ECE103 Lecture 24, Nov. 28, 2018

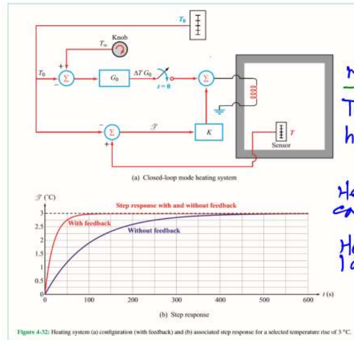
Quiz 8 today (This is the Last Quiz)

Midterm Average = 69.3 $\sigma = 16.3$

Final Exam on Dec. 12 (W) 4–7 pm

- It will cover the entire course materials
- You are allowed to bring 2 pages of formulas only. No solved problems

$\boxed{\text{Course score}} = 2 \times \left(\begin{array}{l}\text{Average of}\\ \text{best 6 Qz scores}\end{array}\right) + 0.3\ \text{Midterm} + 0.5\ \text{Final}$



(a) Closed-loop mode heating system

(b) Step response

Figure 4-32: Heating system (a) configuration (with feedback) and (b) associated step response for a selected temperature rise of 3 °C.

The Laplace transform of $x(t)$ is

$$\mathbf{X}(\mathbf{s}) = \left(\frac{\Delta T\, G_0}{C}\right)\frac{1}{\mathbf{s}},$$

and the output step response is

$$\mathbf{Y}(\mathbf{s}) = \mathbf{Q}(\mathbf{s})\,\mathbf{X}(\mathbf{s}) = \left(\frac{\Delta T\, G_0}{C}\right)\frac{1}{\mathbf{s}(\mathbf{s}+b)}.$$

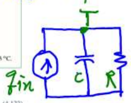
Partial fraction expansion leads to

$$\mathbf{Y}(\mathbf{s}) = \left(\frac{\Delta T\, G_0}{bC}\right)\left[\frac{1}{\mathbf{s}} - \frac{1}{\mathbf{s}+b}\right], \quad (4.123)$$

which has the inverse Laplace transform

$$\mathscr{T}(t) = \frac{\Delta T\, G_0}{bC}\,[1 - e^{-bt}]\,u(t). \quad (4.124)$$

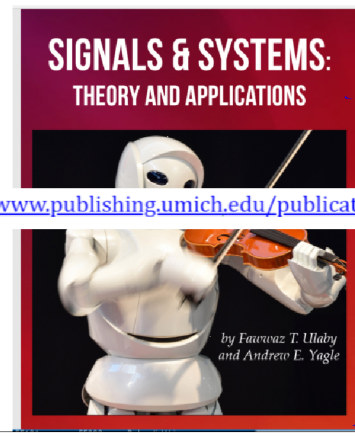
Mech		Elect.
Temp	T	v
heat flow	q	i
Heat capacity	C	C
Heat loss	R	R



