The Art of Electronics Solutions

July 16, 2023

Contents

Exercise. This is the solution to exercise 1.1 in the book.

Solution. Series

$$R = R_1 + R_2 = 15 \text{ k}\Omega$$

Parallel

$$R = \frac{R_1 R_2}{R_1 + R_2} = 3.33 \text{ k}\Omega$$

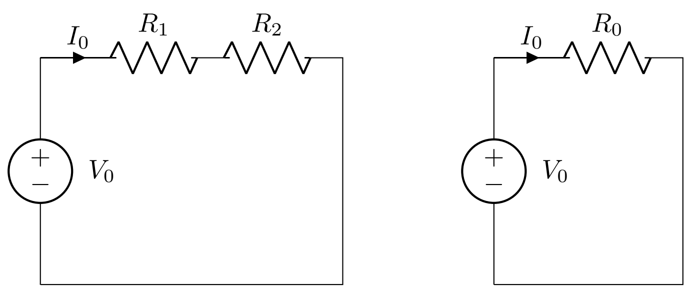
Exercise. This is the solution to exercise 1.2 in the book.

Solution.

$$P = UI = U\frac{U}{R} = \frac{144 \text{ V}^2}{1 \ \Omega} = 144 \text{ W}$$

Exercise. This is the solution to exercise 1.3 in the book.

Solution. For series resistors we have the equivalent circuits:



We have that:

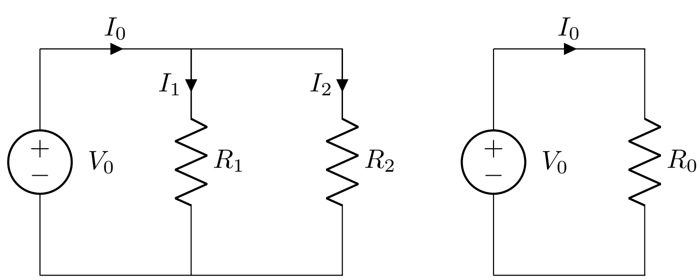
$$\begin{aligned} V_1 &= I_0 R_1 \\ V_2 &= I_0 R_2 \\ V_0 &= V_1 + V_2 \end{aligned}$$

and

$$V_0 = I_0 R_0$$

Therefore:

$$R_0 = R_1 + R_2$$



$$\begin{aligned} V_0 &= I_1 R_1 \\ V_0 &= I_2 R_2 \\ I_0 &= I_1 + I_2 \end{aligned}$$

and

$$V_0 = I_0 R_0$$

hence:

$$\frac{V_0}{R_0} = \frac{V_0}{R_1} + \frac{V_0}{R_2}$$

or

$$\frac{1}{R_0} = \frac{1}{R_1} + \frac{1}{R_2}$$

These can be generalized to more than two resistors by taking two at a time and reducing to one.

Exercise. This is the solution to exercise 1.4 in the book.

Solution. The expression can be re-written as:

$$\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \dots$$

Using the result in exercise 1.3 we can replace R_1 and R_2 with R_{12} to get:

$$\frac{1}{R_{12}} = \frac{1}{R_1} + \frac{1}{R_2}$$

We can then replace R_{12} and R_3 with R_{123}. Thus:

$$\frac{1}{R_{123}} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}$$

Extending this up to more than three resistors we thus get:

$$\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \dots$$

Exercise. This is the solution to exercise 1.5 in the book.

Solution. When the resistor shorts the voltage source the current is:

$$I = \frac{V}{R} = 15 \text{ mA}$$

The power dissipated in the resistor is:

$$P = VI = 15 \times 15 \cdot 10^{-3} \text{ W} = 0.225 \text{ W}$$

Hence, power rating will not be exceeded.

Exercise. This is the solution to exercise 1.6 in the book.

Solution. (a) In order to deliver 10^{10} W of power at 115 V the DC current through the cable must be:

$$I = \frac{P}{U} = \frac{10^{10}}{115} = 86.956 \cdot 10^6 \text{ A}$$

The I^2R losses are then:

$$P_{loss} = (86.956 \cdot 10^6 \text{ A})^2 \cdot 5 \cdot 10^{-8} \frac{\Omega}{\text{ft}} = 378.071 \cdot 10^6 \frac{\text{W}}{\text{ft}}$$

(b) The length is:

$$L = \frac{10^{10}}{378.071 \cdot 10^6} = 26.45 \text{ ft}$$

This is obviously not practical.

(c) From Stefan-Boltzmann law:

$$P = \sigma A T^4$$

where A is the area of the radiating surface. In this case $A = \pi D L$ with $D = 1$ ft and L as determined above. We get:

$$A = 74.785 \cdot 10^3 \text{ cm}^2$$

Using the provided value for σ and the 10^{10} W of power we get:

$$T^4 = 2.23 \cdot 10^{16} \text{ K}^4$$

or

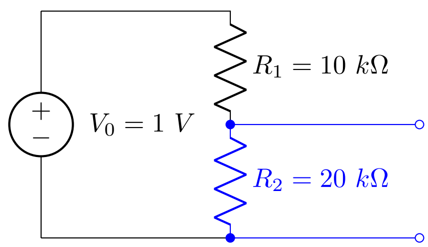
$$T = 12220 \text{ K}$$

which is also not practical.

The solution to this is to transmit power to the city at high voltage. In that case the current is much reduced and $I^2 R$ losses are also much smaller.

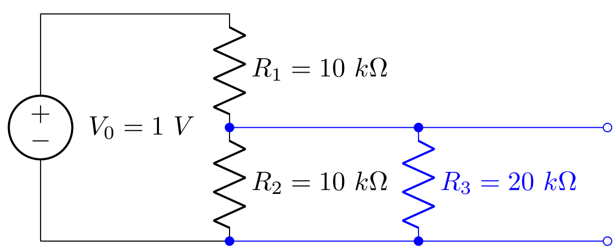
Exercise. This is the solution to exercise 1.7 in the book.

Solution. Remember that the meter will determine the voltage using the current thru the 20kΩ resistor internal to the meter. Hence, when measuring the voltage the following equivalent circuit is exercised, with the part in blue corresponding to the internals of the meter:



We then have:

$$V_{out} = \frac{R_2}{R_1 + R_2} V_0 = \frac{20 \text{ k}\Omega}{30 \text{ k}\Omega} \cdot 1 \text{ V} = 0.67 \text{ V}$$



In this case:

$$R_{23} = \frac{R_2 R_3}{R_2 + R_3} = 6.67 \text{ k}\Omega$$

and

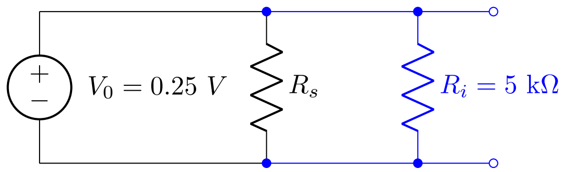
$$V_{out} = \frac{R_{23}}{R_1 + R_{23}} V_0 = \frac{6.67 \text{ k}\Omega}{16.67 \text{ k}\Omega} \cdot 1 \text{ V} = 0.40 \text{ V}$$

Exercise. This is the solution to exercise 1.8 in the book.

Solution. Without the shunt resistance the voltage drop across the meter is:

$$V_m = I_m \cdot R_m = 50\ \mu\text{A} \cdot 5 \text{ k}\Omega = 0.25 \text{ V}$$

To change the scale of the movement from (0,50) μA to (0,1) A, a shunt resistor must be installed between the meter terminals:



The current through the circuit must be 1 A, therefore the equivalent resistance must be:

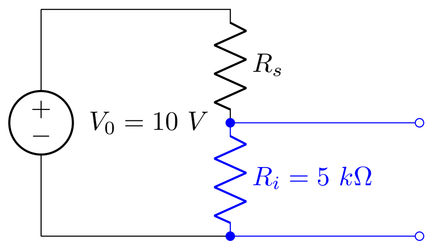
$$R_e = \frac{V_0}{I_0} = \frac{0.25\text{ V}}{1\text{ A}} = 0.25\ \Omega$$

The equivalent resistance is:

$$\frac{1}{R_e} = \frac{1}{R_s} + \frac{1}{R_i}$$
$$\frac{1}{R_s} = \frac{1}{R_e} - \frac{1}{R_i}$$

Therefore the shunt resistance is:

$$R_s = \frac{R_i R_e}{R_i - R_e} = \frac{5000 \cdot 0.25}{5000 - 0.25} = 0.25\ \Omega$$



The current thru the resistors must be the full scale movement current $I_0 = 50\ \mu$A when the voltage is $V_0 = 10$ V. Therefore:

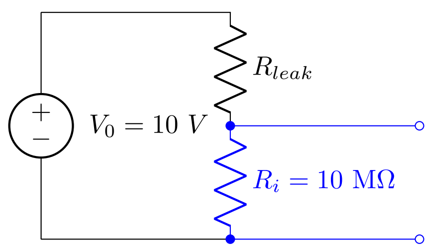
$$R_e = R_i + R_s = \frac{V_0}{I_0} = 200\text{ k}\Omega$$

Therefore the series resistance is:

$$R_i = R_e - R_s = 195\text{ k}\Omega$$

Exercise. This is the solution to exercise 1.9 in the book.

Solution. The measurement can be achieved with the following circuit:



When connecting the 10 V source to the leakage sink the current is:

$$I_{leak} = \frac{10\text{ V}}{(R_{leak} + R_i)\ \Omega} = \frac{10}{R_{leak} + 1 \cdot 10^7}\text{A}$$

The voltage measurement on the meter's 2V scale is in this case:

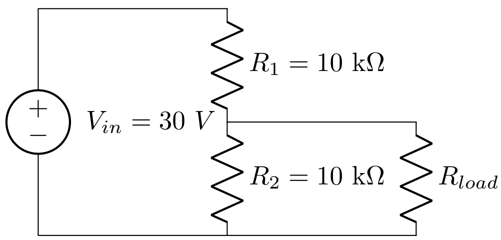
$$V_{leak} = I_{leak} \cdot R_i = \frac{1 \cdot 10^8}{R_{leak} + 1 \cdot 10^7}\text{V}$$

For example if $R_{leak} = 1000$ MΩ then:

$$V_{leak} = \frac{1 \cdot 10^8}{1 \cdot 10^9 + 1 \cdot 10^7} \text{V} = 0.099 \text{ V}$$

Exercise. This is the solution to exercise 1.10 in the book.

Solution.



(a) The open circuit voltage is:

$$V_{open} = \frac{R_2}{R_1 + R_2} V_{in} = 0.5 \cdot 30 \text{ V} = 15 \text{ V}$$

(b) We have the equivalent resistor or R_2 and R_{load} as:

$$R_{out} = \frac{R_2 R_{load}}{R_2 + R_{load}} = \frac{100}{20} \text{ k}\Omega = 5 \text{ k}\Omega$$

$$V_{out} = \frac{R_{out}}{R_1 + R_{out}} V_{in} = \frac{5}{10 + 5} \cdot 30 \text{ V} = 10 \text{ V}$$

(c) The open circuit/Thévenin voltage is:

$$V_{Th} = \frac{R_2}{R_1 + R_2} V_{in}$$

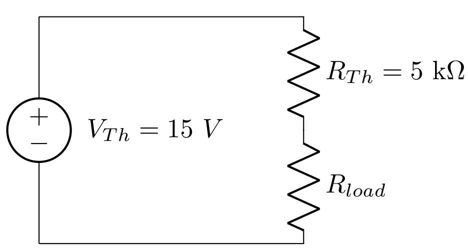
The short-circuit current is:

$$I_{sc} = \frac{V_{in}}{R_1}$$

and the Thévenin resistance is:

$$R_{Th} = \frac{V_{Th}}{I_{sc}} = V_{Th} \frac{R_1}{V_{in}} = \frac{R_1 R_2}{R_1 + R_2} = 5 \text{ k}\Omega$$

Hence the equivalent circuit is:



(d) Using the Thévenin equivalent circuit from (c):

$$V_{out} = \frac{R_{load}}{R_{Th} + R_{load}} V_{Th} = \frac{10}{5 + 10} \cdot 15 \text{ V} = 10 \text{ V}$$

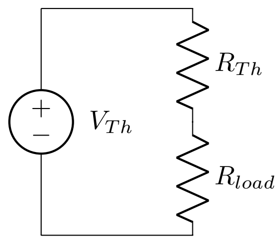
We obtain the same value as in (b).

(e)

$$
\begin{aligned}
P_{load} &= \frac{V_{out}^2}{R_{load}} = \frac{100\ \text{V}^2}{10000\ \Omega} = 0.01\ \text{W} \\
P_1 &= \frac{(V_{in} - V_{out})^2}{R_1} = \frac{400\ \text{V}^2}{10000\ \Omega} = 0.04\ \text{W} \\
P_2 &= \frac{V_{out}^2}{R_2} = \frac{100\ \text{V}^2}{10000\ \Omega} = 0.01\ \text{W}
\end{aligned}
$$

Exercise. This is the solution to exercise 1.11 in the book.

