Physics 127A — Practice Exam Solutions

Spring 2026

Standard idealizations are assumed knowledge and are not printed on the exam: an active-region BJT has $V_{\mathrm{BE}} \approx 0.6\,\mathrm{V}$ and a conducting diode drops $0.6\,\mathrm{V}$; a typical $\beta = 100$ is used where a problem calls for it (Problems 5 and 8 state it explicitly). Thermal voltage $V_T \approx 25\,\mathrm{mV}$ at room temperature, so $r_e = V_T/I_C \approx 25\,\Omega/I_C(\mathrm{mA})$. Where useful I give each number two ways as a consistency check.

Problem 1. Resistive divider, Thevenin equivalent, and power

(a) Unloaded, no current flows except through the series pair, so the divider rule applies:

$$V_{\mathrm{out}} = 12\,\mathrm{V} \cdot \frac{R_2}{R_1 + R_2} = 12 \cdot \frac{2\,\mathrm{k\Omega}}{3\,\mathrm{k\Omega}} = 8\,\mathrm{V}.$$

(b) Killing the ideal source (short it) leaves R_1 and R_2 in parallel as seen from the output node:

$$R_{\mathrm{th}} = R_1 \| R_2 = \frac{1 \cdot 2}{1 + 2}\,\mathrm{k\Omega} = \frac{2}{3}\,\mathrm{k\Omega} \approx 667\,\Omega.$$

(c) Model the divider as its Thevenin source ($V_{\mathrm{th}} = 8\,\mathrm{V}$, $R_{\mathrm{th}} = 667\,\Omega$) feeding $R_L = 1\,\mathrm{k\Omega}$:

$$V_{\mathrm{out}} = 8 \cdot \frac{1000}{1000 + 667} = 4.80\,\mathrm{V}.$$

Check by collapsing the load into the divider directly: $R_2 \| R_L = 1\,\mathrm{k\Omega} \| 2\,\mathrm{k\Omega} = 667\,\Omega$, so $V_{\mathrm{out}} = 12 \cdot 667/(1000 + 667) = 4.80\,\mathrm{V}$. The two routes agree.

(d) With the output at $4.80\,\mathrm{V}$, the current in R_1 is

$$I_{R_1} = \frac{12 - 4.80}{1\,\mathrm{k\Omega}} = 7.2\,\mathrm{mA},$$

so $P_{R_1} = I^2 R_1 = (7.2\,\mathrm{mA})^2 (1\,\mathrm{k\Omega}) = 51.8\,\mathrm{mW}$. Cross-check: $P = I\,V_{R_1} = (7.2\,\mathrm{mA})(7.2\,\mathrm{V}) = 51.8\,\mathrm{mW}$.

Problem 2. Non-ideal meters

(a) The two $1\,\mathrm{M\Omega}$ resistors give an unloaded midpoint of $5\,\mathrm{V}$. The voltmeter's $10\,\mathrm{M\Omega}$ sits in parallel with the lower resistor:

$$1\,\mathrm{M\Omega} \| 10\,\mathrm{M\Omega} = \frac{10}{11}\,\mathrm{M\Omega} = 0.909\,\mathrm{M\Omega}.$$

$$V_{\mathrm{read}} = 10 \cdot \frac{0.909}{1 + 0.909} = 4.76\,\mathrm{V}.$$

Error $= (5 - 4.76)/5 = 4.76\,\%$ low. (Loading a source whose impedance is one-tenth of the meter's costs you a few percent — the usual rule of thumb.)

(b) The ammeter's $1\,\Omega$ adds to the $100\,\Omega$:

$$I_{\text{read}} = \frac{10\,\text{V}}{101\,\Omega} = 99.0\,\text{mA},$$

versus the true $10\,\text{V}/100\,\Omega = 100\,\text{mA}$: about 1 % low.

(c) Setting $|X_C| = 1/(\omega C) = 1\,\text{M}\Omega$:

$$f = \frac{1}{2\pi RC} = \frac{1}{2\pi(1\,\text{M}\Omega)(20\,\text{pF})} = 7.96\,\text{kHz}.$$

Above this frequency the 20 pF reactance falls below the $1\,\text{M}\Omega$ resistance, so the probe loads the circuit more and more heavily and high-frequency signals are attenuated — the scope rolls off the measurement.