Physical Systems

Electrical, Mechanical, Fluidic

QUANTITY	PROTOTYPE	ELECTRICAL	TRANSLATIONAL	ROTATIONAL	FLUID-FLOW
Flow	Flow, f	Current, i	Force, f	Torque, τ	Volume flow, q
Potential	Potential, p	Voltage, v	Velocity, v	Angular vel., ω	Pressure, P
Flow equation	$f = Yp$	$i = Yv$	$f = Yv$	$\tau = Y\omega$	$q = Yp$
Instantaneous power (watts)	$\mathscr{P} = fp$	$\mathscr{P} = iv$	$\mathscr{P} = fv$	$\mathscr{P} = \tau\omega$	$\mathscr{P} = qp$
Other variables		Charge $q = \int i\,dt$	Displacement $x = \int v\,dt$	Angle, $\theta = \int \omega\,dt$	
Integrating element, (as an admittance) and flow equation	$f = A\int p\,dt$	$\Gamma = 1/L$, $i = \frac{1}{L}\int v\,dt$, L = Inductance	$f = K\int v\,dt$, K = Spring constant	$\tau = K\int \omega\,dt$, K = Spring constant	$Y = 1/I$, $q = \frac{1}{I}\int p\,dt$, I = Inertance
i–d Admittance	$Y = A\int dt$	$\frac{1}{L}\int dt$	$K\int dt$	$K\int dt$	$\frac{1}{I}\int dt$
i–d Impedance	$\frac{1}{A}\frac{d}{dt}$	$L\frac{d}{dt}$	$\frac{1}{K}\frac{d}{dt}$	$\frac{1}{K}\frac{d}{dt}$	$I\frac{d}{dt}$
Stored energy (joules)	$\frac{f^2}{2A}$	$\frac{Li^2}{2}$	$\frac{f^2}{2K}$ or $\frac{Kx^2}{2}$	$\frac{\tau^2}{2K}$ or $\frac{K\theta^2}{2}$	$\frac{Iq^2}{2}$
Initial condition	$f(0)$	$i(0)$	$f(0)$ or $x(0)$	$\tau(0)$ or $\theta(0)$	$q(0)$
Proportional element, (as an admittance) and flow equation	$f = Bp$	$G = 1/R$, $i = Gv = v/R$, R = Resistance	$f = Bv$, B = Viscous friction	$\tau = D\omega$, D = Viscous friction	$q = p/R$, R = Fluid resistance
i–d Admittance	B	G or $1/R$	B	D	$1/R$
i–d Impedance	$1/B$	R	$1/B$	$1/D$	R
Power (watts)	$fp = f^2/B = p^2B$	$iv = i^2R = v^2/R$	$fv = f^2/B = v^2B$	$\tau\omega = \tau^2/D = \omega^2D$	$qp = q^2R = p^2/R$
Differentiating element, (as an admittance) and flow equation	$f = C\frac{dp}{dt}$	$i = C\frac{dv}{dt}$, C = Capacitance	$f = M\frac{dv}{dt}$, M = Mass	$\tau = J\frac{d\omega}{dt}$, J = Moment of inertia	$q = C\frac{dp}{dt}$, C = Fluid capacitance
i–d Admittance	$C\frac{d}{dt}$	$C\frac{d}{dt}$	$M\frac{d}{dt}$	$J\frac{d}{dt}$	$C\frac{d}{dt}$
i–d Impedance	$\frac{1}{C}\int dt$	$\frac{1}{C}\int dt$	$\frac{1}{M}\int dt$	$\frac{1}{J}\int dt$	$\frac{1}{C}\int dt$
Stored energy (joules)	$\frac{Cp^2}{2}$	$\frac{Cv^2}{2}$	$\frac{Mv^2}{2}$	$\frac{J\omega^2}{2}$	$\frac{Cp^2}{2}$
Initial condition	$p(0)$	$v(0)$ or $q(0)$	$v(0)$	$\omega(0)$	$p(0)$



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$N(s) = (s+j2)(s-j2)(s-\epsilon+j4)(s-\epsilon-j4)$

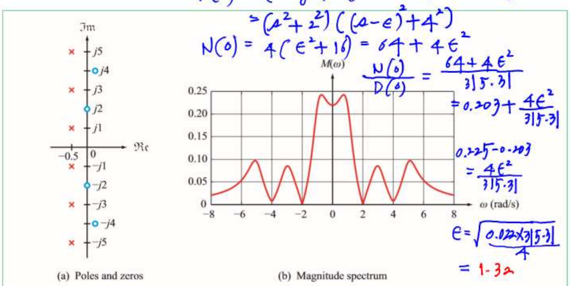
$= (s^2+2^2)((s-\epsilon)^2+4^2)$

$N(0) = 4(\epsilon^2+16) = 64 + 4\epsilon^2$

$\frac{N(0)}{D(0)} = \frac{64+4\epsilon^2}{315.31} = 0.203 + \frac{4\epsilon^2}{315.31}$

$0.225 - 0.203 = \frac{4\epsilon^2}{315.31}$

$\epsilon = \sqrt{\frac{0.022 \times 315.31}{4}} = 1.32$



(a) Poles and zeros

(b) Magnitude spectrum

Figure 6-25: Magnitude spectrum of a frequency response composed of 2 pairs of conjugate zeros and 3 pairs of conjugate poles (Example 6-6).