Solution. Remember that the Thévenin equivalent circuit with a load is:



The load voltage is:

$$V_{load} = \frac{R_{load}}{R_{Th} + R_{load}} V_{Th}$$

and the power dissipated by load is:

$$P_{load} = \frac{V_{load}^2}{R_{load}} = \frac{R_{load} V_{Th}^2}{(R_{Th} + R_{load})^2}$$

To maximize power dissipated in the load with respect to R_{load} (and for a given R_{Th}):

$$\frac{\mathrm{d}P_{load}}{\mathrm{d}R_{load}} = 0 = \frac{V_{Th}^2}{(R_{Th} + R_{load})^2} - 2\frac{R_{load} V_{Th}^2}{(R_{Th} + R_{load})^3}$$

or

$$R_{Th} + R_{load} - 2R_{load} = 0$$

or

$$R_{load} = R_{Th}$$

Exercise. This is the solution to exercise 1.12 in the book.

Solution. The power ratio is:

$$\frac{P_1}{P_2} = 10^{\frac{\mathrm{dB}}{10}}$$

and the amplitude (voltage) ratio is:

$$\frac{A_1}{A_2} = 10^{\frac{\mathrm{dB}}{20}}$$

We get:

(a)

$$\frac{P_1}{P_2} = 2 \qquad \frac{A_1}{A_2} = 1.41$$

(b)

$$\frac{P_1}{P_2} = 3.98 \qquad \frac{A_1}{A_2} = 2$$

(c)

$$\frac{P_1}{P_2} = 10 \qquad \frac{A_1}{A_2} = 3.16$$

(d)

$$\frac{P_1}{P_2} = 100 \qquad \frac{A_1}{A_2} = 10$$

Exercise. This is the solution to exercise 1.13 in the book.

Solution. We can use the fact that 3 dB difference corresponds to a doubling (or halving) of power while a 10 dB difference corresponds to increasing/decreasing the power 10×. So we first find the values for 7 dB, 4 dB and 1 dB. Then we estimate the values for 11 dB and then 8 dB, 5 dB and 2 dB.

dB	ratio (P/P_0)
0	1
1	1.25
2	$\frac{\pi}{2}$
3	2
4	2.5
5	π
6	4
7	5
8	6.25
9	8
10	10
11	12.5

Exercise. This is the solution to exercise 1.14 in the book.

Solution. We have that:

$$\mathrm{d}U = VI\mathrm{d}t$$

but

$$I = C\frac{\mathrm{d}V}{\mathrm{d}t}$$

therefore:

$$\mathrm{d}U = CV\mathrm{d}V$$

or

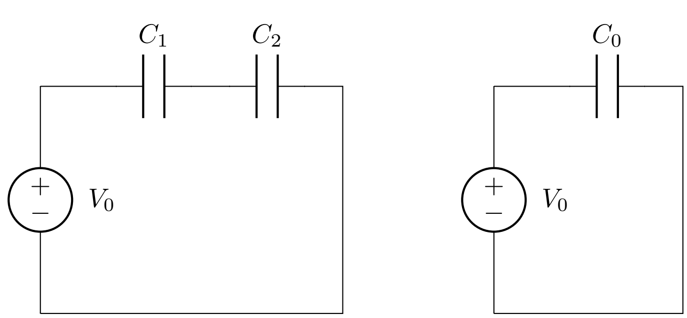
$$\int_0^{U_c} = C\int_0^{V_f} V\mathrm{d}V$$

or

$$U_c = C\frac{V_f^2}{2}$$

Exercise. This is the solution to exercise 1.15 in the book.

Solution.



We have that:

$$\begin{aligned} V_0 &= V_1 + V_2 \\ Q_0 &= C_0 V_0 \\ Q_1 &= C_1 V_1 \\ Q_2 &= C_2 V_2 \end{aligned}$$

Hence:

$$\frac{Q_0}{C_0} = \frac{Q_1}{C_1} + \frac{Q_2}{C_2}$$

Since charge neutrality between C_1 and C_2 must be maintained then $Q_1 = Q_2$. Therefore, the series capacitors have Q_{12} charge on their terminals exposed to V_0. Hence the most obvious choice of an equivalent capacitance is to set $Q_{12} = Q_0$ and thus:

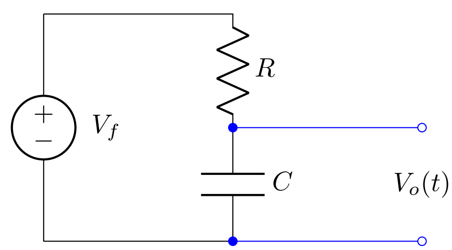
$$\frac{Q_0}{C_0} = Q_{12}\left(\frac{1}{C_1} + \frac{1}{C_2}\right) = Q_0\left(\frac{1}{C_1} + \frac{1}{C_2}\right)$$

and thus:

$$\frac{1}{C_0} = \frac{1}{C_1} + \frac{1}{C_2}$$

Exercise. This is the solution to exercise 1.16 in the book.

Solution.



The current through the resistor and capacitor while the capacitor is charging is:

$$I = C\frac{\mathrm{d}V_o(t)}{\mathrm{d}t} = \frac{V_f - V_o(t)}{R}$$

or

$$C\frac{\mathrm{d}V_o(t)}{\mathrm{d}t} + \frac{V_o(t)}{R} = \frac{V_f}{R}$$

or

$$\frac{\mathrm{d}V_o(t)}{\mathrm{d}t} + \frac{V_o(t)}{RC} = \frac{V_f}{RC}$$

The solution to the homogenous equation is:

$$V_{oh}(t) = A\mathrm{e}^{-\frac{t}{RC}}$$

A particular solution to the non-homogenous equation is:

$$V_{op} = V_f$$

hence:

$$V_o(t) = V_f + A\mathrm{e}^{-\frac{t}{RC}}$$

Using the initial condition:

$$V_o(0) = 0$$

we get:

$$V_o(t) = V_f(1 - \mathrm{e}^{-\frac{t}{RC}})$$

or

$$t = RC \ln \frac{V_f}{V_f - V_o(t)}$$

or

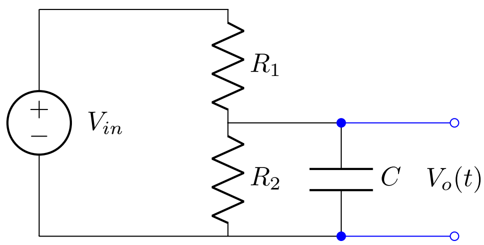
$$\Delta t_{12} = RC \ln \frac{V_f - V_o(t_2)}{V_f - V_o(t_1)}$$

$$\Delta t_{10\% \to 90\%} = RC \ln \frac{V_f - 0.1V_f}{V_f - 0.9V_f}$$

$$\Delta t_{10\% \to 90\%} = 2.2RC$$

Exercise. This is the solution to exercise 1.17 in the book.

Solution.



We have that:

$$\begin{aligned} I_2(t) &= \frac{V_o(t)}{R_2} \\ I_1(t) &= \frac{V_{in} - V_o(t)}{R_1} \\ I_C(t) &= I_1(t) - I_2(t) = C\frac{\mathrm{d}V_o(t)}{\mathrm{d}t} \end{aligned}$$

$$\frac{V_{in} - V_o(t)}{R_1} - \frac{V_o(t)}{R_2} = C\frac{\mathrm{d}V_o(t)}{\mathrm{d}t}$$

$$C\frac{\mathrm{d}V_o(t)}{\mathrm{d}t} + \frac{R_1 + R_2}{R_1 R_2} V_o(t) = \frac{V_{in}}{R_1}$$

The homogenous solution is:

$$V_{oh}(t) = A\mathrm{e}^{-\frac{t(R_1+R_2)}{R_1 R_2 C}}$$

and a particular solution is:

$$V_{op} = V_{in} \frac{R_2}{R_1 + R_2}$$

hence the general solution is:

$$V_o(t) = A\mathrm{e}^{-\frac{t(R_1+R_2)}{R_1 R_2 C}} + V_{in}\frac{R_2}{R_1 + R_2}$$

When $t = 0$ $V_o(t) = 0$ and then:

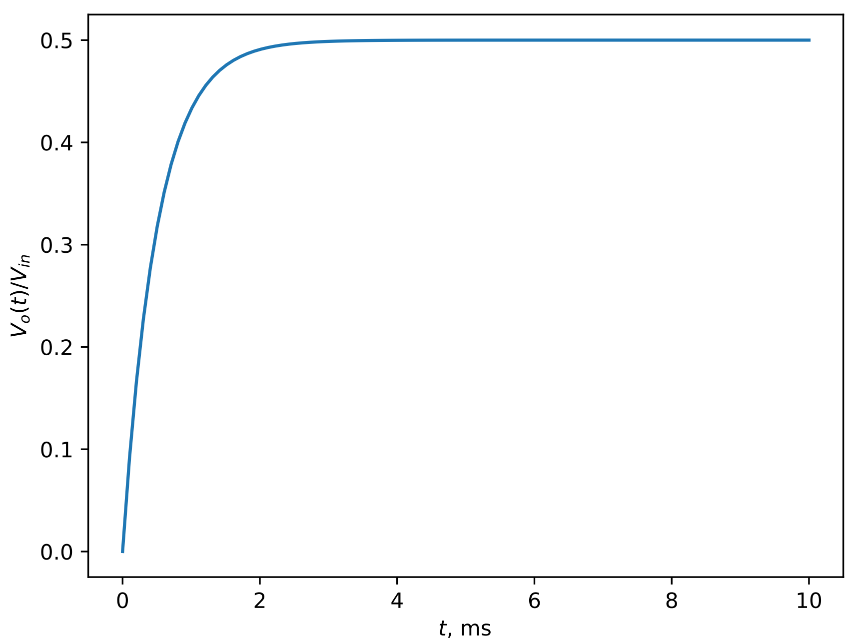
$$A = -V_{in}\frac{R_2}{R_1 + R_2}$$

$$V_o(t) = V_{in}\frac{R_2}{R_1 + R_2}\left[1 - \mathrm{e}^{-\frac{t(R_1+R_2)}{R_1 R_2 C}}\right]$$

Plotting can be done with the following script:

```
import matplotlib.pyplot as plt
import numpy as np

plt.ion()
plt.close('all')

R1=10000
R2=10000
C=1.0E-7

def calc_vr(tin):
    retval = R2/(R1+R2)
    retval*=(1-np.exp(-tin*(R1+R2)/R1/R2/C))
    return retval

t=np.linspace(0,10,100)
vr=np.empty(shape=[100,1])
for i in range(len(t)):
    vr[i] = calc_vr(t[i]/1000.0)

fig,ax=plt.subplots(1,1)
ax.set_xlabel('$t$, ms')
ax.set_ylabel('$V_o(t)/V_{in}$')
ax.plot(t, vr)
#plt.savefig('c01ex17i02.pdf',format='pdf',bbox_inches='tight')
```

to produce the result:



Exercise. This is the solution to exercise 1.18 in the book.

Solution. We have that:

$$I(t) = C\frac{\mathrm{d}V(t)}{\mathrm{d}t}$$

Since the charging current is constant:

$$dV(t) = \frac{I}{C}dt$$

or

$$V_f(t) = \frac{I}{C}t$$

or:

$$t = \frac{1 \cdot 10^{-6}\ \text{F} \cdot 10\ \text{V}}{1 \cdot 10^{-3}\ \text{A}} = 0.01\ \text{s}$$

Exercise. This is the solution to exercise 1.19 in the book.

Solution. Self-inductance of a coil is related to the magnetic flux through one turn of the coil, the number of turns in the coil n and the current through the coil I as:

$$L = \frac{n\Phi_m}{I}$$

The magnetic flux Φ_m through one turn of the coil is:

$$\Phi_m = B \cdot S$$

where B is the magnetic flux density and S is the area of each turn of the coil. The magnetic flux density thru the coil is:

$$B = \mu H = n\mu\frac{I}{l}$$

where μ is the magnetic constant and H is the magnetic field strength produced by the coil with n turns and length l and I is the current through the coil.

