Signals and Systems

University Electrical Engineering

What this course is, and how to use it

This is the sixth course in University Electrical Engineering: **signals and systems** . EE04 asked "what does THIS circuit do at THIS frequency"; this course asks the deeper question: what does a system do to ANY input – answered through the three transforms (convolution, Fourier, z) that the rest of the degree (communications, DSP, control) is built on.

- Unit 6.1 classifies signals and systems.
- Unit 6.2 makes **convolution** the universal time-domain answer.
- Units 6.3–6.4 move to frequency: Fourier series/transforms and sampling.
- Unit 6.5 reuses EE02's Laplace machinery as the system function.
- Units 6.6–6.7 go discrete: difference equations and the z-transform.

The running story is one audio channel: a microphone signal, a filter, a sampler, a digital equaliser. Every transform below moves that channel forward, and every answer is checked the way a signals engineer checks it – by deriving each transform pair at least once, verifying every convolution two ways (graphically and by the Fourier product when available), and testing every discrete result by direct iteration.

Signal representations and system properties

Why it matters. "Is it linear? Is it time-invariant?" are the first two questions of every analysis in this course – the answers decide which of all the machinery is even allowed.

The idea, with a picture. A **signal** is a function of time: continuous-time $x(t)$ or discrete-time $x[n]$. A **system** maps input signals to output signals. Four properties carry the course, and each is a test you run on ONE carefully chosen input, not a feeling:

- **Linearity** : $ax_1 + bx_2 \to ay_1 + by_2$ – feed the sum, check the sum of the outputs.
- **Time invariance** : shift the input by T; the output must shift by exactly T and nothing else.
- **Causality** : the output at time t depends only on x up to t.
- **BIBO stability** BIBO : every bounded input gives a bounded output; for an LTI system this collapses to $\int |h(t)|\,dt < \infty$ – the impulse response must be absolutely integrable.

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linearity

feed x1 + x2
output = y1 + y2?

sum in, sum out

time invariance

shift x by T
output shifts by T only?

shift in, shift out

causality

output at t = 0
depends on x(t > 0)?

no output before input

BIBO stability

bounded x in
bounded y out?

integral |h| < infinity

Each property is one experiment: pick the probe input, compare the two outputs

The derivation. Time invariance fails whenever the system's clock scales: for $y(t) = x(2t)$, shifting the input by T gives $y_{shift}(t) = x(2t - T) = x(2(t - T/2))$ – the output is the original system's output delayed by only $T/2$. Comparing "shift then system" ($x(t - T) \to y(t)$ at argument $2t$, i.e. $x(2t - T)$) against "system then shift" ($y(t - T) = x(2t - 2T)$) exposes the mismatch whenever it exists – that comparison IS the test.

> **Worked example 1 – the time-scaling trap.** Is $y(t) = x(2t)$ time-invariant? With $x(t) = u(t)$: shift-then-system gives $x(t-1) = u(t-1) \to u(2t-1) = u(t-\frac{1}{2})$; system-then-shift gives $y(t-1) = u(2(t-1)) = u(t-1)$. Half shift versus full shift – time-varying ✓. A 2× fast-playback stage in the running audio channel fails exactly this way.

> **Worked example 2 – stability by the integral.** The channel's room-echo model $h(t) = e^{-t}u(t)$: $\int_0^\infty e^{-t}\,dt = 1 < \infty$ ✓ stable. The same filter with $h(t) = u(t)$ (an integrator) gives $\int_0^\infty 1\,dt = \infty$ – a bounded step input produces an unbounded ramp output, so the integrator is BIBO-unstable however harmless it looks. One integral, one verdict.

Common mistakes. Testing linearity with one input (you need two, plus scaling). Declaring time invariance from a diagram (run the shift test). And confusing causality with stability: $h(t) = e^{-t}u(t)$ is both, but $h(t) = -e^{t}u(-t)$ (anti-causal, decaying) is stable and NOT causal – the two properties are independent.