$D(s) = (s+0.5+j1)(s+0.5-j1)(s+0.5+j3)(s+0.5-j3)(s+0.5+j5)(s+0.5-j5)$

$= ((s+0.5)^2+1)((s+0.5)^2+3^2)((s+0.5)^2+5^2)$

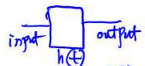
$D(0) = 1.25 \times 9.25 \times 25.25 = 315.31$

$$H(s) = K\left[\frac{(s+z_1)(s+z_2)\cdots}{(s+p_1)(s+p_2)\cdots}\right]$$

$\times 4$

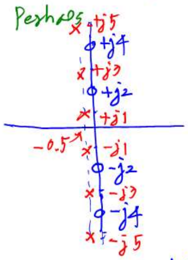
$(0.25 \to 1.0)$

⇒ Some errors in this example



$H(s), H(\omega)$

Perhaps



$$H(s)=\frac{(s+j2)(s-j2)(s+j4)(s-j4)}{(s+0.5+j1)(s+0.5-j1)(s+0.5+j3)(s+0.5-j3)\times(s+0.5+j5)(s+0.5-j5)}$$

$$=\frac{(s^2+4)(s^2+16)}{((s+0.5)^2+1)((s+0.5)^2+9)((s+0.5)^2+25)}$$

$$h(0_+)=\lim_{s\to\infty} sH(s)\simeq\lim_{s\to\infty}\frac{s^5}{s^6}=0$$

$e^{-0.5t}$ effect ⟶ $$h(\infty)=\lim_{s\to 0} sH(s)=\lim_{s\to 0}\frac{s\,(4)(16)}{(1.25)(9.25)(25.25)}=0$$

Exercise 6-8: An LTI system has zeros at $\pm j3$. What sinusoidal signals will it eliminate?

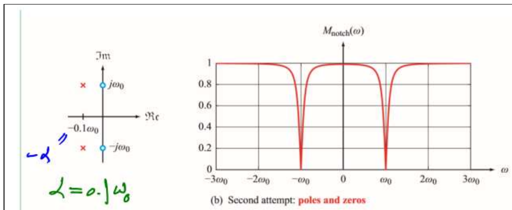
Answer: $A\cos(3t+\theta)$ for any constant A and θ. (See S2) which includes $A\sin 3t$

Exercise 6-9: An LTI system has poles at $-0.1\pm j4$. What sinusoidal signals will it emphasize?

Answer: $A\cos(4t+\theta)$ for any constant A and θ. (See S2)

$$\frac{N(s)}{(s+0.1+j4)(s+0.1-j4)}=\frac{}{(s+0.1)^2+\underbrace{4^2}_{\omega_0^2}}$$

$$\rightarrow e^{-0.1t}\cos(4t+\theta)$$



(b) Second attempt: **poles and zeros**

Figure 6-27: Two attempts at designing a filter to reject $\omega_0=\pm 120\pi$ (rad/s); i.e., $f_0=60$ Hz.

by replacing $j\omega$ with $\mathbf{s}$ in Eq. (6.74),

$$\begin{aligned}\mathbf{H}_{\text{notch}}(\mathbf{s})&=\mathbf{H}_{\text{notch}}(\omega)\big|_{j\omega=\mathbf{s}}\\&=\frac{(\mathbf{s}-j\omega_0)(\mathbf{s}+j\omega_0)}{(\mathbf{s}+\alpha-j\omega_0)(\mathbf{s}+\alpha+j\omega_0)}\\&=\frac{\mathbf{s}^2+\omega_0^2}{(\mathbf{s}+\alpha)^2+\omega_0^2}\\&=\frac{\mathbf{s}^2+2\alpha\mathbf{s}+\alpha^2+\omega_0^2}{(\mathbf{s}+\alpha)^2+\omega_0^2}-\frac{(2\alpha\mathbf{s}+\alpha^2)}{(\mathbf{s}+\alpha)^2+\omega_0^2}\\&=1-\frac{(2\alpha\mathbf{s}+\alpha^2)}{(\mathbf{s}+\alpha)^2+\omega_0^2},\end{aligned}\qquad(6.76)$$