Therefore:

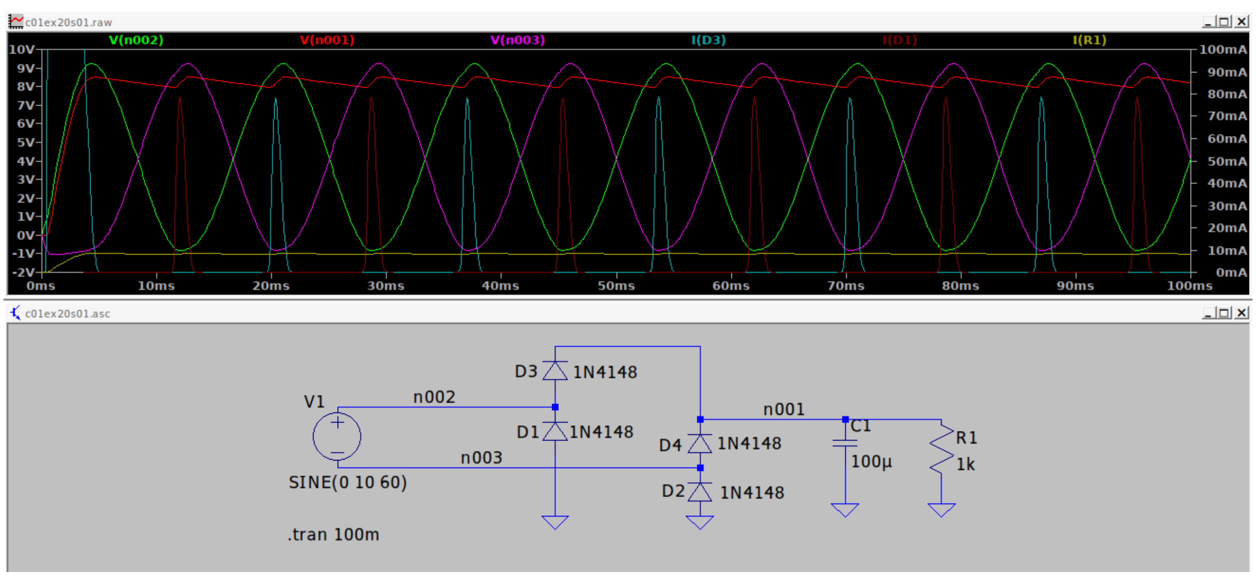
$$L = n^2\mu\frac{S}{l}$$

Notice that l is the total length of the coil, which does not depend on the number of turns (more turns can be accomodated if the surface of each turn is reduced).

Exercise. This is the solution to exercise 1.20 in the book.

Solution. Let's try with a sine wave with amplitude of 10V and 60 Hz frequency.

Please refer to the SPICE simulation below:



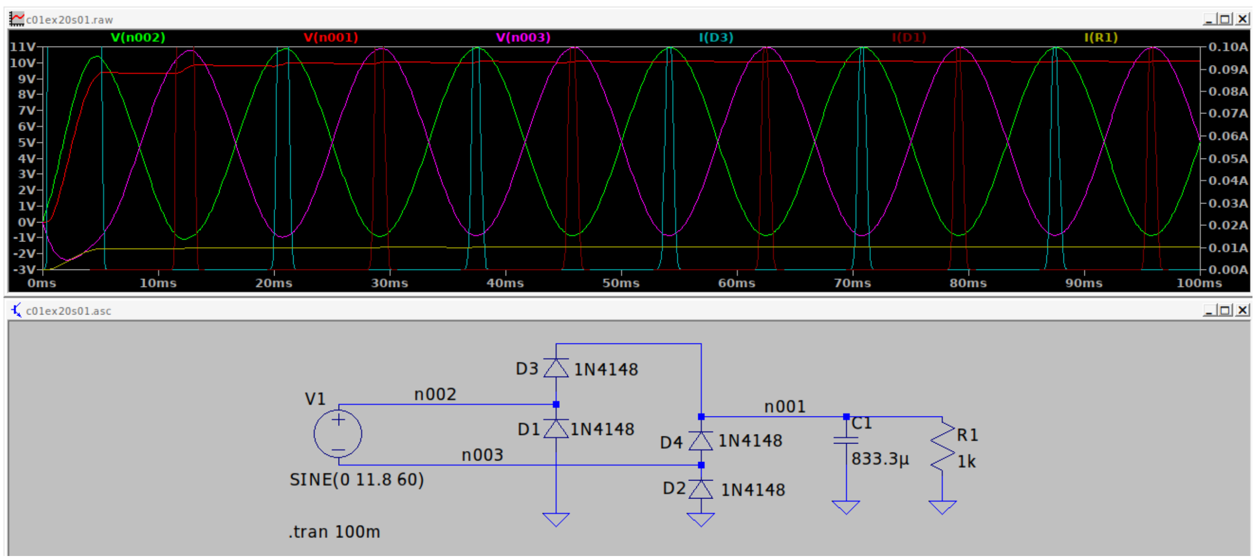
Notice that following the initial transient that charges up the capacitor $V_{dc}(t)$ oscillates between 8.52V and 7.91V. Hence the desired DC voltage has too much ripple and averages around 8.22V. Additionally notice that the peak AC voltage is 9.23 volts. The difference in peak voltage between AC and DC is due to the voltage drop across the forward biased diodes which is 0.7V in this case. Notice that the current thru R1 is indeed approximately 10mA. The DC ripple is expected base on the capacitor we chose:

$$\Delta V = \frac{I_{load}}{2fC} = \frac{8.22 \cdot 10^{-3}}{120 \cdot 100 \cdot 10^{-6}} = 0.68V$$

where we have used the load current of 8.22mA given the average V_{dc}=8.22V. Hence we need the source voltage to be a bit larger to bring the DC voltage to 10V. Additionally the capacitor should be:

$$C = \frac{I_{load}}{2f\Delta V} = \frac{10 \cdot 10^{-3}}{120 \cdot 0.1} = 833\mu F$$

to reduce the ripple in the DC voltage. Hence we increase the amplitude of the AC voltage to 11.8V and change the capacitor accordingly:



In this case the DC voltage ranges from 9.91V to 9.99V hence the ripple is reduced to spec and $V_{dc} \approx$10V. And the load current is 10mA.

Here are the LTSpice schematic and plotting files.

Exercise. This is the solution to exercise 1.21 in the book.

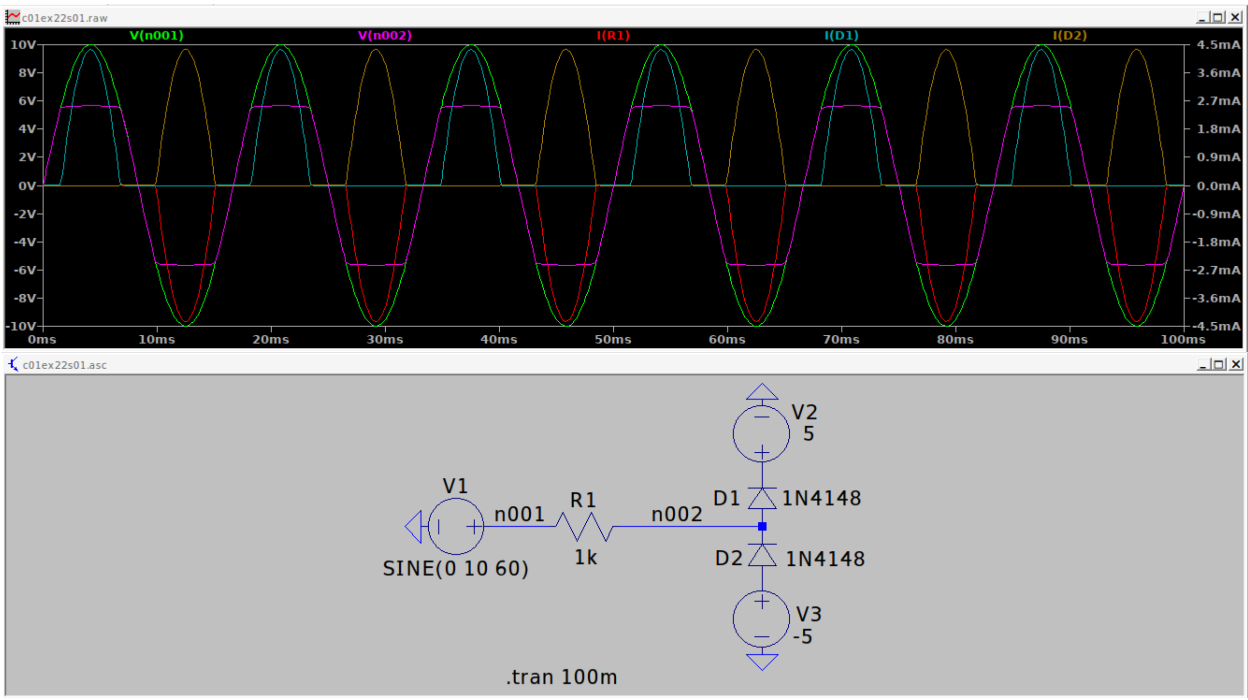
Solution. We use the hint provided to calculate the average I^2 current for the given waveform. If $\tau = \frac{t}{T}$ where T is the period of one cycle. Therefore:

$$\langle I^2 \rangle = \int_0^1 I^2(t) \mathrm{d}\tau = \int_{1/2}^1 4 \mathrm{d}\tau = 2 \text{ A}^2$$

Therefore the fuse must be minimally rated for $I_{fuse} = \sqrt{\langle I^2 \rangle} = 1.41$ A.

Exercise. This is the solution to exercise 1.22 in the book.

Solution. An implementation is provided below:



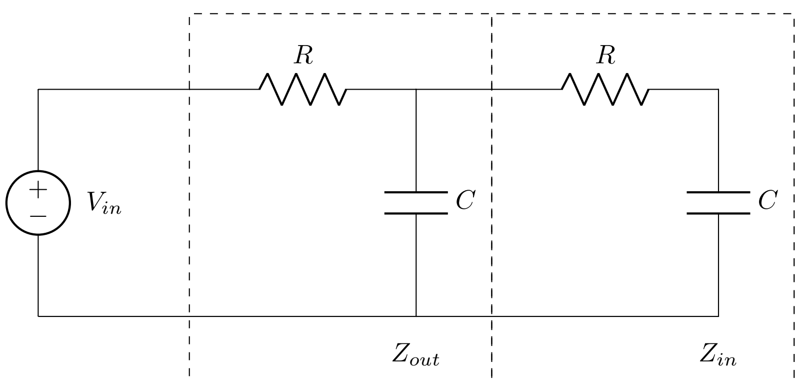
Notice that as the voltage in front of $R1$ ($V(n001)$) becomes greater than 5.6 V diode D1 becomes forward biased and current starts to flow through it from source V_1. There is no current flowing thru D2. Similarly, as the voltage decreases below -5.6 V then diode D2 becomes forward biased and the current flows back into source V_1. There is no current flowing thru D1 in this case.

Hence $V(n002)$ is clamped between -5.6 V and 5.6 V.

Here are the LTSpice schematic and plotting files.

Exercise. This is the solution to exercise 1.23 in the book.

Solution. Let's consider the following circuit. It encompasses an RC low-pass filter as source output impedance and another RC low-pass filter as load input impedance.



We have that:

$$Z_{out} = \frac{RX_C}{R + X_C} = \frac{-\mathrm{i}/\omega/C \cdot R}{R - \mathrm{i}/\omega/C} = \frac{R}{1 + \mathrm{i}\omega RC}$$

$$|Z_{out}| = \frac{R}{\sqrt{1 + \omega^2 R^2 C^2}}$$

Therefore the maximum output impedance is for $\omega = 0$:

$$|Z_{out}| = R$$

We also have that:

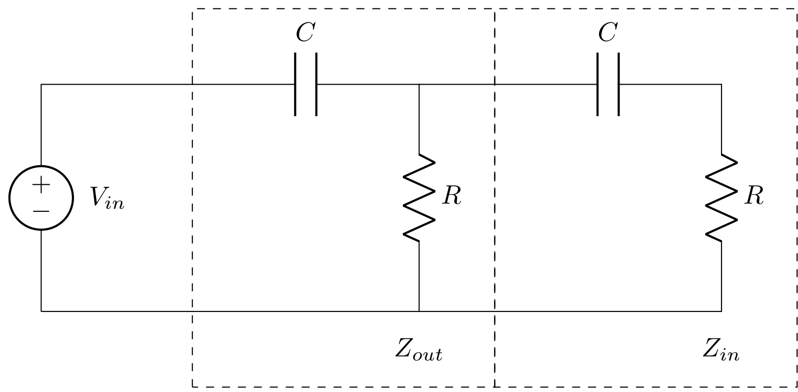
$$Z_{in} = R + X_C = R - \frac{\mathrm{i}}{\omega C}$$

$$|Z_{in}| = \sqrt{R^2 + \frac{1}{\omega^2 C^2}}$$

The minimum input impedance is for $\omega \to \infty$:

$$|Z_{in}| = R$$

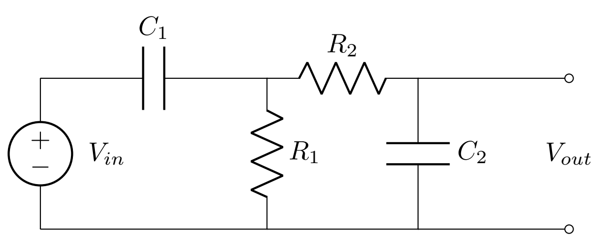
Now consider the case of a high-pass filter:



Notice that impedances are calculated in exactly the same way. Hence, indeed, the worst-case impedance is indeed just R.

Exercise. This is the solution to exercise 1.24 in the book.