Problem 3. Diodes and a Zener regulator

(a) Circuit (i): the diode drops 0.6 V, leaving $5\,\text{V} - 0.6\,\text{V} = 4.4\,\text{V}$ across the $1\,\text{k}\Omega$:

$$I = \frac{4.4}{1\,\text{k}\Omega} = 4.4\,\text{mA}.$$

(b) No load: all the series current flows through the Zener.

$$I = \frac{15 - 5.1}{470\,\Omega} = 21.06\,\text{mA}, \qquad P_Z = (5.1\,\text{V})(21.06\,\text{mA}) = 107\,\text{mW}.$$

(c) A 10 mA load steals current from the Zener:

$$I_Z = 21.06 - 10 = 11.06\,\text{mA} > 0,$$

so the Zener still conducts and the output stays regulated at 5.1 V.

(d) Regulation holds until the Zener current reaches zero, i.e. the load takes the entire series current:

$$I_{L,\max} = 21.06\,\text{mA}.$$

Check: that load looks like $R_L = 5.1/21.06\,\text{mA} = 242\,\Omega$; the divider $15{\cdot}242/(242{+}470) = 5.1\,\text{V}$ sits right at the edge of dropout. Consistent.

Problem 4. RC low-pass filter

(a) Output taken across the capacitor, whose impedance falls with frequency, so this is a *low-pass* filter. The $-3\,\text{dB}$ corner is

$$f_c = \frac{1}{2\pi RC} = \frac{1}{2\pi(1\,\text{k}\Omega)(0.1\,\mu\text{F})} = 1.59\,\text{kHz}.$$

(b) The transfer function is the capacitor's share of the divider:

$$H(\omega) = \frac{1/i\omega C}{R + 1/i\omega C} = \frac{1}{1 + i\omega RC}.$$

At $f = f_c$, $\omega RC = 1$, so $|H| = 1/\sqrt{2} = 0.707$, i.e. $-3.0\,\text{dB}$.

(c) At $f = 10f_c$, $\omega RC = 10$:

$$|H| = \frac{1}{\sqrt{1+100}} = 0.0995 \approx -20.0\,\text{dB},$$

and the phase is $\angle H = -\arctan(10) = -84.3°$. (A single-pole low-pass rolls off at $-20\,\text{dB/decade}$ and approaches $-90°$ — both checks consistent.)

(d) At $f = f_c$ the gain magnitude is 0.707 and the phase is $-\arctan(1) = -45°$. A 2 V-amplitude input gives a $2(0.707) = 1.41$ V-amplitude output, lagging by 45°.

Problem 5. Emitter follower

(a) The base divider ($R_1 = R_2 = 100\,\text{k}\Omega$) sets, to first order, $V_B = 15\,\text{V}/2 = 7.5\,\text{V}$, hence

$$V_E = V_B - V_{\text{BE}} = 6.9\,\text{V}, \qquad I_E = \frac{6.9}{6.8\,\text{k}\Omega} = 1.01\,\text{mA} \approx I_C.$$

Consistency note: the divider current is $\sim 75\,\mu\text{A}$ while the base current is $I_B = I_C/\beta \approx 10\,\mu\text{A}$, a ratio near 7:1. That is close to the usual 10:1 stiffness target but not perfectly stiff; base loading pulls V_B down by roughly $I_B(R_1\|R_2) = (10\,\mu\text{A})(50\,\text{k}\Omega) \approx 0.5\,\text{V}$, to about 7.0 V. The simple model ignores this; either answer is acceptable if the assumption is stated. Using $I_C \approx 1\,\text{mA}$ gives $r_e \approx 25\,\Omega$.

(b) Looking into the base, the emitter resistor is multiplied by $\beta+1$:

$$Z_{\text{in,base}} = \beta(r_e + R_E) = 100(25 + 6800)\,\Omega \approx 683\,\text{k}\Omega.$$

In parallel with the divider:

$$Z_{\text{in}} = R_1\|R_2\|Z_{\text{in,base}} = 50\,\text{k}\Omega\|683\,\text{k}\Omega \approx 46.6\,\text{k}\Omega.$$

The bias network, not the transistor, dominates the input impedance here.

(c) Driven from $R_s = 1\,\text{k}\Omega$, the output impedance at the emitter is the source impedance divided down by β, plus r_e:

$$Z_{\text{out}} = r_e + \frac{R_s\|R_1\|R_2}{\beta} = 25 + \frac{1\,\text{k}\Omega\|50\,\text{k}\Omega}{100} \approx 25 + 9.8 \approx 35\,\Omega$$

(R_E in parallel is negligible).