Summary. Linearity + time invariance unlock convolution; causality orders cause and effect; BIBO stability is $\int |h| < \infty$. Every property is a one-probe experiment.

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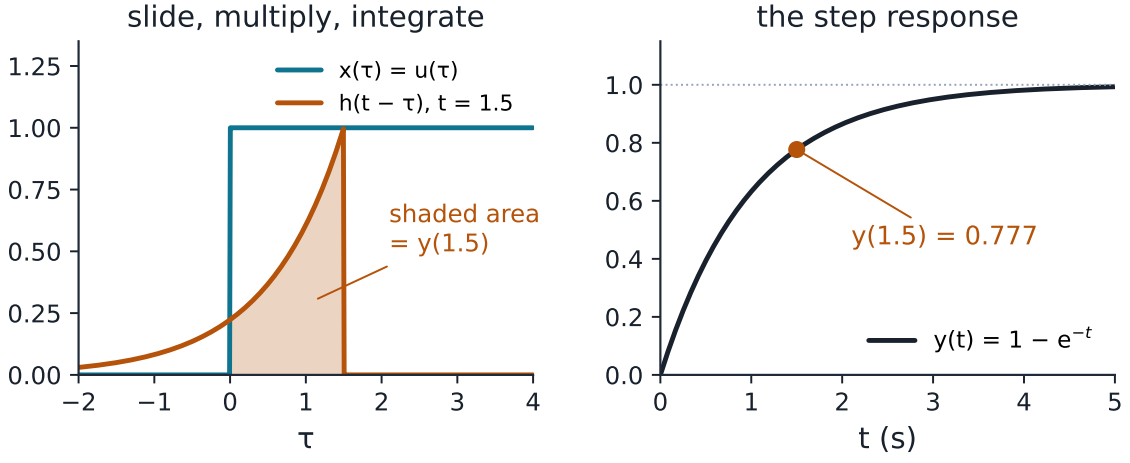
LTI systems and convolution

Why it matters. Convolution is the ONE formula that gives the output of any linear time-invariant system to ANY input – no transform, no transfer function, just the impulse response and an integral.

The idea, with a picture. For a **linear time-invariant** system, the **impulse response** $h(t)$ determines everything: decompose the input into delayed, scaled impulses; each impulse $\delta(t-\tau)\,x(\tau)\,d\tau$ produces $h(t-\tau)$ scaled; superposition integrates them all:

$$y(t) = \int_{-\infty}^{\infty} x(\tau)\,h(t-\tau)\,d\tau = (x * h)(t).$$

Convolution is slide–multiply–integrate: flip h, slide it past x, and each output value is the area of the overlap.



Each value of $y(t)$ is the overlap area as h slides past x

The derivation. Sifting: $x(t) = \int x(\tau)\delta(t-\tau)\,d\tau$. Linearity moves the integral outside; time invariance turns $\delta(t-\tau) \to h(t-\tau)$; the result is the convolution integral – three properties, one formula.

> **Worked example 1 – the classic step response.** The channel's filter $x(t) = u(t)$ into $h(t) = e^{-t}u(t)$: for $t \geq 0$ the overlap is $\tau \in [0, t]$, so
>
> $$y(t) = \int_0^t e^{-(t-\tau)}\,d\tau = e^{-t}(e^t - 1) = 1 - e^{-t}.$$

Check the ends: $y(0) = 0$ ✓ (no output before any input arrives); $y(\infty) = 1$ ✓ (DC gain 1 – and indeed $H(0) = 1$). At $t = 1.5$, $y = 1 - e^{-1.5} = 0.777$ – the shaded overlap area in the figure.

