] \\ v[k + 1] \\ v[k] \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} u[k].$$

We now add zeros.

$$G(z) = \frac{b_1 z^2 + b_2 z + b_3}{z^3 + a_1 z^2 + a_2 z + a_3} = \frac{Y(z)}{U(z)}.$$  

Break up transfer function into two parts. $\frac{V(z)}{U(z)}$ contains all of the poles of $\frac{Y(z)}{U(z)}$. Then,

$$Y(z) = \left[b_1 z^2 + b_2 z + b_3\right] V(z).$$

Or,

$$y[k] = b_1 v[k + 2] + b_2 v[k + 1] + b_3 v[k].$$

Then

$$x[k + 1] = \begin{bmatrix} -a_1 & -a_2 & -a_3 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} v[k + 2] \\ v[k + 1] \\ v[k] \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} u[k]$$

$$y[k] = \begin{bmatrix} b_1 & b_2 & b_3 \end{bmatrix} x[k] + \begin{bmatrix} 0 \end{bmatrix} u[k].$$
Many discrete-time transfer functions are not strictly proper. Solve by polynomial long division, and setting $D$ equal to the quotient.

In MATLAB, $[A, B, C, D] = \text{tf2ss}(\text{num}, \text{den})$ converts a transfer function form to state-space form (same syntax as for cts.-time, but $\text{num}$ and $\text{den}$ must have same order, maybe zero padded).

As with continuous-time systems, we have a lot of freedom when making state-space models (i.e., in choosing components of $x[k]$).

**Canonical forms**

In discrete-time we have the same canonical forms: Controller, observer, controllability, observability, modal, and Jordan.

They are derived in the same way, as demonstrated above for the controller form.

A block diagram for controller form is:

![Block diagram for controller form](image)

**Time (dynamic) response**

**Homogeneous part**

First, consider the scalar case

$x[k + 1] = ax[k], \quad x[0]$. 

Take $z$-transform. $X(z) = (z - a)^{-1}z.x[0]$.

Inverse $z$-transform. $x[k] = a^k x[0]$.

Similarly, the full solution (vector case) is

$$x[k] = A^k x[0].$$

**Aside: Nilpotent systems**

- $A$ is nilpotent if some power of $n$ exists such that

$$A^n = 0.$$

- $A$ does not just decay to zero, it is exactly zero!

- This might be a desirable control design! (Why?) You might imagine that all the eigenvalues of $A$ must be zero for this to work.

**Forced solution**

- The full solution is:

$$x[k] = A^k x[0] + \sum_{j=0}^{k-1} A^{k-1-j} Bu[j].$$

This can be proved by induction from the equation

$$x[k + 1] = Ax[k] + Bu[k], \quad x[0]$$

- Clearly, if $y[k] = Cx[k] + Du[k]$,

$$y[k] = CA^k x[0] + \sum_{j=0}^{k-1} CA^{k-1-j} Bu[k] + Du[k].$$
3.4: More on discrete-time state-space models

State-space to transfer function

- Start with the state equations
  \[ x[k + 1] = Ax[k] + Bu[k] \]
  \[ y[k] = Cx[k] + Du[k]. \]

- \( z \)-transform
  \[ zX(z) - zx[0] = AX(z) + BU(z) \]
  \[ Y(z) = CX(z) + DU(z) \]
  or
  \[ (zI - A)X(z) = BU(z) + zx[0] \]
  \[ X(z) = (zI - A)^{-1}BU(z) + (zI - A)^{-1}zx[0] \]
  and
  \[ Y(z) = [C(zI - A)^{-1}B + D]U(z) + C(zI - A)^{-1}zx[0]. \]

- So,
  \[ \frac{Y(z)}{U(z)} = C(zI - A)^{-1}B + D \]

- Same form as for continuous-time systems.
- Poles of system are roots of \( \det[zI - A] = 0 \).

Transformation

- State-space representations are not unique. Selection of state \( x \) are quite arbitrary.
Analyze the transformation of

\[ x[k + 1] = Ax[k] + Bu[k] \]

\[ y[k] = Cx[k] + Du[k] \]

Let \( x[k] = Tw[k] \), where \( T \) is an invertible (similarity) transformation matrix.

\[ w[k + 1] = \begin{pmatrix} T^{-1} & 0 \end{pmatrix} \begin{pmatrix} A & 0 \\ B & 0 \end{pmatrix} \begin{pmatrix} w[k] \\ u[k] \end{pmatrix} \]

\[ y[k] = \begin{pmatrix} C \\ D \end{pmatrix} \begin{pmatrix} w[k] \\ u[k] \end{pmatrix} \]

so, \( w[k + 1] = \tilde{A}w[k] + \tilde{B}u[k] \)

\[ y[k] = \tilde{C}w[k] + \tilde{D}u[k]. \]

Same as for continuous-time.