(d) The follower gain is just below unity:

$$A_v = \frac{R_E}{R_E + r_e} = \frac{6800}{6825} = 0.996 \approx 1.$$

Problem 6. Common-emitter amplifier

(a) With the base pinned at $V_B = 2.6\,\text{V}$, $V_E = 2.6 - 0.6 = 2.0\,\text{V}$, so

$$I_E = \frac{2.0}{2\,\text{k}\Omega} = 1.0\,\text{mA} \approx I_C, \qquad V_C = 15 - (1\,\text{mA})(5\,\text{k}\Omega) = 10\,\text{V}.$$

$V_{CE} = V_C - V_E = 8\,\text{V} \gg 0.2\,\text{V}$: comfortably active. Also $r_e = 25\,\Omega/1 = 25\,\Omega$.

(b) With the emitter fully bypassed at signal frequencies, the gain is set by R_C and r_e:

$$A_v = -\frac{R_C}{r_e} = -\frac{5000}{25} = -200.$$

(c) In decibels, $20\log_{10}(200) = 46.0\,\mathrm{dB}$.

(d) Leaving $100\,\Omega$ unbypassed in the emitter adds emitter degeneration:

$$A_v = -\frac{R_C}{r_e + R'_E} = -\frac{5000}{25+100} = -40.$$

The gain drops fivefold but becomes far less sensitive to r_e (hence to temperature and bias) — the usual reason to degenerate.

Problem 7. Current source and current mirror

(a) Ebers–Moll gives $I_C = I_S(e^{V_{\mathrm{BE}}/V_T} - 1) \approx I_S e^{V_{\mathrm{BE}}/V_T}$. Then

$$r_e \equiv \frac{dV_{BE}}{dI_C} = \left(\frac{dI_C}{dV_{BE}}\right)^{-1} = \frac{V_T}{I_C}.$$

At $I_C = 1\,\mathrm{mA}$ and $V_T = 25\,\mathrm{mV}$, $r_e = 25\,\Omega$.

(b) Circuit (i): the 2.0 V base sets $V_E = 2.0 - 0.6 = 1.4\,\mathrm{V}$, so

$$I_{\mathrm{out}} = I_E = \frac{1.4}{1.4\,\mathrm{k}\Omega} = 1.0\,\mathrm{mA}.$$

The collector current is essentially independent of the load voltage — that is the point of a current source.

(c) The transistor stays in the active region as long as $V_{CE} \geq 0.2\,\mathrm{V}$, i.e. the collector stays at least $V_E + 0.2\,\mathrm{V} = 1.6\,\mathrm{V}$. With a 15 V supply the load can therefore drop anywhere from 0 up to $15 - 1.6 = 13.4\,\mathrm{V}$: a compliance range of about 13.4 V.

(d) For the mirror, the diode-connected Q_1 carries the program current set by R_{prog} from the 15 V rail:

$$R_{\mathrm{prog}} = \frac{15 - V_{\mathrm{BE}}}{I_{\mathrm{prog}}} = \frac{14.4}{1\,\mathrm{mA}} = 14.4\,\mathrm{k}\Omega.$$

Because Q_1 and Q_2 share the same V_{BE} and are matched, Ebers–Moll forces equal collector currents, so $I_{\mathrm{out}} = I_{\mathrm{prog}} = 1\,\mathrm{mA}$. The output current is "mirrored" from the programmed side.

Problem 8. Differential amplifier

(a) The 2 mA tail splits equally when the inputs are balanced: $I_C = 1\,\mathrm{mA}$ per side, $r_e = 25\,\Omega$. Each collector sits at

$$V_C = 15 - (1\,\mathrm{mA})(7.5\,\mathrm{k}\Omega) = 7.5\,\mathrm{V}.$$

(b) For a differential input, each emitter is a virtual signal ground through the other transistor, so the single-ended gain is

$$\frac{v_{o1}}{V_1 - V_2} = \frac{R_C}{2r_e} = \frac{7500}{50} = 150,$$

and the collector-to-collector (double-ended) gain is twice that, $R_C/r_e = 300$.

(c) With a finite tail resistance $R_{\text{tail}} = 500\,\text{k}\Omega$, a common-mode input sees both emitters move together through $2R_{\text{tail}}$:

$$A_{\text{cm}} = -\frac{R_C}{2R_{\text{tail}}} = \frac{7500}{1{,}000{,}000} = 0.0075.$$

Then

$$\text{CMRR} = \left|\frac{A_{\text{diff}}}{A_{\text{cm}}}\right| = \frac{150}{0.0075} = 20000 = 86.0\,\text{dB}.$$

(Cross-check: CMRR $= R_{\text{tail}}/r_e$ in the single-ended form $= 500\text{k}/25 = 20000$. Consistent.)