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Worked example 2 – escalate to the pulse. Now a 2-second rectangular input of height 1 (a held note) into the same filter. Two checkpoints by superposition: the step ON at $t = 0$ plus a step OFF at $t = 2$, so $y(t) = (1-e^{-t})u(t) - (1-e^{-(t-2)})u(t-2)$. At $t = 1$: $1 - e^{-1} = 0.632$. At $t = 3$: $(1 - e^{-3}) - (1 - e^{-1}) = e^{-1} - e^{-3} = 0.318$ – the release tail. Check the peak at $t = 2^-$: $1 - e^{-2} = 0.865$, and at $t = 2^+$ the second term subtracts exactly its start, so the curve is continuous with a slope discontinuity of -1 – the kink the ear hears as the note-off.

Common mistakes. Forgetting the flip (convolution is NOT $\int x(\tau)h(\tau)\,d\tau$). Dropping the limits when the signals start at 0 (the overlap window, not $(-\infty, \infty)$, does the work). And assuming $x * h = x \cdot h$ – multiplication is what happens in the FREQUENCY domain, never in time.

Summary. $y = x * h$; flip, slide, integrate; widths add ($W_1 + W_2$); commutative; cascades convolve. Every transform later exists to avoid doing this integral.

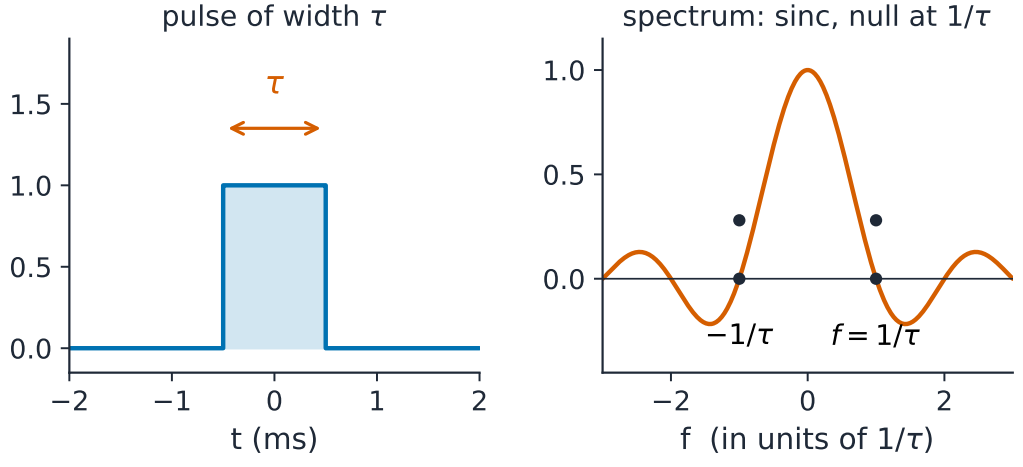
Fourier series and transforms

Why it matters. The channel's noise, its harmonics, its equaliser bands – all live in frequency. The Fourier transform is the bridge, and its convolution-to-multiplication property is why filters are designed there.

The idea, with a picture. Periodic signals expand into harmonics (EE02's series); the **Fourier transform** extends the idea to aperiodic signals:

$$X(\omega) = \int_{-\infty}^{\infty} x(t)\, e^{-j\omega t}\, dt, \qquad x(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} X(\omega)\, e^{j\omega t}\, d\omega.$$

Convolution in time becomes multiplication in frequency: $Y(\omega) = X(\omega)H(\omega)$ – the property the whole subject turns on.



The sinc spectrum of a rectangular pulse: null at 1/τ, narrower in time means wider in frequency

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The derivation. For a rectangular pulse of amplitude A and width τ centred at 0:

$$X(\omega) = \int_{-\tau/2}^{\tau/2} A\, e^{-j\omega t}\, dt = A\, \frac{e^{j\omega\tau/2} - e^{-j\omega\tau/2}}{j\omega} = A\tau\, \frac{\sin(\omega\tau/2)}{\omega\tau/2},$$

the sinc function. At $\omega = 0$ the transform equals the AREA $A\tau$ (and $\mathrm{sinc}(0) = 1$ returns it ✓); the first null sits at $\omega\tau/2 = \pi$, i.e. $f = 1/\tau$ – the **time–bandwidth trade-off** .