$\mathcal{L}^{-1}$

where in the last step, we applied the division relationship given by Eq. (3.55) to convert $\mathbf{H}(\mathbf{s})$ from a ***proper rational function*** (numerator and denominator of the same degree) into a form $\mathbf{H}_{\text{notch}}=1+\mathbf{G}(\mathbf{s})$, where $\mathbf{G}(\mathbf{s})$ is a ***strictly*** proper rational function (degree of numerator smaller than that of denominator). Application of partial fraction expansion followed by transformation to the time domain leads to the impulse response:

$$h_{\text{notch}}(t)=\delta(t)-Ae^{-\alpha t}\cos(\omega_0 t+\theta)\,u(t)\qquad(6.77)$$

with

$$A=\left[4\alpha^2+\frac{\alpha^4}{\omega_0^2}\right]^{1/2},\quad\text{and}\quad\theta=\tan^{-1}\left(\frac{\alpha}{2\omega_0}\right).\qquad(6.78)$$

$$\frac{2\alpha s+\alpha^2}{(s+\alpha)^2+\omega_0^2}=\frac{2\alpha(s+\alpha)-\alpha^2}{(s+\alpha)^2+\omega_0^2}$$

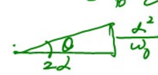
$$=2\alpha\left[\frac{(s+\alpha)}{(s+\alpha)^2+\omega_0^2}\right]-\frac{\alpha^2}{\omega_0}\,\frac{\omega_0}{(s+\alpha)^2+\omega_0^2}$$

$\downarrow \mathcal{L}^{-1}$

$$e^{-\alpha t}\left(2\alpha\cos\omega_0 t-\frac{\alpha^2}{\omega_0}\sin\omega_0 t\right)$$

$$=e^{-\alpha t}\underbrace{\sqrt{(2\alpha)^2+\left(\frac{\alpha^2}{\omega_0}\right)^2}}_{A}\left(\frac{2\alpha}{A}\cos\omega_0 t-\frac{\alpha^2/\omega_0}{A}\sin\omega_0 t\right)$$

$$=B\cos(\omega_0 t+\theta)$$



$$\tan\theta=\frac{\alpha^2/\omega_0}{2\alpha}$$

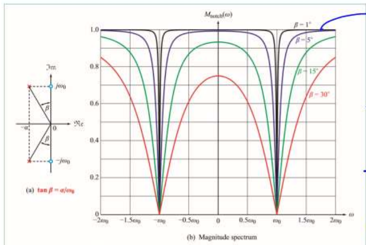
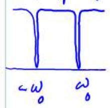


Figure 6-28: Notch-filter magnitude spectra for various values of β.

For $\beta = 1°$



$\tan 1° = \frac{\alpha}{\omega_0}$

$\alpha = 0.017\,\omega_0$

Example 6-7: Rejecting an Interfering Tone

Design a notch filter to reject a 1 kHz interfering sinusoid. The filter's impulse response should decay to less than 0.01 within 0.1 s.

Solution: At 1 kHz,

$$\omega_0 = 2\pi f_0 = 2\pi \times 10^3 = 6283 \text{ rad/s}.$$

To satisfy the impulse-response decay requirement, the value of α should be such that the coefficient of the cosine term in Eq. (6.77) is no larger than 0.01 at $t = 0.1$ s. That is,

$$\left(4\alpha^2 + \frac{\alpha^4}{\omega_0^2}\right)^{1/2} e^{-0.1\alpha} \le 0.01.$$

Solution of this inequality (using MATLAB or MathScript) leads to $\alpha \approx 100 \text{ s}^{-1}$ and, in turn, to

$$\beta = \tan^{-1}\left(\frac{\alpha}{\omega_0}\right) = \tan^{-1}\left(\frac{100}{6283}\right) = 0.9°.$$

With both ω_0 and α specified, we now can calculate and plot the impulse response given by Eq. (6.77) (excluding the impulse function), as shown in Fig. 6-29.