### Converting plant dynamics to discrete time

- Combine the dynamics of the zero-order hold and the plant.

- The continuous-time dynamics of the plant are:

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

\[ y(t) = Cx(t) + Du(t). \]

- Evaluate \( x(t) \) at discrete times. Recall

\[ x(t) = \int_0^t e^{A(t-\tau)} Bu(\tau) \, d\tau \]

\[ x[k + 1] = x((k + 1)T) = \int_0^{(k+1)T} e^{A((k+1)T-\tau)} Bu(\tau) \, d\tau \]
With malice aforethought, break up the integral into two pieces. The first piece will become \( A_d \) times \( x(kT) \). The second part will become \( B_d \) times \( u(kT) \).

\[
\begin{align*}
\int_0^{kT} e^{A((k+1)T-\tau)} Bu(\tau) \, d\tau &+ \int_{kT}^{(k+1)T} e^{A((k+1)T-\tau)} Bu(\tau) \, d\tau \\
&= \int_0^{kT} e^{AT} e^{A(kT-\tau)} Bu(\tau) \, d\tau + \int_{kT}^{(k+1)T} e^{A((k+1)T-\tau)} Bu(\tau) \, d\tau \\
&= e^{AT} x(kT) + \int_{kT}^{(k+1)T} e^{A((k+1)T-\tau)} Bu(\tau) \, d\tau.
\end{align*}
\]

In the remaining integral, note that \( u(\tau) \) is constant from \( kT \) to \( (k+1)T \), and equal to \( u(kT) \); let \( \sigma = (k+1)T - \tau \); \( \tau = (k+1)T - \sigma \); \( d\tau = -d\sigma \).

\[
x((k+1)T) = e^{AT} x(kT) + \left[ \int_0^T e^{A\sigma} B \, d\sigma \right] u(kT)
\]

or, \( x[k + 1] = e^{AT} x[k] + \left[ \int_0^T e^{A\sigma} B \, d\sigma \right] u[k] \).

So, we have a discrete-time state-space representation from the continuous-time representation.

\[
x[k + 1] = A_d x[k] + B_d u[k]
\]

where \( A_d = e^{AT} \), \( B_d = \int_0^T e^{A\sigma} B \, d\sigma \).

Similarly,

\[
y[k] = C x[k] + D u[k].
\]

That is, \( C_d = C \); \( D_d = D \).

**Calculating \( A_d \), \( B_d \), \( C_d \) and \( D_d \)**

- \( C_d \) and \( D_d \) require no calculation since \( C_d = C \) and \( D_d = D \).
\( A_d \) is calculated via the *matrix* exponential \( A_d = e^{AT} \). This is different from taking the exponential of each element in \( AT \).

- If MATLAB is handy, you can type in \( A_d = \text{expm}(A \times T) \)
- If MATLAB is not handy, then you need to work a little harder. Recall from the previous set of notes that \( e^{At} = \mathcal{L}^{-1}[(sI - A)^{-1}] \). So,

\[
e^{AT} = \mathcal{L}^{-1}[(sI - A)^{-1}]_{|t=T},
\]

which is probably the “easiest” way to work it out by hand. Or,

\[
e^{AT} = I + AT + \frac{A^2 T^2}{2!} + \frac{A^3 T^3}{3!} + \cdots
\]

which is a convergent series so may be approximated with only a few terms.

- Now we focus on computing \( B_d \). Recall that

\[
B_d = \int_0^T e^{A\sigma} B \, d\sigma
\]

\[
= \int_0^T \left( I + A\sigma + \frac{A^2 \sigma^2}{2} + \cdots \right) B \, d\sigma
\]

\[
= \left( IT + A \frac{T^2}{2!} + A^2 \frac{T^3}{3!} + \cdots \right) B
\]

\[
= A^{-1}(e^{AT} - I)B
\]

\[
= A^{-1}(A_d - I)B,
\]

if \( A^{-1} \) exists.

- So, calculating \( B_d \) is easy once we have already calculated \( A_d \).
- Also, in MATLAB, \([\text{Ad}, \text{Bd}] = \text{c2d}(A, B, T)\)
3.5: Linear time-varying and nonlinear discrete-time systems

**Linear time-varying discrete-time systems**

- A linear time-varying system can be written as
  \[
  y[k] = C[k]x[k] + D[k]u[k].
  \]