> **Worked example 1 – the channel's pulse budget.** A 1 ms click ($A = 1$, $\tau = 10^{-3}$ s): $X(f) = 10^{-3}\,\mathrm{sinc}(f \cdot 10^{-3})$, first null at $f = 1$ kHz. A ten-times-shorter click (0.1 ms) pushes the null to 10 kHz – the shorter the event, the wider the spectrum, which is why a click irritates every band of the equaliser at once. Check the area: the value at $f = 0$ is $A\tau$ – 10^{-3} for the 1 ms click and 10^{-4} for the 0.1 ms one ✓; the shorter click puts less into each band but spreads it over ten times as many.

> **Worked example 2 – escalate to the ideal filter's price.** The channel wants a brick-wall low-pass: gain 1 to 4 kHz, 0 above. Its impulse response is the inverse transform of the passband rectangle: $h(t) = \frac{\sin(\omega_c t)}{\pi t}$ with $f_c = 4$ kHz – a sinc in TIME, stretching infinitely in both directions, so the ideal filter is non-causal. Truncate it to $|t| \le 5$ ms (a 10 ms window) and the passband gains ripple (Gibbs), with ripples about $1/(10\text{ ms}) = 100$ Hz apart, and the edge can be no sharper than about that same 100 Hz. The trade-off is honest: steepness in frequency is paid in time-domain length, the same law from the other side.

Common mistakes. Evaluating sinc at ω rather than $\omega\tau/2$ (dimensionless argument, always). Forgetting the $1/2\pi$ on the inverse. And reading $|X(\omega)|$ as "the signal" – the transform is complete only with its phase.

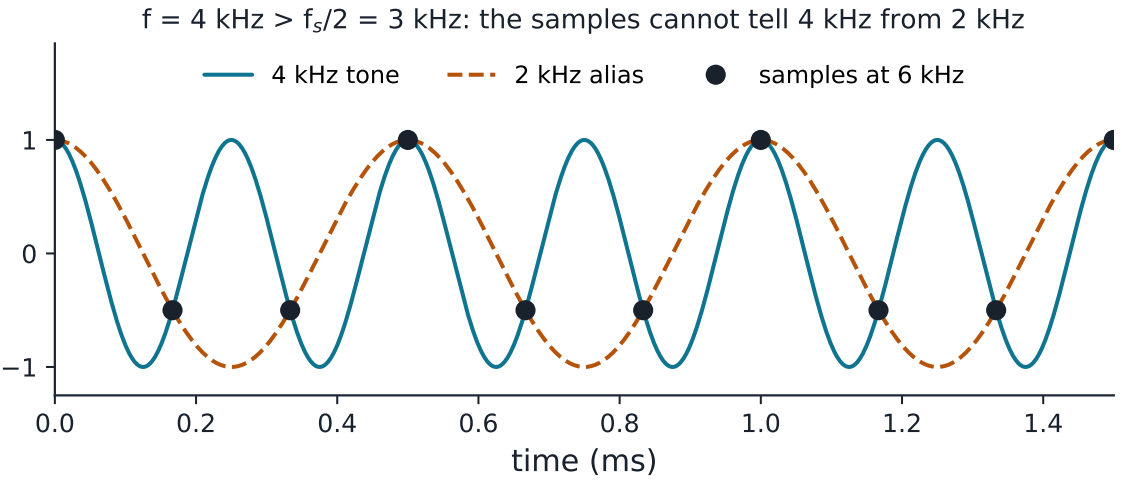
Summary. Transform pairs by the integral; pulse ↔ sinc with null at $1/\tau$; convolution ↔ multiplication. Steepness in one domain is length in the other.

Sampling

Why it matters. The channel goes digital at one point: the sampler. Everything downstream – the equaliser, the storage, the reconstruction – inherits whatever the sampling step did, and it can never be undone afterwards.