Solution. Consider the circuit as follows:



Since the source impedance is 100 Ω then we must choose the minimum input impedance of the high pass RC filter to be 1 kΩ. Hence we choose $R_1 = 1$ kΩ and given the breakpoint of 100 Hz:

$$\omega = 2\pi f = \frac{1}{RC}$$

hence:

$$C_1 = \frac{1}{2\pi f R_1}$$

or:

$$C_1 = 1.59\ \mu\text{F}$$

The worst-case output impedance of the second low-pass filter must be 10 k Ω hence we choose $R_2 = 10$ kΩ. Using:

$$C_2 = \frac{1}{2\pi f R_2}$$

with $f = 10$ kHz we get:

$$C_2 = 1.59\ \text{nF}$$

Since the output impedance of the low-pass filter is $R_2 = 10$ kΩ then the minimum input impedance of the load must be:

$$|Z_{load}| = 10R_2 = 100 \text{ k}\Omega$$

Exercise. This is the solution to exercise 1.25 in the book.

Solution. When the capacitors are in series then impedances add:

$$Z = Z_1 + Z_2$$

But:

$$Z_1 = -\frac{\mathrm{i}}{\omega C_1} \qquad Z_2 = -\frac{\mathrm{i}}{\omega C_2}$$

and

$$Z = -\frac{\mathrm{i}}{\omega C}$$

where C is the capacitance of the capacitor equivalent to the two capacitors in series. Hence:

$$-\frac{\mathrm{i}}{\omega C} = -\frac{\mathrm{i}}{\omega C_1} - \frac{\mathrm{i}}{\omega C_2}$$

or

$$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2}$$

or

$$C = \frac{C_1 C_2}{C_1 + C_2}$$

which is exactly equation 1.18 in the text. When the capacitors are in parallel the impedances are related as:

$$\frac{1}{Z} = \frac{1}{Z_1} + \frac{1}{Z_2}$$

or

$$\mathrm{i}\omega C = \mathrm{i}\omega C_1 + \mathrm{i}\omega C_2$$

or

$$C = C_1 + C_2$$

which is exactly equation 1.17 in the text.

Exercise. This is the solution to exercise 1.26 in the book.

Solution. Following the hint and using:

$$\begin{aligned} \mathbf{A} &= A\mathrm{e}^{\theta_A} \\ \mathbf{B} &= B\mathrm{e}^{\theta_B} \\ \mathbf{C} &= C\mathrm{e}^{\theta_C} \end{aligned}$$

we get:

$$A = BC\mathrm{e}^{\theta_B + \theta_C - \theta_A}$$

Since A, B, C are real numbers we must have that $\mathrm{e}^{\theta_B + \theta_C - \theta_A}$ is also a real number. Hence $\mathrm{e}^{\theta_B + \theta_C - \theta_A} = 1$ and:

$$A = BC$$

Exercise. This is the solution to exercise 1.27 in the book.

Solution. We can compute the power over one period as:

$$P = \frac{1}{T}\int_0^T V(t)I(t)\mathrm{d}t$$

Since we can have:

$$V(t) = V_0 \sin 2\pi f t$$

and

$$I(t) = I_0 \cos 2\pi f t$$

such that voltage and current are 90° out of phase then:

$$P = \frac{V_0 I_0}{T} \int_0^T \sin 2\pi f t \cos 2\pi f t \mathrm{d}t = \frac{V_0 I_0}{2T} \int_0^T \sin 4\pi f t \mathrm{d}t$$

or

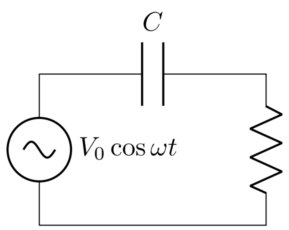
$$P = \frac{V_0 I_0}{8\pi f T} \cos 4\pi f t|_T^0 = \frac{V_0 I_0}{8\pi f T}(1 - \cos 4\pi f T)$$

But $fT = 1$ hence:

$$P = \frac{V_0 I_0}{8\pi}(1 - \cos 4\pi) = 0$$

Exercise. This is the solution to exercise 1.28 in the book.

Solution.



Consider that $I(t)$ is the current through the circuit. We have then:

$$I(t) = C\frac{\mathrm{d}V_C(t)}{\mathrm{d}t} \qquad V_C(t) = V(t) - RI(t)$$

or

$$I(t) = C\frac{\mathrm{d}V(t)}{\mathrm{d}t} - RC\frac{\mathrm{d}I(t)}{\mathrm{d}t}$$

Using the expression:

$$V(t) = V_0 \cos \omega t$$

$$\frac{\mathrm{d}I(t)}{\mathrm{d}t} + \frac{1}{RC}I(t) = -\frac{\omega V_0 \sin \omega t}{R}$$

The differential equation can be solved and using $I(0) = 0$ we get:

$$I(t) = \frac{V_0 \omega C(\omega RC \cos \omega t - \sin \omega t)}{1 + \omega^2 R^2 C^2} - \frac{V_0 \omega^2 RC^2}{1 + \omega^2 R^2 C^2}\mathrm{e}^{-\frac{t}{RC}}$$

Since we are interested in the steady state current when the RC decays:

$$I(t) = \frac{V_0 \omega C(\omega RC \cos \omega t - \sin \omega t)}{1 + \omega^2 R^2 C^2}$$

The power dissipated in the resistor is:

$$P_R(t) = I(t)^2 R = \frac{V_0^2 \omega^2 RC^2(\omega RC \cos \omega t - \sin \omega t)^2}{(1 + \omega^2 R^2 C^2)^2}$$

The average power dissipated over one period is:

$$\bar{P}_R = \frac{1}{T}\int_0^T P_R(t)\mathrm{d}t = \frac{V_0^2 \omega^2 RC^2}{T(1 + \omega^2 R^2 C^2)^2}\int_0^T (\omega RC \cos \omega t - \sin \omega t)^2 \mathrm{d}t$$

or

$$\bar{P}_R = \frac{V_0^2 \omega^2 R C^2}{(1+\omega^2 R^2 C^2)^2} \int_0^1 (\omega RC \cos 2\pi\tau - \sin 2\pi\tau)^2 \mathrm{d}\tau$$

The integral can be evaluated with Mathematica for example to yield:

$$\bar{P}_R = \frac{V_0^2 \omega^2 R C^2}{(1+\omega^2 R^2 C^2)^2} \frac{1+\omega^2 R^2 C^2}{2}$$

or

$$\bar{P}_R = \frac{V_0^2 \omega^2 R C^2}{2(1+\omega^2 R^2 C^2)}$$

The instantaneous power produced by the source (after the decay of the RC transient) is:

$$P_S(t) = V(t)I(t) = \frac{V_0^2 \omega C}{1+\omega^2 R^2 C^2} (\omega RC \cos^2 \omega t - \sin \omega t \cos \omega t)$$

The average power produced over one period is:

$$\bar{P}_S = \frac{1}{T} \int_0^T P_S(t) \mathrm{d}t = \frac{V_0^2 \omega C}{T(1+\omega^2 R^2 C^2)} \int_0^T (\omega RC \cos^2 \omega t - \sin \omega t \cos \omega t) \mathrm{d}t$$

or

$$\bar{P}_S = \frac{V_0^2 \omega C}{(1+\omega^2 R^2 C^2)} \int_0^1 (\omega RC \cos^2 2\pi\tau - \sin 2\pi\tau \cos 2\pi\tau) \mathrm{d}\tau$$

The integral can be evaluated with Mathematica for example to yield:

$$\bar{P}_S = \frac{V_0^2 \omega C}{(1+\omega^2 R^2 C^2)} \frac{\omega RC}{2}$$

or

$$\bar{P}_S = \frac{V_0^2 \omega^2 R C^2}{2(1+\omega^2 R^2 C^2)}$$

Therefore,

$$\bar{P}_S = \bar{P}_R$$

that is all average power produced by the source is dissipated by the resistor.

For $V_0 = 115\sqrt{2}$ V, $\omega = 120\pi$ s^{-1}, $C = 1 \cdot 10^{-6}$ F, and $R = 1000\ \Omega$.

$$P = \frac{115^2 \cdot (120\pi)^2 \cdot 10^{-12} \cdot 1000}{1 + (120\pi)^2 \cdot 10^6 \cdot 10^{-12}} \text{ W} = 1.646 \text{ W}$$

Exercise. This is the solution to exercise 1.29 in the book.

Solution. For an RLC circuit in series the impedance is:

$$Z = R + X_L + X_C$$

$$Z = R + \mathrm{i}\omega L - \frac{\mathrm{i}}{\omega C} = R + \mathrm{i}\omega L - \frac{\mathrm{i}\omega^2 L}{\omega} = R$$

hence the impedance of the circuit is real and the reactance component is zero. Thus the power factor is 1 since impedance is purely resistive.

For a parallel RLC circuit:

$$\frac{1}{Z} = \frac{1}{R} + \frac{1}{X_L} + \frac{1}{X_C}$$

$$Z = \frac{1}{R} - \frac{\mathrm{i}}{\omega L} + \mathrm{i}\omega C = \frac{1}{R} - \frac{\mathrm{i}}{\omega L} + \frac{\mathrm{i}\omega}{\omega^2 L} = R$$

hence the impedance of the circuit is real and the reactance component is zero. Thus the power factor is 1 since impedance is purely resistive.

Exercise. This is the solution to exercise 1.30 in the book.

Solution. We have that:

$$\frac{V_{out}}{V_{in}} = \frac{X_C}{R + X_C} = \frac{-\frac{\mathrm{i}}{\omega C}}{R - \frac{\mathrm{i}}{\omega C}} = \frac{1}{1 + \mathrm{i}\omega RC}$$

or

$$V_{out} = \frac{1 - \mathrm{i}\omega RC}{1 + \omega^2 R^2 C^2} V_{in}$$

Hence:

$$|V_{out}|^2 = \frac{1 + \omega^2 R^2 C^2}{(1 + \omega^2 R^2 C^2)^2} |V_{in}|^2$$

or

$$|V_{out}| = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}} |V_{in}|$$

Exercise. This is the solution to exercise 1.31 in the book.

Solution. For lowpass filter we have that:

$$\mathbf{V}_{out} = \mathbf{V}_{in} \frac{-\frac{\mathrm{i}}{\omega C}}{R - \frac{\mathrm{i}}{\omega C}}$$

or

$$\mathbf{V}_{out} = \mathbf{V}_{in} \frac{1}{1 + \mathrm{i}\omega RC}$$

or

$$\mathbf{V}_{out} = \mathbf{V}_{in} \frac{1 - \mathrm{i}\omega RC}{1 + \omega^2 R^2 C^2}$$

Therefore the phase angle is:

$$\phi = -\arctan \frac{\frac{\omega RC}{1+\omega^2 R^2 C^2}}{\frac{1}{1+\omega^2 R^2 C^2}} = -\arctan \omega RC$$

or

$$\phi = -\arctan(2\pi f RC) = -\arctan \frac{f}{f_{3\mathrm{dB}}}$$

We therefore get:

$$\phi_{0.1\times} = -5.71^\circ \qquad \phi_{10\times} = -84.29^\circ$$

Therefore $\phi_{0.1\times}$ is about 6° from the value $\phi_{0\times} = 0^\circ$ and $\phi_{10\times}$ is about 6° from the asymptotic value $\phi_{\infty\times} = -90^\circ$.

Exercise. This is the solution to exercise 1.32 in the book.

Solution. We have that:

$$\begin{aligned} \mathbf{V}_{out} &= \mathbf{I}R \\ \mathbf{V}_{in} &= \mathbf{I}\left(R - \frac{\mathrm{i}}{\omega C}\right) \end{aligned}$$

or

$$\mathbf{V_{out}} = \mathbf{V}_{in} \frac{R}{R - \frac{\mathrm{i}}{\omega C}}$$

Therefore the magnitude is:

$$|\mathbf{V_{out}}| = |\mathbf{V}_{in}| \frac{R}{\sqrt{R^2 + \frac{1}{\omega^2 C^2}}}$$

Exercise. This is the solution to exercise 1.33 in the book.

Solution.

$$\begin{aligned}\mathbf{V}_{out} &= -\mathbf{I}\frac{\mathrm{i}}{\omega C}\\ \mathbf{V}_{in} &= \mathbf{I}\left(R - \frac{\mathrm{i}}{\omega C}\right)\end{aligned}$$

or

$$\mathbf{V_{out}} = \mathbf{V}_{in}\frac{-\frac{\mathrm{i}}{\omega C}}{R - \frac{\mathrm{i}}{\omega C}}$$

or

$$\mathbf{V_{out}} = \mathbf{V}_{in}\frac{1}{1 + \mathrm{i}\omega RC}$$

Therefore the magnitude is:

$$|\mathbf{V_{out}}| = |\mathbf{V}_{in}|\frac{1}{\sqrt{1+\omega^2R^2C^2}}$$

For 6 dB the ratio must be:

$$\frac{|\mathbf{V_{out}}|}{|\mathbf{V}_{in}|} = \frac{1}{2} = \frac{1}{\sqrt{1+\omega^2R^2C^2}}$$

$$\omega^2R^2C^2 = 3$$

or

$$\omega = \frac{\sqrt{3}}{RC}$$

or

$$f_{6\mathrm{dB}} = \frac{\sqrt{3}}{2\pi RC} = \sqrt{3}f_{3\mathrm{dB}}$$

Since for lowpass RC filter we determined in exercise 1.31:

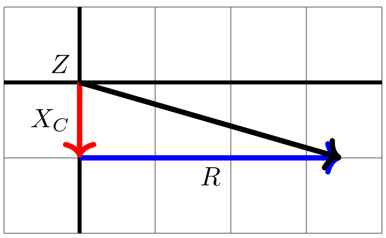
$$\phi_{6\mathrm{dB}} = -\arctan\frac{f_{6\mathrm{dB}}}{f_{3\mathrm{dB}}} = -\arctan\sqrt{3}$$

Thus the phase shift is:

$$\phi_{6\mathrm{dB}} = -60^\circ$$

Exercise. This is the solution to exercise 1.34 in the book.