Note that although the maximum value of $h_{notch}(t)$, excluding the impulse, is 200, $|h_{notch}(0.1)| = 0.009$, which is smaller than the specified value of 0.01. The magnitude spectrum of this notch filter is essentially the $\beta = 1°$-plot shown in Fig. 6-28(b).

Concept Question 6-9: Why do we need poles in a notch filter? Why not just use zeros only? (See S²)

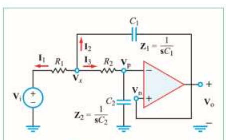


Figure 6-43: Sallen-Key circuit.

(a) Obtain the transfer function $\mathbf{H}(\mathbf{s}) = \mathbf{V}_o(\mathbf{s})/\mathbf{V}_i(\mathbf{s})$.

~~(b) Obtain the transfer function of a second-order Butterworth~~ filter with cutoff frequency $\omega_c = 10^3$ rad/s and arbitrary dc gain $\mathbf{C}$.

(c) ~~Match the transfer functions in (a) and (b) to specify the~~ values of R_1, R_2, C_1, and C_2.

Solution:

(a) At node $\mathbf{V}_x$, KCL gives

$$\frac{\mathbf{V}_x - \mathbf{V}_i}{R_1} + \frac{\mathbf{V}_x - \mathbf{V}_p}{R_2} + \frac{\mathbf{V}_x - \mathbf{V}_0}{\mathbf{Z}_1} = 0, \qquad (6.102)$$

and since $\mathbf{V}_o = \mathbf{V}_n = \mathbf{V}_p$, voltage division gives

$$\mathbf{V}_x = \mathbf{V}_o\left(\frac{R_2 + \mathbf{Z}_2}{\mathbf{Z}_2}\right). \qquad (6.103)$$

The combination of the two equations leads to

$$\mathbf{H}(\mathbf{s}) = \frac{\frac{1}{R_1 R_2 C_1 C_2}}{\mathbf{s}^2 + \left(\frac{1}{R_1 C_1} + \frac{1}{R_2 C_1}\right)\mathbf{s} + \frac{1}{R_1 R_2 C_1 C_2}}. \qquad (6.104)$$

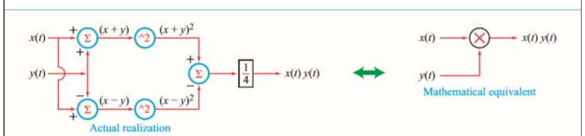
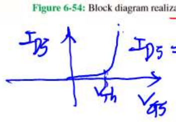
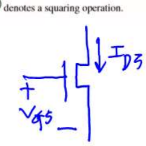


Figure 6-54: Block diagram realization of the product $x(t)\ y(t)$. The symbol ^2 denotes a squaring operation.



$I_{DS} = \frac{k}{2}(V_{GS} - V_{th})^2$



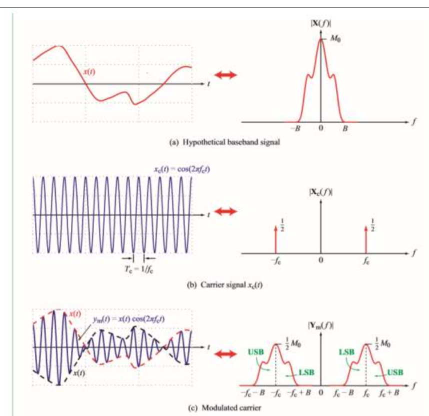


Figure 6-56: Baseband signal $x(t)$, carrier signal $x_c(t)$, and modulated carrier $y_m(t)$.