The idea, with a picture. Sampling multiplies the signal by an impulse train; in frequency, the spectrum replicates at multiples of f_s. The **Nyquist criterion** : a signal band-limited to f_{max} is perfectly recoverable if $f_s > 2f_{max}$. Below that, replicas overlap and **aliasing** folds high frequencies into the band as false lows.

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Sampled at 6 kHz, a 4 kHz tone is indistinguishable from 2 kHz

The derivation. A sampler output $x_s(t) = x(t)\sum_n \delta(t-nT)$ has, by the convolution-to- multiplication duality (multiplication in time = convolution in frequency), the spectrum $(1/T)\sum_k X(\omega - k\omega_s)$ – replicas at every multiple of the sampling frequency. If X occupies $|\omega| < \omega_s/2$, the replicas never touch; any narrower spacing overlaps them, and within the overlap no filter can tell a true low frequency from a folded high one.

> **Worked example 1 – the 4 kHz alias.** A 4 kHz cosine sampled at 6 kHz: $4 > 6/2 = 3$, so the sequence is indistinguishable from $|f - f_s| = 2$ kHz. Check sample by sample: $\cos(2\pi\cdot\frac{4}{6}n) = \cos(2\pi\cdot\frac{2}{3}n)$ and $\cos(2\pi\cdot\frac{2}{6}n) = \cos(2\pi\cdot\frac{1}{3}n)$; the two sequences agree because $\cos(2\pi\cdot\frac{2}{3}n) = \cos(2\pi(\frac{2}{3}-1)n) = \cos(2\pi\cdot(-\frac{1}{3})n) = \cos(2\pi\cdot\frac{1}{3}n)$ ✓ (cosine is even). A 4 kHz tone sampled at 6 kHz IS a 2 kHz sequence.

> **Worked example 2 – escalate to the anti-alias budget.** The channel's microphone carries content to 20 kHz; the sampler runs at 44.1 kHz ($f_s/2 = 22.05$ kHz). A 25 kHz interference tone would alias to $44.1 - 25 = 19.1$ kHz – inside the stated 20 kHz band. The **anti-alias filter** must cut it before the sampler: at 44.1 kHz CD sampling the guard band is only $22.05 - 20 = 2.05$ kHz wide, which is why professional capture oversamples at 96 or 192 kHz (guard bands of 28 and 76 kHz) and filters gently. The fix is always before the sampler, never after – a folded frequency carries no flag saying where it came from.

Common mistakes. Counting $f_s = 2f_{max}$ as safe (the inequality is strict; at equality a sine at f_{max} can be sampled at its zero crossings and vanish). Filtering after sampling (the fold is already inside the band). And assuming digital upsampling recovers what a low f_s missed – the information was never recorded.

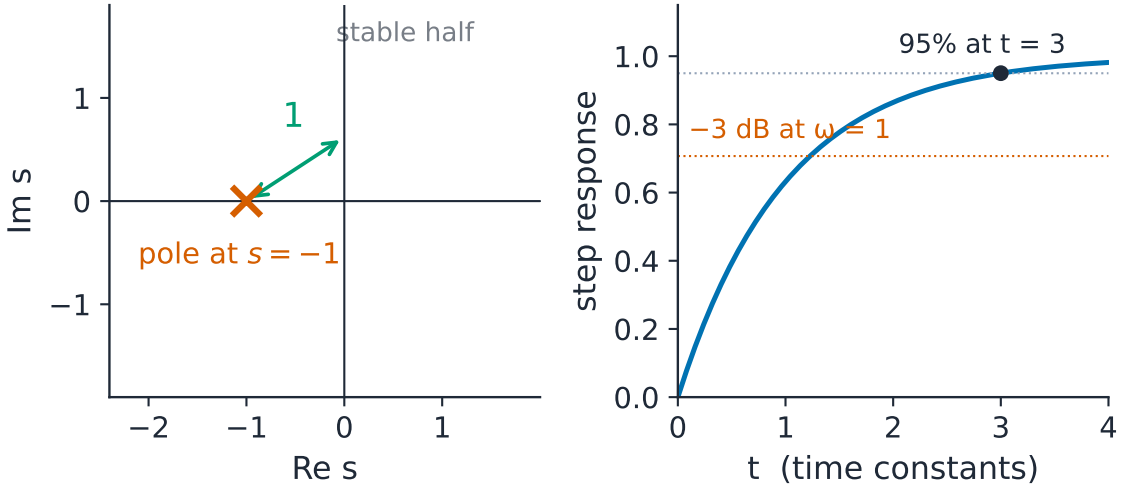
Summary. Sampling replicates the spectrum at f_s; $f_s > 2f_{max}$ or aliases fold; the anti-alias filter goes before the sampler; the guard band between f_{max} and $f_s/2$ is the design budget.