Solution.



The phase shift between V_{out} and V_{in} is the angle between phasor X_C and the resultant phasor $X_C + R$. The filter response V_{out}/V_{in} is the ratio of the two phasors:

$$\frac{\mathbf{V_{out}}}{\mathbf{V_{in}}} = \frac{X_C}{X_C + R} = \frac{-\frac{\mathrm{i}}{\omega C}}{R - \frac{\mathrm{i}}{\omega C}}$$

or taking the magnitude:

$$|\mathbf{V_{out}}| = |\mathbf{V_{in}}|\frac{\frac{1}{\omega C}}{\sqrt{R^2 + \frac{1}{\omega^2 C^2}}}$$

or

$$|\mathbf{V_{out}}| = |\mathbf{V_{in}}|\frac{1}{\sqrt{1+\omega^2 C^2 R^2}}$$

which is exactly the expression of equation 1.36 in the text.

Exercise. This is the solution to exercise 1.35 in the book.

Solution. We have that:

$$\frac{\mathbf{V_{out}}}{\mathbf{V_{in}}} = \frac{X_C + X_L}{R + X_C + X_L} = \frac{\mathrm{i}\omega L - \frac{\mathrm{i}}{\omega C}}{R + \mathrm{i}\left(\omega L - \frac{1}{\omega C}\right)}$$

Hence, taking the magnitude:

$$|\mathbf{V_{out}}| = |\mathbf{V_{in}}|\frac{\omega L - \frac{1}{\omega C}}{\sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}}$$

or

$$|\mathbf{V_{out}}| = |\mathbf{V_{in}}|\frac{\omega^2 LC - 1}{\sqrt{\omega^2 R^2 C^2 + \left(\omega^2 LC - 1\right)^2}}$$

Therefore when $\omega = \omega_0 = \frac{1}{\sqrt{LC}}$ then:

$$\frac{|\mathbf{V_{out}}|}{|\mathbf{V_{in}}|} = 0$$

When $\omega^2 LC \ll 1$ then:

$$\frac{|\mathbf{V_{out}}|}{|\mathbf{V_{in}}|} = \frac{1}{\sqrt{1+\omega^2 R^2 C^2}}$$

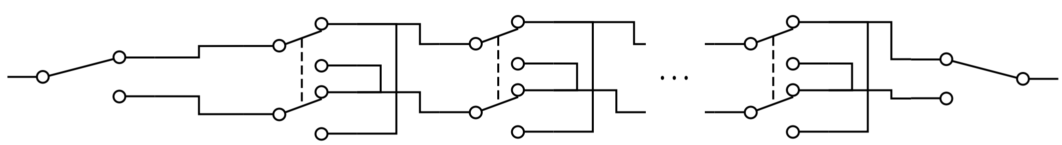
which is the response of an RC circuit. When $\omega^2 LC \gg 1$ then:

$$\frac{|\mathbf{V_{out}}|}{|\mathbf{V_{in}}|} = \frac{\omega L}{\sqrt{R^2 + \omega^2 L^2}}$$

which is the response of the RL circuit.

Exercise. This is the solution to exercise 1.36 in the book.

Solution. We can replace the assembly of two SPST switches with the assembly of 2 SPST and $N-2$ DPDT switches:



Exercise. This is the solution to exercise 1.37 in the book.

Solution. First we place a short across the load (opposite to calculating the Thévenin equivalent voltage where load is open). The Norton current is:

$$I_N = \frac{10\ \mathrm{V}}{10\ \mathrm{k}\Omega} = 1\ \mathrm{mA}$$

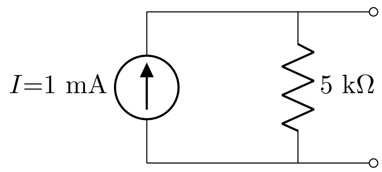
Second find the equivalent Norton resistance by replacing the voltage source with a short. The resistance is:

$$\frac{1}{R_N} = \frac{1}{10\ \mathrm{k}\Omega} + \frac{1}{10\ \mathrm{k}\Omega} = \frac{1}{5\ \mathrm{k}\Omega}$$

therefore:

$$R_N = 5 \text{ k}\Omega$$

Hence the Norton equivalent is:



When the load is 5 kΩ the equivalent resistance is:

$$\frac{1}{R_{eq}} = \frac{1}{5 \text{ k}\Omega} + \frac{1}{5 \text{ k}\Omega}$$

or

$$R_{eq} = 2.5 \text{ k}\Omega$$

and the voltage is $V_{out} = IR_{eq} = 2.5V$.

If the load of $R_{load} = 5$ kΩ is used in the original circuit the voltage across the load is:

$$V_{out} = V_{in}\frac{R_{eq}}{10 \text{ k}\Omega + R_{eq}}$$

where

$$\frac{1}{R_{eq}} = \frac{1}{10 \text{ k}\Omega} + \frac{1}{5 \text{ k}\Omega}$$

or

$$R_{eq} = \frac{10}{3} \text{ k}\Omega$$

Hence:

$$V_{out} = 10 \text{ V} \cdot \frac{\frac{10}{3}}{10 + \frac{10}{3}} = 2.5 \text{ V}$$

Hence the same output value is obtained as that found above from the Norton equivalent circuit.

Exercise. This is the solution to exercise 1.38 in the book.

Solution. For exercise 1.37 the Thévenin equivalent is:

$$V_{Th} = V_{in}\frac{R_1}{R_1 + R_2} = 10 \text{ V}\frac{10 \text{ k}\Omega}{10 \text{ k}\Omega + 10 \text{ k}\Omega} = 5 \text{ V}$$

and

$$R_{Th} = \frac{R_1 R_2}{R_1 + R_2} = \frac{10 \text{ k}\Omega \cdot 10 \text{ k}\Omega}{10 \text{ k}\Omega + 10 \text{ k}\Omega} = 5 \text{ k}\Omega$$

For this exercise the open circuit voltage leads to:

$$V_{Th} = IR_2 = 0.5 \text{ mA} \cdot 10 \text{ k}\Omega = 5 \text{ V}$$

The short-circuit resistance is:

$$R_{Th} = \frac{V_{Th}}{I} = \frac{5}{0.5} = 10 \text{ k}\Omega$$

Hence the two circuits are not equivalent, meaning that even though the open circuit $V_+ = 10V$ we cannot replace the current source with a voltage source of 10 V.

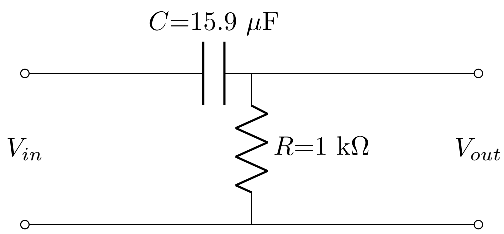
Exercise. This is the solution to exercise 1.39 in the book.

Solution. This is a high-pass RC filter. If $f_{3\text{ dB}} = 10$ Hz then:

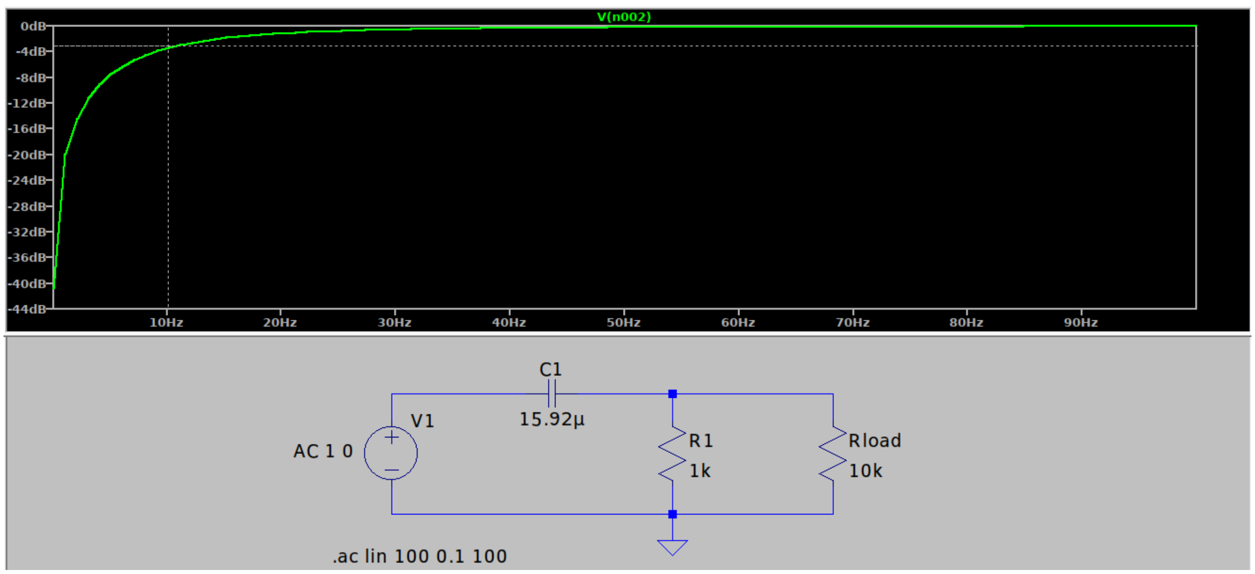
$$RC = \frac{1}{2\pi f_{3\text{ dB}}} = 0.015915 \text{ s}$$

Since the minimum load impedance is 10 kΩ then the output impedance of the filter must be at most 1 kΩ (following the 1:10 source to load impedance ratio). We thus choose $R = 1$ kΩ and:

$$C = 15.92\ \mu\text{F}$$



We can simulate the circuit with the schematic from here and the plot here:



Exercise. This is the solution to exercise 1.40 in the book.

Solution. The scratch filter is a low-pass filter. Again for $f_{3\text{ dB}} = 10$ kHz we get:

$$RC = \frac{1}{2\pi f_{3\text{ dB}}} = 15.92\ \mu\text{s}$$

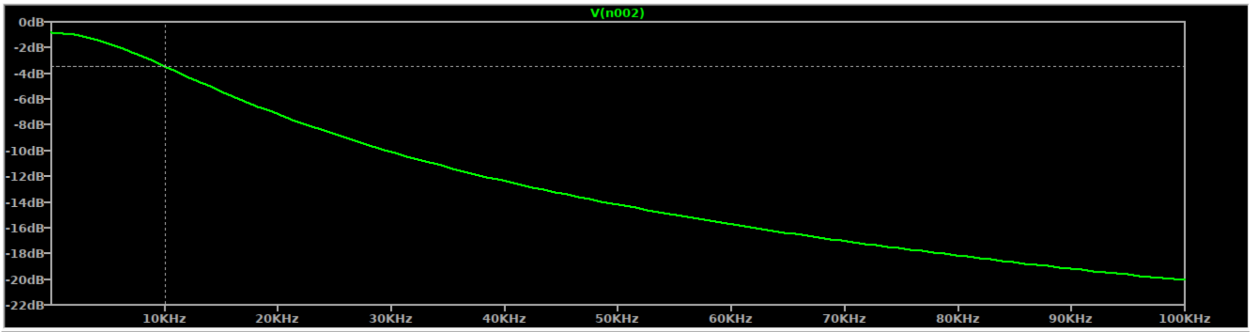
Using a 10 kΩ minimum load impedance and using the 1:10 rule for source to load impedance we set the capacitor impedance at 10 kHz to 1 kΩ and hance:

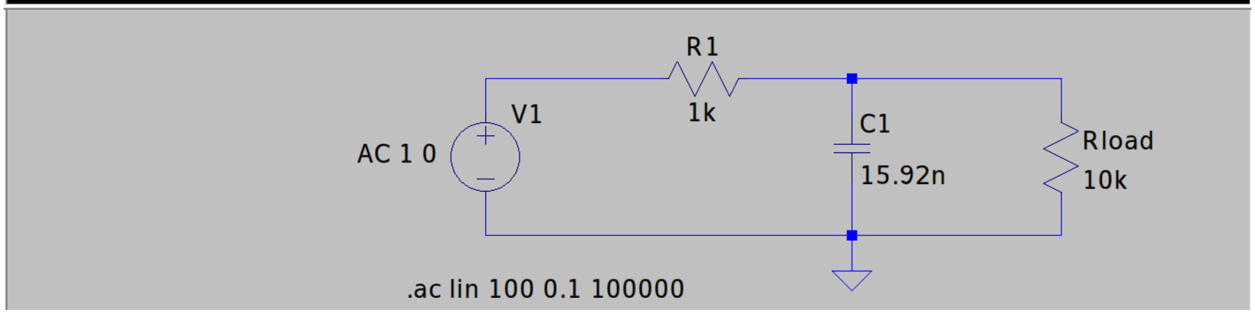
$$C = \frac{1}{2\pi \cdot 10 \text{ kHz} \cdot 1 \text{ k}\Omega} = 15.92 \text{ nF}$$

Resistor value is then:

$$R = 1 \text{ k}\Omega$$

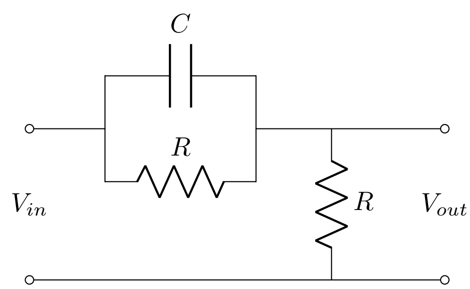
We can simulate the circuit with the schematic from here and the plot here:





Exercise. This is the solution to exercise 1.41 in the book.