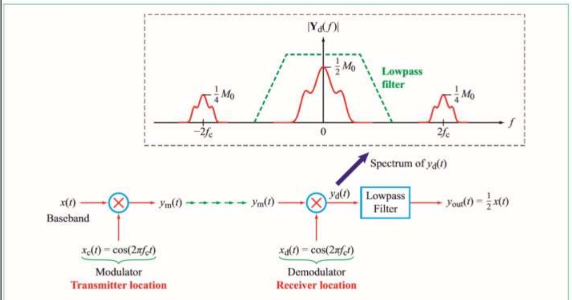


Figure 6-57: The DSB demodulator uses a replica of the carrier signal, followed by a lowpass filter, to recover the baseband signal $x(t)$.

Table 5-4: Fourier series expressions for a select set of periodic waveforms.

Waveform	Fourier Series
1. Square Wave	$x(t) = \sum_{n=1}^{\infty} \frac{4A}{n\pi} \sin\left(\frac{n\pi}{2}\right) \cos\left(\frac{2n\pi t}{T_0}\right)$
2. Time-Shifted Square Wave	$x(t) = \sum_{\substack{n=1\\ n=\text{odd}}}^{\infty} \frac{4A}{n\pi} \sin\left(\frac{2n\pi t}{T_0}\right)$
3. Pulse Train	$x(t) = \frac{A\tau}{T_0} + \sum_{n=1}^{\infty} \frac{2A}{n\pi} \sin\left(\frac{n\pi\tau}{T_0}\right) \cos\left(\frac{2n\pi t}{T_0}\right)$
4. Triangular Wave	$x(t) = \sum_{\substack{n=1\\ n=\text{odd}}}^{\infty} \frac{8A}{n^2\pi^2} \cos\left(\frac{2n\pi t}{T_0}\right)$
5. Shifted Triangular Wave	$x(t) = \sum_{\substack{n=1\\ n=\text{odd}}}^{\infty} \frac{8A}{n^2\pi^2} \sin\left(\frac{n\pi}{2}\right) \sin\left(\frac{2n\pi t}{T_0}\right)$
6. Sawtooth	$x(t) = \sum_{n=1}^{\infty} (-1)^{n+1} \frac{2A}{n\pi} \sin\left(\frac{2n\pi t}{T_0}\right)$
7. Backward Sawtooth	$x(t) = \frac{A}{2} + \sum_{n=1}^{\infty} \frac{A}{n\pi} \sin\left(\frac{2n\pi t}{T_0}\right)$
8. Full-Wave Rectified Sinusoid	$x(t) = \frac{2A}{\pi} + \sum_{n=1}^{\infty} \frac{4A}{\pi(1-4n^2)} \cos\left(\frac{2n\pi t}{T_0}\right)$
9. Half-Wave Rectified Sinusoid	$x(t) = \frac{A}{\pi} + \frac{A}{2} \sin\left(\frac{2\pi t}{T_0}\right) + \sum_{\substack{n=2\\ n=\text{even}}}^{\infty} \frac{2A}{\pi(1-n^2)} \cos\left(\frac{2n\pi t}{T_0}\right)$

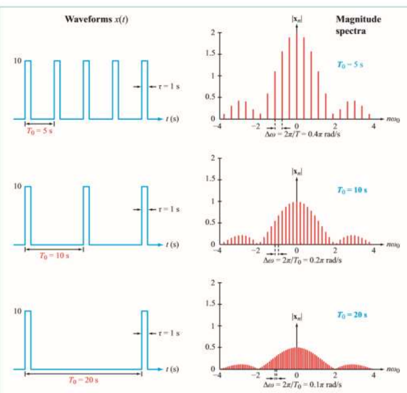


Figure 5-12: Line spectra for pulse trains with $T_0/\tau = 5$, 10, and 20.

Table 5-6: Examples of Fourier transform pairs. Note that constant $a \geq 0$.