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Laplace transform and system functions

Why it matters. The channel's analogue stage – coupling capacitors, tone controls – is a rational system, and the poles of its system function predict ringing, speed and stability at a glance.

The idea, with a picture. EE02's Laplace transform becomes the **system function** $H(s) = Y(s)/X(s)$; for an LTI system, $h(t) \leftrightarrow H(s)$ as a transform pair. Rational $H(s)$ factors into **poles and zeros** : each pole contributes a natural mode e^{pt}; the region of convergence (ROC) decides which; causal stability requires all poles in the open left half-plane.



One pole at −1: the decay rate in time, the −3 dB frequency in the response – one number, three readings

The derivation. $\mathcal{L}\{e^{-at}u(t)\} = 1/(s+a)$ for $\mathrm{Re}(s) > -a$ (one integration by parts). The unit step $1/s$. So for $H(s) = 1/(s+1)$ driven by a step:

$$Y(s) = \frac{1}{s(s+1)} = \frac{1}{s} - \frac{1}{s+1} \implies y(t) = 1 - e^{-t},$$

partial fractions doing the convolution of Unit 6.2 with three lines of algebra – and both routes land on the same curve ✓.

> **Worked example 1 – read the pole three ways.** $H(s) = 1/(s+1)$: the pole at $s = -1$ IS the e^{-t} in the impulse response (time constant 1 s); the step response settles at $1 - e^{-t}$ reaching 95% at $t = 3$ s; the **frequency response** $H(j\omega)$ has $|H(j \cdot 1)| = 1/\sqrt{2} = 0.707$ – the −3 dB point sits exactly at the pole's distance from the origin, and the phase there is −45°.

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Worked example 2 – escalate to the two-pole tone control. A bass boost with a notch, $H(s) = \dfrac{s^2 + 2s + 100}{(s+1)(s+10)}$: zeros at $-1 \pm j9.95$ (close to the imaginary axis near ± 10 rad/s, so they carve a notch there: $|H(j10)| = 20/(10.05 \times 14.14) = 0.14$, about -17 dB), poles at -1 and -10 (both left – stable ✓). DC gain $100/10 = 10$ (20 dB of bass lift ✓); high-frequency gain 1 (the numerator's s^2 matches the denominator's). Check the mid-dip: at $\omega = 0$, 20 dB; at $\omega \to \infty$, 0 dB; the transition lives between the pole frequencies – the whole equaliser curve is legible from five marked points on the s-plane.