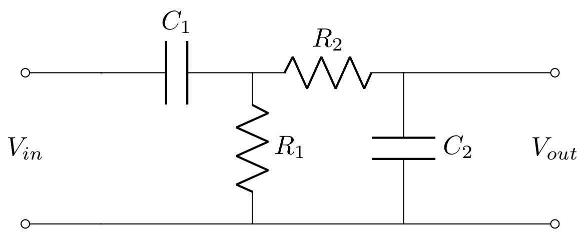
Solution. At low frequencies the circuit behaves like a voltage divider with $R_1 = R_2$ such that $V_{out} = \frac{1}{2}V_{in}$. At high frequencies the circuit behaves as if the voltage source is perfect with zero source resistance. Hence a possible circuit is:



The 3 dB frequency $\omega_{3\text{ dB}}$ is for $\omega_{3\text{ dB}} = \frac{1}{RC} = \omega_0$. Hence once a value for R was chosen, the capacitor value is $C = \frac{1}{\omega_0 R}$.

Exercise. This is the solution to exercise 1.42 in the book.

Solution. According to the numbering the high-pass filter is the first stage and the low-pass filter is the second stage:



The requirement for a 1:10 ratio of source to load impedance implies that $R_2 = 10R_1$. We have that:

$$\omega_1 = \frac{1}{R_1 C_1} \qquad \omega_2 = \frac{1}{R_2 C_2}$$

R_2 value is selected based on the impedance of the load downstream of the bandpass filter. Then C_2 is selected as:

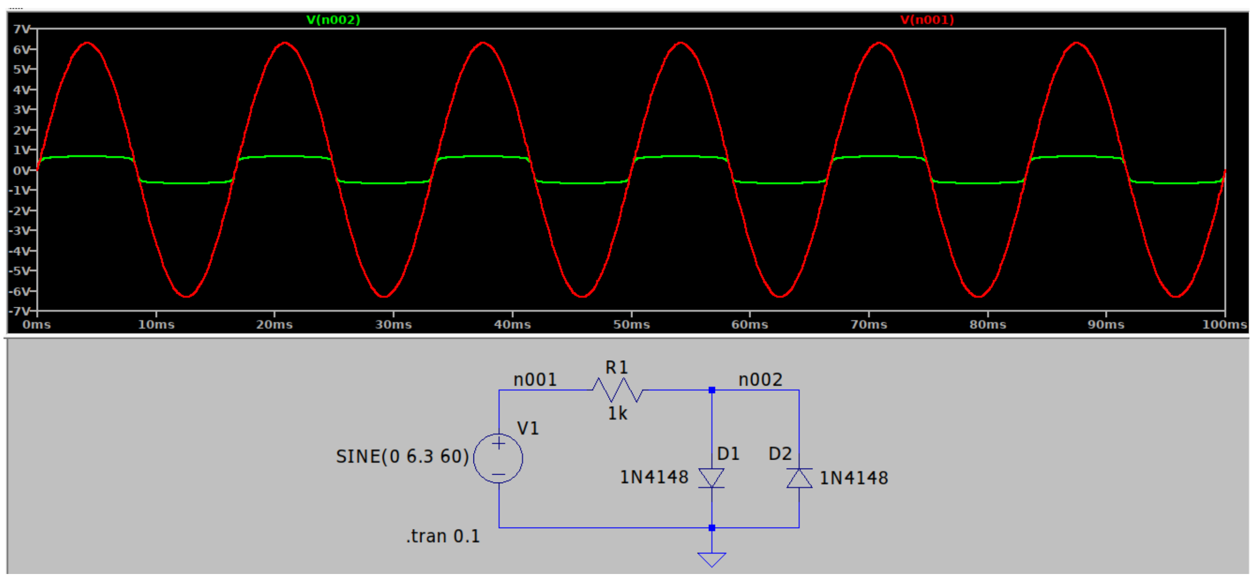
$$C_2 = \frac{1}{\omega_2 R_2}$$

and then we select C_1 as:

$$C_1 = \frac{1}{\omega_1 R_1} = \frac{10}{\omega_1 R_2}$$

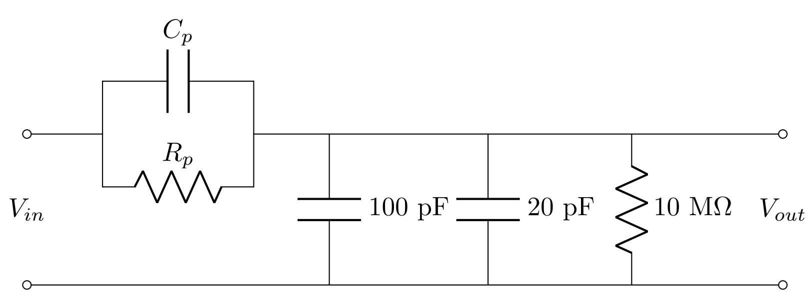
Exercise. This is the solution to exercise 1.43 in the book.

Solution. The circuit is just a voltage clamping circuit. It clamps the output V_{out} between -0.6V and +0.6V. We can simulate the circuit with the schematic from here and the plot here:



Exercise. This is the solution to exercise 1.44 in the book.

Solution. The circuit is:



The parallel capacitors are equivalent to another capacitor with $C_s = 120$ pF. Hence the impedance at the scope input is:

$$Z_s = R_s - \frac{\mathrm{i}}{\omega C_s} = 1\ \mathrm{M}\Omega - \frac{\mathrm{i}}{\omega 120\ \mathrm{pF}}$$

The probe parallel RC results in impedance:

$$Z_p = R_p - \frac{\mathrm{i}}{\omega C_p}$$

Therefore the input-output relationship can be written:

$$\frac{V_{out}}{V_{in}} = \frac{Z_s}{Z_s + Z_p} = \frac{1}{10}$$

for 20 dB attenuation. Thus:

$$Z_p = 9Z_s$$

and therefore:

$$\begin{aligned} R_p &= 9R_s \\ C_p &= \frac{C_s}{9} \end{aligned}$$

or

$$R_p = 9 \text{ M}\Omega \qquad C_p = 13.33 \text{ pF}$$

Exercise. This is the solution to exercise 2.1 in the book.

Solution. If V_{led} is the voltage drop across the LED then the LED current must be (with the given collector resistor of 330 Ω):

$$I_{led} = 10 - 3V \text{ mA}$$

The current should be much less than 10 mA so that the LED is not too bright but much larger than1-2 mA so that the LED is not too faint. For this constraint to hold then the LED must be a red, amber or green (GaP) led. If V=1.7 V then $I_{led} = I_C = 5$ mA.

Assuming a B to E voltage drop of a typical diode of 0.6 V then $V_B = 0.6$ V and thus:

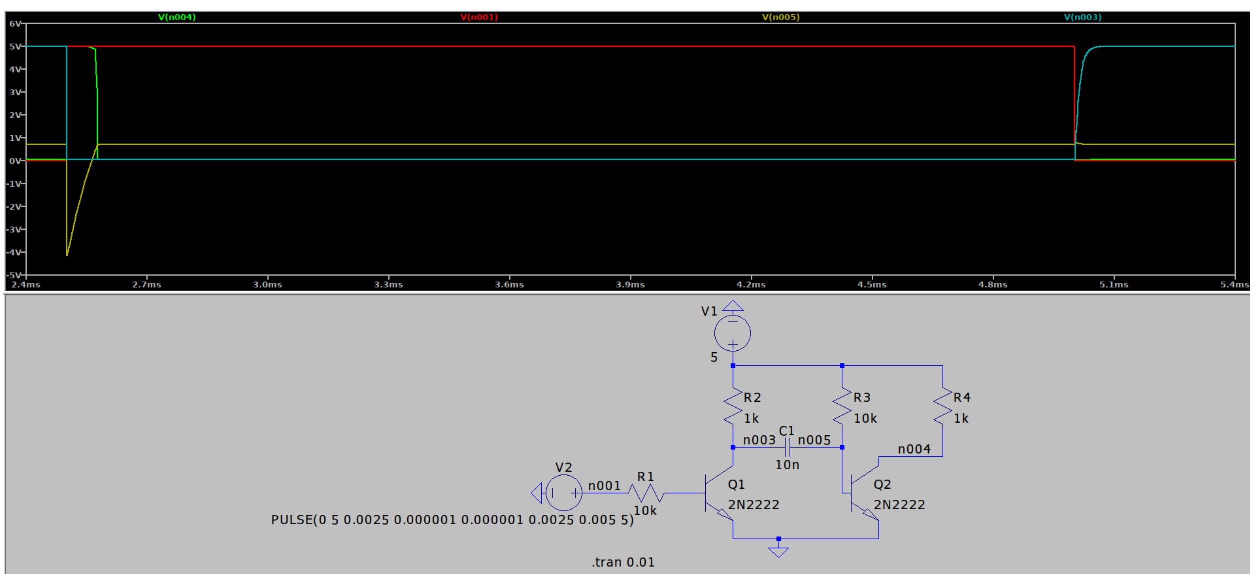
$$I_B = \frac{3.3 - 0.6}{10000} = 0.27 \text{ mA}$$

Hence:

$$\beta = \frac{I_C}{I_B} = \frac{5}{0.27} = 18.5$$

Exercise. This is the solution to exercise 2.2 in the book.

Solution. Spice simulation of the circuit shows the following behavior (schematic and plotting):



Before Q1 is saturated, $V_{B,Q2} = 0.6$V and $I_{B,Q2} = 0.5$ mA Q2 is saturated and $V_{C,Q2} = 0$V. $V_{C,Q1} = 5$V and hence there's a 4.4V drop across the capacitor $\Delta V_{n003-n005}$ and the capacitor is charged.

When V2 is turned on this immediately saturates Q1 as $I_{B,Q1} = \frac{5-0.6}{10k} = 0.44$ mA. and $I_{C,Q1} = \frac{5}{1k} = 5$ mA and most likely $\beta > 12$ and thus its collector drops from 5V to 0V. The drop in potential at the positive plate of C1 from 5V to 0V results in the potential at the negative plate of C1 to drop from 0.6V to -4.4V. This implies that the base-emitter diode of Q2 becomes unbiased and current starts being diverted from flowing to base of Q2 to flowing into capacitor to raise $V_{B,Q2}$.