	$x(t)$		$\mathbf{X}(\omega) = \mathcal{F}[x(t)]$
	BASIC FUNCTIONS		
1.	$\delta(t)$	⟷	1
1a.	$\delta(t - t_0)$	⟷	$e^{-j\omega t_0}$
2.	1	⟷	$2\pi\,\delta(\omega)$
3.	$u(t)$	⟷	$\pi\,\delta(\omega) + 1/j\omega$
4.	$\text{sgn}(t)$	⟷	$2/j\omega$
5.	$\text{rect}(t/\tau)$	⟷	$\tau\,\text{sinc}(\omega\tau/2)$
6.	$\frac{e^{-t^2/(2\sigma^2)}}{\sqrt{2\pi\sigma^2}}$	⟷	$e^{-\omega^2\sigma^2/2}$
7a.	$e^{-at}\,u(t)$	⟷	$1/(a + j\omega)$
7b.	$e^{at}\,u(-t)$	⟷	$1/(a - j\omega)$
8.	$\cos\omega_0 t$	⟷	$\pi[\delta(\omega - \omega_0) + \delta(\omega + \omega_0)]$
9.	$\sin\omega_0 t$	⟷	$j\pi[\delta(\omega + \omega_0) - \delta(\omega - \omega_0)]$
	ADDITIONAL FUNCTIONS		
10.	$e^{j\omega_0 t}$	⟷	$2\pi\,\delta(\omega - \omega_0)$
11.	$te^{-at}\,u(t)$	⟷	$1/(a + j\omega)^2$
12a.	$[e^{-at}\sin\omega_0 t]\,u(t)$	⟷	$\omega_0/[(a + j\omega)^2 + \omega_0^2]$
12b.	$[\sin\omega_0 t]\,u(t)$	⟷	$(\pi/2j)[\delta(\omega - \omega_0) - \delta(\omega + \omega_0)] + [\omega_0^2/(\omega_0^2 - \omega^2)]$
13a.	$[e^{-at}\cos\omega_0 t]\,u(t)$	⟷	$(a + j\omega)/[(a + j\omega)^2 + \omega_0^2]$
13b.	$[\cos\omega_0 t]\,u(t)$	⟷	$(\pi/2)[\delta(\omega - \omega_0) + \delta(\omega + \omega_0)] + [j\omega/(\omega_0^2 - \omega^2)]$

According to Fig. 6-83, for $f_u/B = 1.5$, the sampling rate should be

$$f_s = 3B = 3 \times 100 = 300 \text{ Hz}.$$

Per Eq. (6.158), the spectrum of the sampled signal is given by

$$\mathbf{X}_s(f) = f_s \sum_{n=-\infty}^{\infty} \mathbf{X}(f - nf_s).$$

Spectrum $\mathbf{X}_s(t)$ consists of the original spectrum, plus an infinite number of duplicates, shifted to the right by nf_s for $n > 0$ and to the left by $|nf_s|$ for $n < 0$. All spectra are scaled by a multiplicative factor of f_s. Figure 6-86(b) displays the spectra for $n = 0$ and ± 1. We note that the spectra do not overlap, which means that the original signal can be reconstructed by passing spectrum $\mathbf{X}_s(f)$ through a lowpass filter with a spectral response that extends between -150 Hz and $+150$ Hz.

6-13.14 Practical Aspects of Sampling

So far, all our discussions of signal sampling assumed the availability of an ideal impulse train composed of impulses. In practice, the sampling is performed by finite-duration pulses that may resemble rectangular or Gaussian waveforms. Figure 6-84 depicts the sampling operation for a signal $x(t)$, with the sampling generated by a train of rectangular pulses given by

$$p_{T_s}(t) = \sum_{n=-\infty}^{\infty} \text{rect}\left(\frac{t - nT_s}{\tau}\right),$$

where T_s is the sampling interval and τ is the pulse width. From Table 5-4, the Fourier series representation of the pulse train is

$$p_{T_s}(t) = \frac{\tau}{T_s} + \sum_{m=1}^{\infty} \frac{2}{m\pi} \sin\left(\frac{m\pi\tau}{T_s}\right) \cos\left(\frac{2m\pi t}{T_s}\right). \quad (6.166)$$