- Analysis is somewhat easier than for the continuous-time counterpart.
- Consider first the homogeneous case
  \[
  x[k + 1] = A[k]x[k], \quad x[k_0] = x_0, \quad k \geq 0.
  \]
- The solution to this is
  \[
  x[k] = \Phi[k, k_0]x_0
  \]
  where
  \[
  \Phi[k, k_0] = \begin{cases} 
  I, & k = k_0 \\
  A[k - 1]A[k - 2] \cdots A[k_0], & k > k_0.
  \end{cases}
  \]
- We see that the state-transition matrix can be computed readily, and does not involve any difficult integrals.
- Some properties of the state-transition matrix include:
  \[
  \Phi[k + 1, k_0] = A[k]\Phi[k, k_0] \\
  \Phi[k_0, k_0] = I \\
  \Phi[k, s]\Phi[s, \tau] = \Phi[k, \tau].
  \]
- The solution to the nonhomogeneous case can be shown to be
  \[
  x[k] = \Phi[k, k_0]x_0 + \sum_{\tau = k_0}^{k-1} \Phi[k, \tau + 1]B[\tau]u[\tau]
  \]
\[ y[k] = C[k]\Phi[k, k_0]x_0 + \sum_{\tau=k_0}^{k-1} C[k]\Phi[k, \tau + 1]B[\tau]u[\tau] + D[k]u[k]. \]

**Nonlinear discrete-time systems**

- The approach to working with nonlinear discrete-time systems is similar to that used for nonlinear continuous-time systems.
  - We can linearize around an equilibrium point or around a solution trajectory.
- We define a nonlinear state-space form as
  \[ x[k + 1] = f(x[k], u[k]) \]
  \[ y[k] = g(x[k], u[k]). \]

**Linearizing around an equilibrium point**

- We can linearize around an equilibrium point if there exists an equilibrium constant solution
  \[ u[k] = u^{eq}, \quad x[k] = x^{eq}, \quad y[k] = y^{eq}. \]
- Then, let the actual input signal and initial state be written as
  \[ u[k] = u^{eq} + \delta u[k], \quad x[k_0] = x^{eq} + \delta x^{eq}[k_0] \]
  for small \( \delta u[k] \).
- Then, we can write actual state and output as
  \[ x[k] = x^{eq} + \delta x[k], \quad y[k] = y^{eq} + \delta y[k]. \]
- Following the same process used in the prior chapter, the linearized perturbation system is
\[ \delta x[k + 1] = A\delta x[k] + B\delta u[k] \]
\[ \delta y[k] = C\delta x[k] + D\delta u[k], \]

where the linearized state-space matrices are

\[ A = \left( \frac{df(x_{eq}, u_{eq})}{dx} \right) ; \quad B = \left( \frac{df(x_{eq}, u_{eq})}{du} \right) \]
\[ C = \left( \frac{dg(x_{eq}, u_{eq})}{dx} \right) ; \quad D = \left( \frac{dg(x_{eq}, u_{eq})}{du} \right) \]

and the overall state and output can be computed as

\[ x[k] = x_{eq} + \delta x[k] \quad \text{and} \quad y[k] = y_{eq} + \delta y[k]. \]

**Linearizing around a solution trajectory**

- Alternately, suppose it is known that
  
  \[ u^{sol}[k], \quad x^{sol}[k], \quad y^{sol}[k] \]

  form a time-varying solution to the nonlinear dynamics of the system.

- This means that
  
  \[ x^{sol}[k + 1] = f(x^{sol}[k], u^{sol}[k]) \]
  \[ y^{sol}[k] = g(x^{sol}[k], u^{sol}[k]). \]

- Then, let general input, state, and output be
  
  \[ u[k] = u^{sol}[k] + \delta u[k] \]
  \[ x[k] = x^{sol}[k] + \delta x[k] \]
  \[ y[k] = y^{sol}[k] + \delta y[k]. \]

- Proceeding as before, we find the perturbation system
\[
\delta x[k + 1] \approx \left( \frac{df(x^\text{sol}, u^\text{sol})}{dx[k]} \right) \delta x[k] + \left( \frac{df(x^\text{sol}, u^\text{sol})}{du[k]} \right) \delta u[k]
\]
\[
\delta y[k] \approx \left( \frac{dg(x^\text{sol}, u^\text{sol})}{dx[k]} \right) \delta x[k] + \left( \frac{dg(x^\text{sol}, u^\text{sol})}{du[k]} \right) \delta u[k].
\]

- The overall state and output can be computed as
\[
x[k] = x^\text{sol} + \delta x[k] \quad \text{and} \quad y[k] = y^\text{sol} + \delta y[k].
\]
- Notice that even if the nonlinear system is time-invariant, the linearized system will be time-varying, in general.

**Where to from here?**

- We have now seen how to model both continuous-time and discrete-time systems in state-space form.
- There are many commonalities between the two, which we will take advantage of in the remainder of the course.
- The main difference is in interpreting the result. For example,
  - Eigenvalues of the continuous-time \( A \) matrix correspond to \( s \)-plane locations, and should be in the left-half \( s \)-plane for stability;
  - Eigenvalues of the discrete-time \( A \) matrix correspond to \( z \)-plane locations, and should be in the unit circle for stability.
- In fact, stability is a very important concept, which must be evaluated and ensured before we talk more about control.
- This will be our next topic.