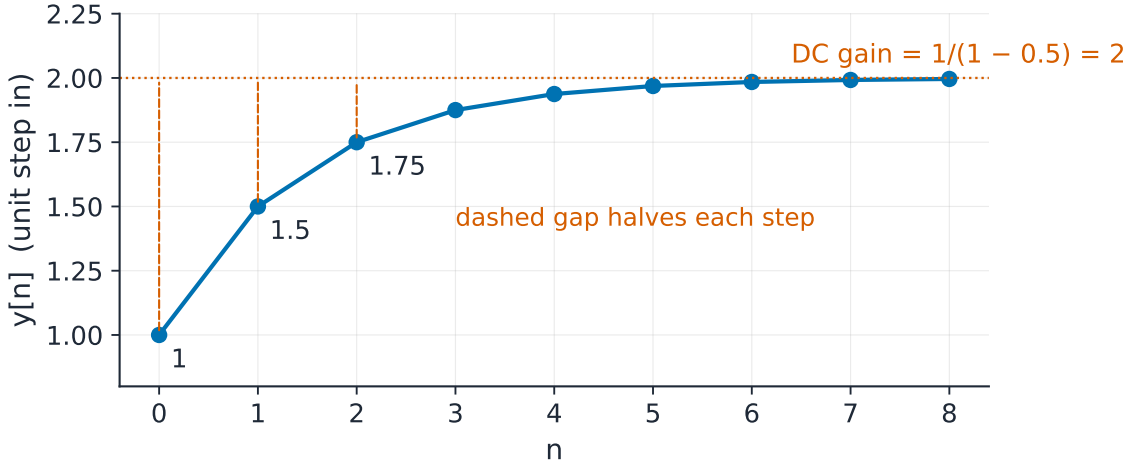
Common mistakes. Reading stability from the poles without the ROC (a pole at -1 with a left-extending ROC is anti-causal and unstable). Forgetting $H(s)$ only exists with its ROC. And trusting a pole-zero plot scaled wrong: the -3 dB point is the pole's DISTANCE from the origin only for a first-order factor with no nearby zeros.

Summary. $H(s) = Y/X$; partial fractions turn transforms back into exponentials; poles set the modes, the ROC the causality, the left half-plane the stability, and $s = j\omega$ the frequency response.

Discrete-time systems

Why it matters. Once sampled, the channel's equaliser is a difference equation running on a processor – iterating it by hand is how its behaviour is actually understood.

The idea, with a picture. The discrete mirror of convolution: $y[n] = \sum_k x[k]\, h[n-k]$. Systems described by **difference equations** such as $y[n] = 0.5\, y[n-1] + x[n]$ – an accumulator with decay. The homogeneous mode is a^n where a is the root of the characteristic equation; iterating one step at a time is the direct solution.



Each step halves the remaining gap to the limit – a geometric approach you can verify on your fingers

The derivation. For $y[n] = a\, y[n-1] + x[n]$ with step input and zero initial condition,

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the first samples are $1,\ 1+a,\ 1+a+a^2,\ \ldots$ – partial sums of a geometric series, so

$$y[n] = \frac{1-a^{n+1}}{1-a} \to \frac{1}{1-a} \quad (|a|<1),$$

and the limit $1/(1-a)$ is the DC gain: the discrete echo of the continuous $\int h < \infty$ condition.

> **Worked example 1 – iterate and converge.** $y[n] = 0.5y[n-1]+x[n]$, $y[-1]=0$, unit step: $y[0]=1$, $y[1]=1.5$, $y[2]=1.75$, $y[3]=1.875$ – closing on the DC gain $1/(1-0.5)=2$ ✓. The gap to the limit halves each step (the residue is 0.5^{n+1}), so after 10 steps the gap is $0.5^{11} \approx 0.05\%$ – one line of arithmetic predicting the whole table.

> **Worked example 2 – escalate to an oscillating root.** Set $a=-0.8$: $y[n] = -0.8\,y[n-1]+x[n]$ on a step gives $y[n] = \frac{1}{1.8}(1-(-0.8)^{n+1})$ – the approach to 0.556 oscillates: $y[0]=1$, $y[1]=0.2$, $y[2]=0.84$, $y[3]=0.328$. Check $|a|=0.8<1$ – still stable (the oscillation decays), and the sign flips every sample are the fingerprint of a negative root, the discrete analogue of a complex pole pair ringing at half the sampling rate ($z=-0.8$ maps to $s=(\ln 0.8 \pm j\pi)/T$).

Common mistakes. Iterating with a wrong $y[-1]$ (state the initial condition before the first line). Declaring stability from the first few samples (only $|a|<1$ decides). And reading the DC gain from $n=0$ alone – it is $1/(1-a)$, the LIMIT, visible only as the samples settle.