The signal sampled by the pulse train is

$$x_s(t) = x(t)\; p_{T_s}(t)$$
$$= \frac{\tau}{T_s}\, x(t) + \sum_{m=1}^{\infty} \frac{2}{m\pi}\, x(t) \sin\left(\frac{m\pi\tau}{T_s}\right) \cos\left(\frac{2m\pi t}{T_s}\right), \quad (6.167)$$

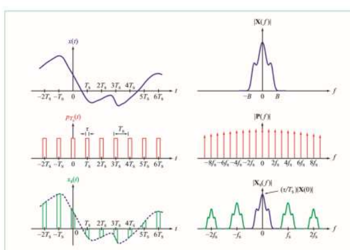


Figure 6-84: Sampling a signal $x(t)$ by a train of rectangular pulses.

which can be cast in the form

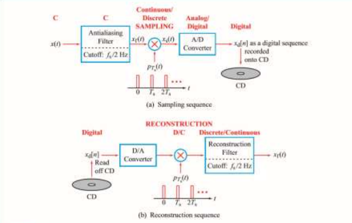
$$x_s(t) = A_0\, x(t) + A_1\, x(t) \cos(2\pi f_s t) + A_2\, x(t) \cos(4\pi f_s t) + \cdots, \quad (6.168)$$

with

$$f_s = \frac{1}{T_s}, \quad A_0 = \frac{\tau}{T_s},$$
$$A_1 = \frac{2}{\pi}\sin(\pi f_s \tau), \quad A_2 = \frac{1}{\pi}\sin(2\pi f_s \tau), \ldots .$$

The sequence given by Eq. (6.168) consists of a dc term, $A_0\, x(t)$, and a sum of sinusoids at frequency f_s and its harmonics. The Fourier transform of the A_0 term is $A_0\, \mathbf{X}(f)$, where $\mathbf{X}(f)$ is the transform of $x(t)$. It is represented by the central spectrum of $\mathbf{X}_s(f)$ in Fig. 6-84. The cosine terms in Eq. (6.168) generate image spectra centered at $\pm f_s$ and its harmonics, but their amplitudes are modified by the values of A_1, A_2, etc.

Signal $x(t)$ can be reconstructed from $x_s(t)$ by lowpass filtering it, just as was done earlier with the ideal impulse sampling. The Shannon sampling requirement that f_s should be greater than $2B$ still holds.



(a) Sampling sequence

(b) Reconstruction sequence

Figure 6-85: DSP system with C/D and D/C to convert between continuous time and discrete time, and A/D and D/A to convert between analog and digital.

By way of an example, we show in Fig. 6-85 a block diagram of a typical DSP system that uses a compact disc (CD) for storage of digital data. The sampling part of the process starts with a continuous-time signal $x(t)$. After bandlimiting the signal to $f_s/2$ by an antialiasing filter, the filtered signal $x_f(t)$ is sampled at a rate f_s and then converted into a digital sequence $x_d[n]$ that gets recorded onto the CD.

Reconstruction performs the reverse sequence, starting with $x_d[n]$ as read off of the CD and concluding in $x_f(t)$, the bandlimited version of the original signal.

Concept Question 6-19: Does sampling a signal at exactly the Nyquist rate guarantee that it can be reconstructed from its discrete samples? (See S[2])

Concept Question 6-20: What is signal aliasing? What causes it? How can it be avoided? (See S[2])

Concept Question 6-21: If brick-wall lowpass filters are used in connection with a signal bandlimited to f_{max} and sampled at f_s, what should the filter's cutoff frequency be when used as (a) an anti-aliasing filter and (b) a reconstruction filter? (See S[2])

Exercise 6-18: What is the Nyquist sampling rate for a signal bandlimited to 5 kHz?

Answer: 10 kHz. (See S[2])

Exercise 6-19: A 500 Hz sinusoid is sampled at 900 Hz. No anti-alias filter is used. What is the frequency of the reconstructed sinusoid?

Answer: 400 Hz. (See S[2])