(d) Looking into one base for a differential signal, $Z_{\text{in}} = 2\beta r_e = 2(100)(25) = 5\,\text{k}\Omega$. The output impedance at one collector is essentially $R_C = 7.5\,\text{k}\Omega$.

Problem 9. Schmitt trigger

(a) Returning the feedback resistor R_2 to the non-inverting (+) input makes any output change reinforce itself: a small rise in V_{out} raises V_+, which (since the signal enters the inverting input) drives the output still higher until it slams to a rail. That regenerative action is positive feedback. The source driving V_{in} connects only to the op amp's inverting input terminal, which (ideal op amp) draws no current, so the input impedance is essentially infinite — the high-impedance feature of this configuration. (Contrast the alternative in which the signal is injected through a resistor into the feedback node, which presents a far lower impedance.)

(b) There is no negative feedback, so $V_- = V_{\text{in}}$ tracks the input directly and the + input sits at a fraction of the output set by the R_1–R_2 divider:

$$V_+ = V_{\text{out}}\,\frac{R_1}{R_1 + R_2} = V_{\text{out}}\,\frac{10}{30} = \frac{V_{\text{out}}}{3}.$$

With the two saturated output states $V_{\text{out}} = \pm 12\,\text{V}$,

$$V_+ = +4\,\text{V (when } V_{\text{out}} = +12), \qquad V_+ = -4\,\text{V (when } V_{\text{out}} = -12).$$

The output switches when V_{in} crosses the current V_+, so the thresholds are

$$V_{\text{th}}^+ = +4\,\text{V}, \qquad V_{\text{th}}^- = -4\,\text{V}.$$

(Concretely: starting with $V_{\text{out}} = +12$, the input must rise past $+4\,\text{V}$ to trip it low; it then stays low until the input falls past $-4\,\text{V}$.)

(c) The hysteresis is the gap between the thresholds,

$$\Delta V = V_{\text{th}}^+ - V_{\text{th}}^- = 4 - (-4) = 8\,\text{V}.$$

(d) The width scales with $R_1/(R_1 + R_2)$, namely $\Delta V = 2V_{\text{sat}}\, R_1/(R_1 + R_2)$. Halving it to 4 V requires $R_1/(R_1 + R_2) = 1/6$, i.e. $R_2 = 5R_1 = 50\,\text{k}\Omega$. Check: $V_+ = \pm 12\,(10/60) = \pm 2\,\text{V}$, giving thresholds ± 2 V and a width of 4 V. Consistent.

Problem 10. Operational amplifiers

(a) Golden rules for an op amp with negative feedback: (1) the output does whatever it must to drive the voltage difference between the inputs to zero ($V_+ = V_-$); (2) the inputs draw no current.

(b) Inverting amplifier: $A_v = -R_f/R_{\text{in}} = -100/10 = -10$. The source sees the input resistor terminating on a virtual ground, so $Z_{\text{in}} = R_{\text{in}} = 10\,\text{k}\Omega$.

(c) Non-inverting amplifier: $A_v = 1 + R_f/R_1 = 1 + 100/10 = 11$. Its input impedance is the op amp's own (ideally infinite), enormously larger than the $10\,\text{k}\Omega$ of the inverting circuit — the main reason to pick the non-inverting topology when you must not load the source.

(d) Replacing R_f with C_f makes an integrator:

$$V_{\text{out}}(t) = -\frac{1}{R_{\text{in}} C_f} \int V_{\text{in}}(t)\, dt.$$

Its gain magnitude is $|A| = 1/(\omega R_{\text{in}} C_f)$, which equals unity when

$$\omega = \frac{1}{R_{\text{in}} C_f} = \frac{1}{(10\,\text{k}\Omega)(0.01\,\mu\text{F})} = 10^4\,\text{rad/s},$$

i.e. $f = \omega/2\pi = 1.59\,\text{kHz}$. (Same number as the Problem 4 corner, since RC is identical — a nice consistency check.)

(e) Slew-rate limit: the maximum sinusoidal slope is $\omega V_p = 2\pi f V_p$. Setting this to the slew rate,

$$f_{\max} = \frac{\text{SR}}{2\pi V_p} = \frac{1\,\text{V}/\mu\text{s}}{2\pi(10\,\text{V})} = 15.9\,\text{kHz}.$$