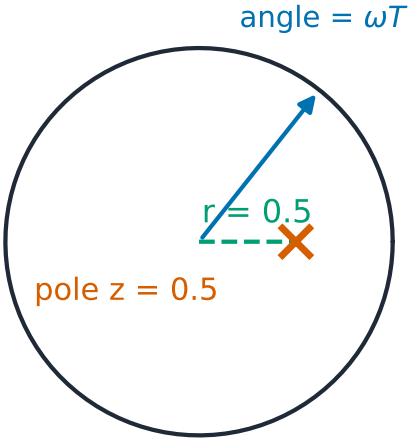
Summary. Difference equation → iterate; geometric partial sums; DC gain $1/(1-a)$; root inside the unit circle = stability; negative or complex roots = oscillation.

The z-transform

Why it matters. The equaliser ships as coefficients a_k, b_k of a difference equation; the z-transform is how those numbers are chosen, and the **unit circle** is where their frequency response is read off.

The idea, with a picture. The **z-transform** z $X(z) = \sum_n x[n]z^{-n}$ plays Laplace's role in discrete time. The fundamental pair: $a^n u[n] \leftrightarrow 1/(1-az^{-1})$ with ROC $|z|>|a|$. For a causal system, poles INSIDE the unit circle mean stability. The inverse transform is long division or partial fractions.

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the circle is the frequency axis; radius 0.5 = damping -693 s^{-1} at T = 1 ms

The unit circle is the frequency axis; the pole radius is the damping – one picture joins the two domains

The derivation. $\sum_{n=0}^{\infty} a^n z^{-n}$ is geometric with ratio az^{-1}; it converges to $1/(1-az^{-1})$ exactly when $|az^{-1}| < 1$, i.e. $|z| > |a|$ – the ROC is the exterior of a circle through the pole. Stability needs the ROC to include the unit circle (the frequency axis $z = e^{j\omega}$), which forces $|a| < 1$ – the unit-circle condition in one line.

Worked example 1 – the accumulator again, by transform. $H(z) = 1/(1 - 0.5z^{-1})$, pole at $z = 0.5$ – inside the unit circle, stable ✓. Step response: $Y(z) = H(z)\,\dfrac{z}{z-1} = \dfrac{z^2}{(z-0.5)(z-1)}$; partial fractions give $y[n] = 2(1 - 0.5^{n+1})$. Check $n = 0$: $2(1 - 0.5) = 1$ ✓ matches the iteration of Unit 6.6, and $n \to \infty$ gives 2 ✓ – the DC gain again. Two methods, one sequence.

Worked example 2 – the plane-to-plane dictionary. The mapping $s \to z = e^{sT}$ sends $s = j\omega$ to the unit circle $e^{j\omega T}$ (frequency = angle), and a continuous decay rate σ to a radius $e^{\sigma T}$. The accumulator's pole at $z = 0.5$ with $T = 1$ ms corresponds to $\sigma = \ln(0.5)/0.001 = -693$ s^{-1} – a time constant of $1/693 = 1.44$ ms. Check against Unit 6.6's iteration: the gap halves per SAMPLE ($0.5 = e^{-0.693}$), and $0.693/T$ per second is the same -693 s^{-1} ✓. The two transforms are one subject seen from two banks of the same river.

Common mistakes. Quoting $H(z)$ without its ROC (the same rational function can describe different systems when the ROC changes). Reading frequency response off the real axis (it lives on $|z| = 1$). And confusing z and z^{-1} conventions – for the causal sequence $a^n u[n]$, the ROC is $|z| > |a|$; a left-sided sequence has a different ROC.

Summary. $a^n \leftrightarrow 1/(1 - az^{-1})$, ROC exterior; for a causal system, poles inside the unit circle = stability; $z = e^{sT}$ maps decay to radius and frequency to angle – and the

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DC gain $1/(1-a)$ closes the loop with the iteration two units ago.

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