The current charging the capacitor is:

$$I(t) = C\frac{\mathrm{d}V(t)}{\mathrm{d}t}$$

and

$$V_{CC} - I(t)R_3 = V(t)$$

This holds while $V(t) < 0.6$ V and the base-emitter diode of Q2 is unbiased, hence no current flows into the base of Q2. Therefore:

$$\frac{\mathrm{d}V(t)}{\mathrm{d}t} + \frac{V(t)}{R_3C} = \frac{V_{CC}}{R_3C}$$

The solution is the sum of the homogenous solution and a particular solution. We get:

$$V(t) = V_{CC} + V_0\mathrm{e}^{-\frac{t}{R_3C}}$$

Since $V(0) = -4.4$V and $V_{CC} = 5$V then $V_0 = -9.4$V. Therefore the pulse resets when $V(t) = 0.6$V and Q2 turns on again. Therefore:

$$0.6 = 5 - 9.4\mathrm{e}^{-\frac{t_{pulse}}{R_3C}}$$

or

$$\frac{t_{pulse}}{R_3C} = 0.759$$

or

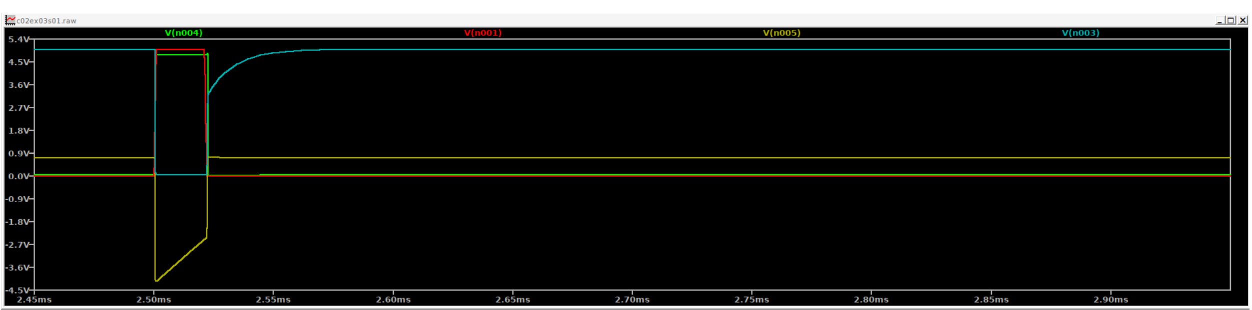
$$t_{pulse} = 0.759R_3C$$

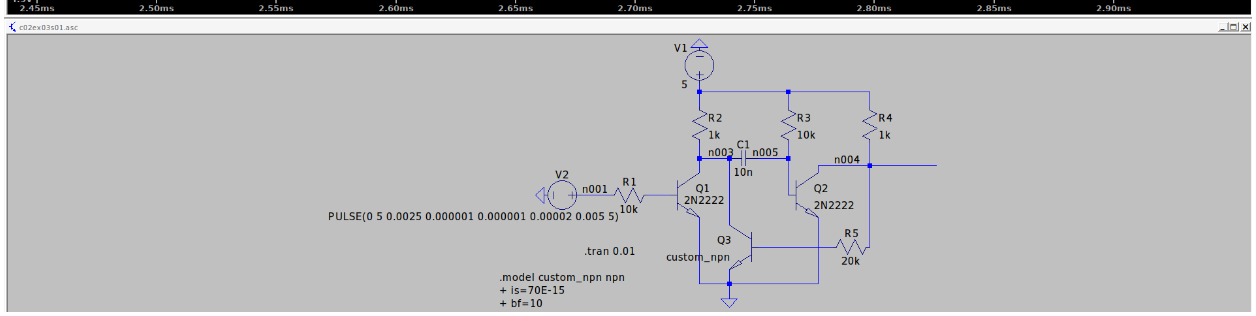
For the values $C = 10$nF and $R_3 = 10$kΩ:

$$t_{pulse} = 0.759 \cdot 10^{-4} \text{ s} = 76\mu\text{s}$$

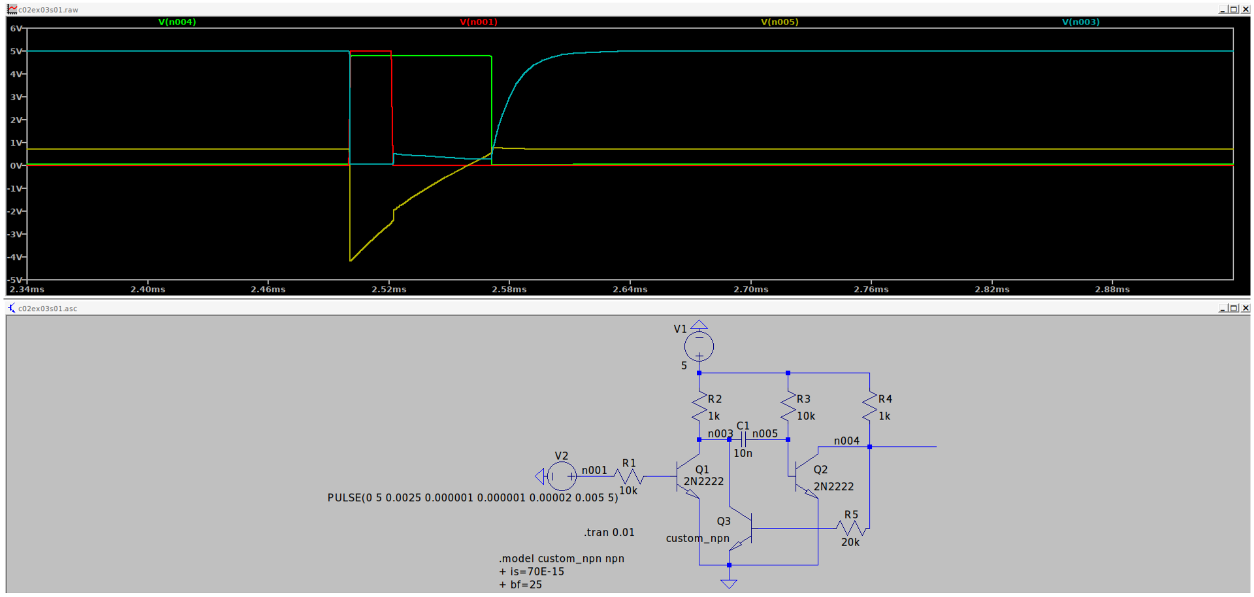
Exercise. This is the solution to exercise 2.3 in the book.

Solution. Spice simulation of the circuit shows the following behavior (schematic and plotting):





when transistor β value is below the minimum required ($\beta = 10$) and when it is above that minimum ($\beta = 25$):



Notice that in the first case the output pulse width follows the input pulse width (in this case some 20 μs) rather than obeying the duration determined by the R_3C time constant of 76 μs. In the second case the output pulse width is as expected.

Now let's calculate the voltage of the output pulse and the minimum β for Q3.

During the output pulse $V_{B,Q3} = 0.6$V and Q2 is turned off. Therefore the current through R4 and R5 is:

$$I_{R4,R5} = I_{B,Q3} = \frac{V_{CC}}{R4 + R5} = \frac{5}{21000} = 0.238 \text{ mA}$$

Hence

$$V_{out} = V_{CC} - I_{R4,R5}R4 = 4.76 \text{ V}$$

While the input pulse is ongoing Q1 is saturated and it is irrelevant if Q3 is saturated yet or not. However, Q3 must reach saturation by the time the input pulse is reset. Then Q3 must be able to handle:

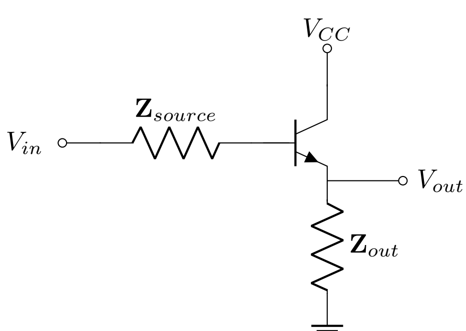
$$I_{C,Q3} = \frac{V_{CC}}{R2} = \frac{5}{1000} = 5 \text{ mA}$$

Therefore, minimally:

$$\beta_{Q3} = \frac{I_{C,Q3}}{I_{B,Q3}} = 21$$

Exercise. This is the solution to exercise 2.4 in the book.

Solution. Consider the generic circuit:



We follow the provided hint to compute the change in emitter voltage:

$$\Delta I_E = \frac{\Delta V_{out}}{\mathbf{Z}_{out}}$$

We also have:

$$I_B = \frac{V_{in} - V_B}{\mathbf{Z}_{source}}$$

hence:

$$\Delta I_B = \frac{\Delta V_B}{\mathbf{Z}_{source}} = \frac{\Delta V_{out}}{\mathbf{Z}_{source}}$$

Also:

$$\Delta I_C = \beta \Delta I_B = \beta \frac{\Delta V_B}{\mathbf{Z}_{source}}$$

But:

$$\Delta I_E = \Delta I_B + \Delta I_C$$

and thus:

$$\frac{\Delta V_{out}}{\mathbf{Z}_{out}} = \frac{\Delta V_{out}}{\mathbf{Z}_{source}} + \beta \frac{\Delta V_{out}}{\mathbf{Z}_{source}}$$

or

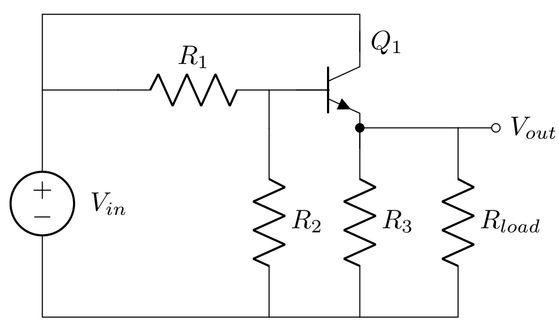
$$\frac{1}{\mathbf{Z}_{out}} = \frac{\beta + 1}{\mathbf{Z}_{source}}$$

or

$$\mathbf{Z}_{out} = \frac{\mathbf{Z}_{source}}{\beta + 1}$$

Exercise. This is the solution to exercise 2.5 in the book.

Solution. Consider the circuit below:



The value of V_{out} is 5V with no load and it can only decrease to 4.75V with a load that draws 25mA. This means that the minimum load impedance is $R_{load} = \frac{5 \cdot 0.95}{25 \cdot 10^{-3}} = 190\Omega$. Using a transistor approximation of $V_{BE} = 0.6$V this means that $V_{B,min} = 5.35$V and $V_{B,max} = 5.6$V. We'll choose $\beta = 78.32$ to match the 2N2369 *npn* we'll use for simulation. This means that the largest base current needed to support the highest load current of 25mA is:

$$I_B(\beta + 1) = I_E = 25 \text{ mA}$$

or

$$I_B \leq 319\mu\text{A}$$

Therefore

$$\frac{V_{in} - V_B}{R_1} - \frac{V_B}{R_2} \leq 319\mu\text{A}$$

or

$$\frac{V_{in}}{R_1} - V_B \frac{R_1 + R_2}{R_1 R_2} \leq 319\mu\text{A}$$

Since

$$\frac{V_{in}}{R_1} - V_B \frac{R_1 + R_2}{R_1 R_2}$$

is largest when $V_B = 5.35$V then:

$$\frac{15}{R_1} - 5.35\frac{R_1 + R_2}{R_1 R_2} \leq 319\mu\text{A}$$

or

$$9.65R_2 - 5.35R_1 \leq 319 \cdot 10^{-6} R_1 R_2$$

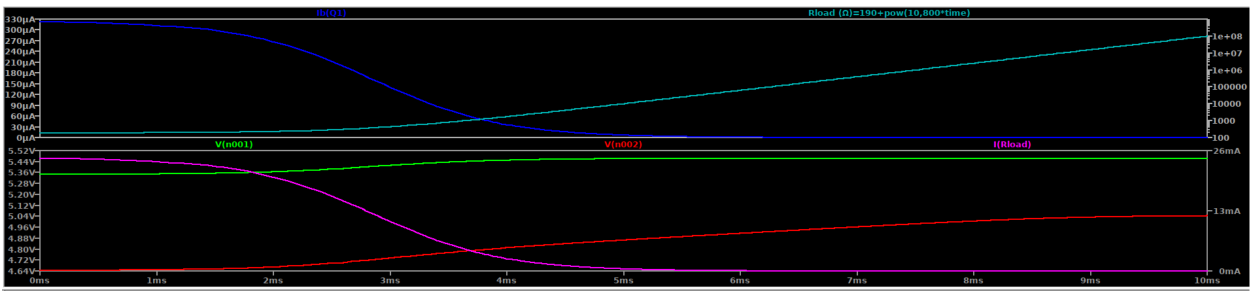
or

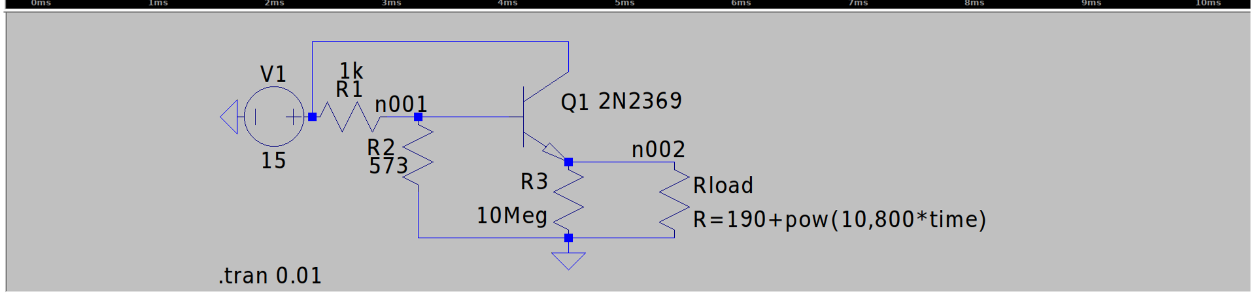
$$R_2 \leq \frac{5.35R_1}{9.65 - 319 \cdot 10^{-6} R_1}$$

Therefore we must choose

$$R_1 < 30.250\text{k}\Omega$$

Let's choose $R_1 = 1\text{k}\Omega$ and thus we must choose $R_2 \leq 573.36\Omega$. We can simulate the circuit in LTSpice using the following schematic and plotting to get:





We can determine that using this particular transistor V_{BE} is load dependent with $V_{BE,max} = 0.698$V at low load impedance and $V_{BE,min} = 0.421$V at high load impedance. This results in V_{out} dropping out of spec on low impedance load and the no load case surpassing the desired 5V. Hence we must adjust the $V_{B,min}$ used for finding the R_1, R_2 resistor constraints above. In this case $V_{B,min} = 5.448$V and $V_{B,max} = 5.421$V. This brings the requirement above to:

$$\frac{15}{R_1} - 5.448\frac{R_1 + R_2}{R_1 R_2} \leq 319\mu\text{A}$$

or

$$R_2 \leq \frac{5.448R_1}{9.552 - 319 \cdot 10^{-6} R_1}$$

Therefore we must choose

$$R_1 < 29.944\text{k}\Omega$$

When there is no load (R_{load} very large) there is essentially no emitter current because we choose a very large value for R_3 and hence the base current is also extremely small. The base current must be non-negative and hence we get:

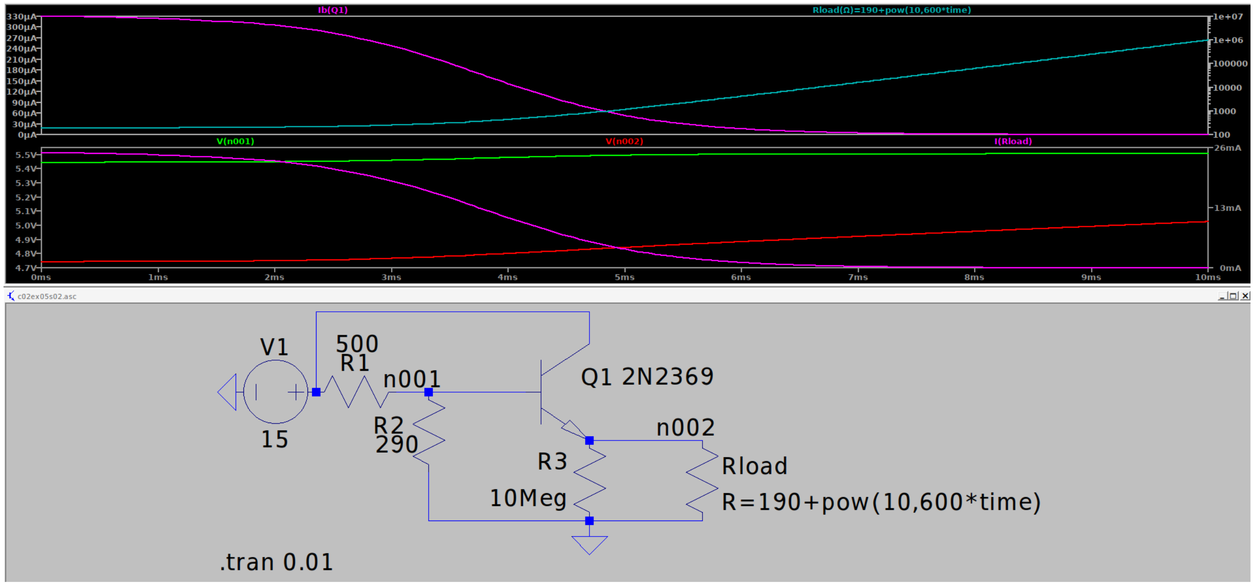
$$R_2 \geq \frac{V_B R_1}{V_{in} - V_B}$$

The largest support value for R_2 is determined for $V_{B,min} = 5.448$V and hence:

$$R_2 \geq 0.5704R_1$$

For example for $R_1 = 1\text{k}\Omega$ we must have $570 \leq R_2 \leq 590\Omega$. However, the plot above was realized with these constraints with this constraint satisfied. Hence, it must be that V_{BE} is further dependent on the R_1. With

$R_1 = 2\text{k}\Omega$ and $1140 \leq R_2 \leq 1222\Omega$ the V_{BE} min/max values for various loads are further reduced. Hence, we must reduce $R_1 < 1\text{k}\Omega$. For $R_1 = 500\Omega$ the valid range R_2 is $285 \leq R_2 \leq 290$. We can simulate the circuit in LTSpice using the following schematic and plotting to get:



Exercise. This is the solution to exercise 2.6 in the book.

Solution. We have the following requirements:

$$V_{in,min} = 20 \text{ V} \qquad V_{out} = 10 \text{ V} \qquad I_{out,max} = 100 \text{ mA}$$

Hence we choose R such that:

$$\frac{10 \text{ V}}{R} > 0.1 \text{ A}$$

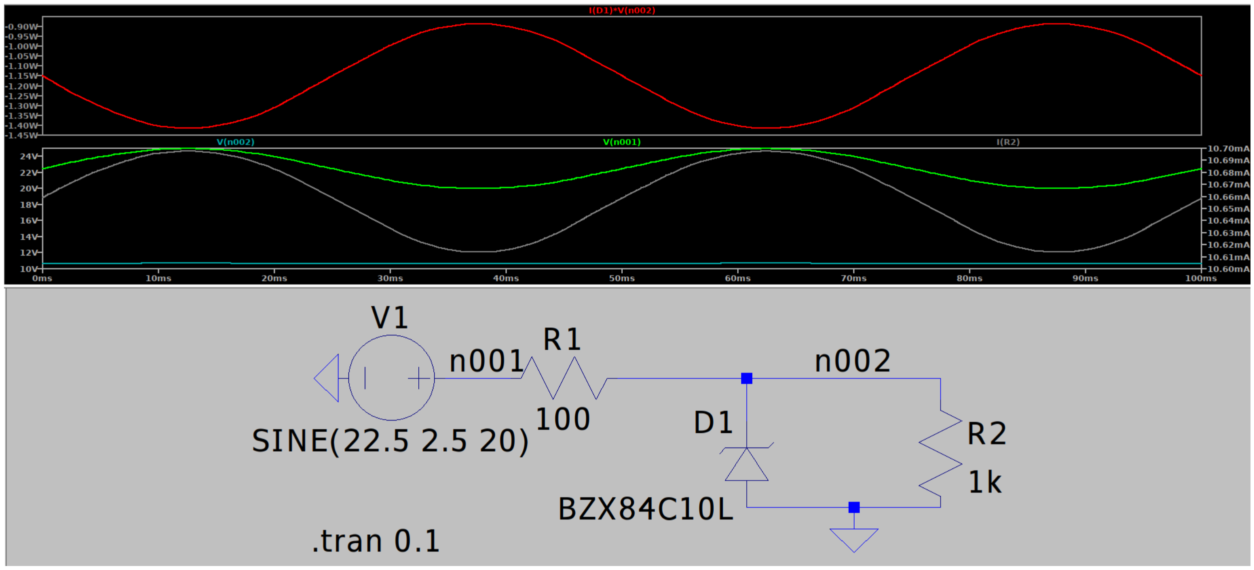
or

$$R < 100 \ \Omega$$

In worst case $I_{out} = 10\text{mA}$ hence:

$$P_{zener} = (I_{out,max} - I_{out,min})V_{zener} = (100\text{mA} - 10\text{mA}) \cdot 10\text{V} = 0.9\text{W}$$

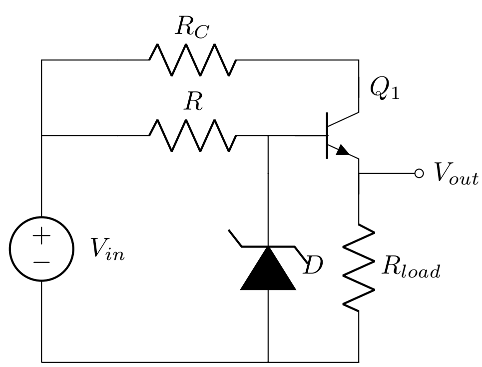
A simulation (schematic,plotting) is shown below:



Indeed the Zener does clamp the voltage however the current through the load does follow the fluctuations of the unregulated source.

Exercise. This is the solution to exercise 2.7 in the book.

Solution.



The Zener is reverse biased and the transistor BE diode is forward biased. Hence we need a Zener with breakdown voltage of 10.6V assuming a 0.6V drop across BE.

We must have:

$$0 \le I_{E,max} \le I_{load,max}$$

or

$$0 \le (\beta + 1) I_B \le I_{load,max}$$

$$I_R = I_B + I_{zener}$$

$$I_{zener} \le I_R \le I_{zener} + I_{load,max}$$

or

$$I_{zener} \le \frac{V_{in} - V_B}{R} \le I_{zener} + I_{load,max}$$

or

$$\frac{V_{in} - V_B}{I_{zener} + \frac{I_{load,max}}{\beta+1}} \le R \le \frac{V_{in} - V_B}{I_{zener}}$$

Using the values with $V_{in,max} = 25\text{V}$, $V_{B,min} = 0.6\text{V}$, $\beta = 78.32$ and $I_{zener,min} = 10\text{mA}$ for the particular transistor we've chosen for simulation we get:

$$2164 \leq R \leq 2440\ \Omega$$

Since

$$I_C = \beta I_B$$

$$0 \leq I_C \leq \frac{\beta I_{load,max}}{\beta + 1}$$

and thus:

$$0 \leq \frac{V_{in} - V_C}{R_C} \leq \frac{\beta I_{load,max}}{\beta + 1}$$

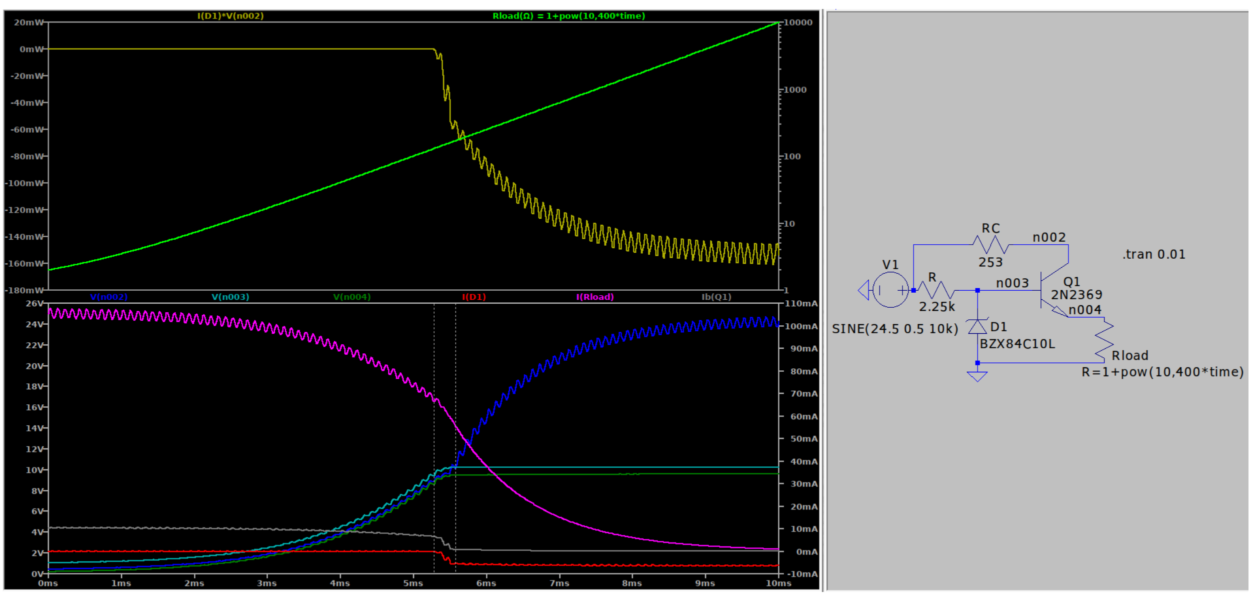
or

$$\frac{(\beta + 1)(V_{in} - V_C)}{\beta I_{load,max}} \leq R_C$$

Again using $V_{in,max} = 25\text{V}$ and $V_{C,min} = 0\text{V}$ (for short circuit) and the other values as above:

$$R_C \geq 253\ \Omega$$

We can simulate the circuit in LTSpice using the following schematic and plotting to get:

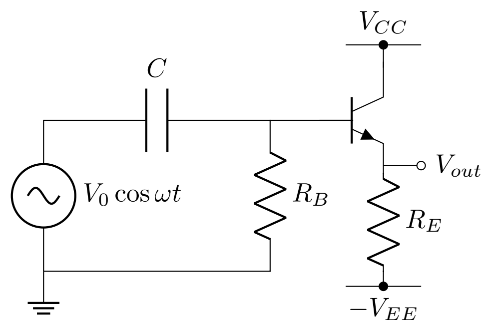


At the low load impedance or load short the Zener is forward biased as the transistor is saturated and $V_{CE} \approx 0$. As the load increases the current through the load decreases and the transistor becomes desaturated up to the point when the base reaches the breakdown voltage of the Zener and the Zener starts drawing reverse bias current. As the load current further decreases/load impedance increases, the transistor is still active but I_C further decreases and the transistor base and emitter are clamped by the breakdown voltage of the Zener. As I_C decreases the collector voltage increases to V_{in} up to the point where the transistor turns off and all current flows through the Zener reverse bias.

If we plot the power thru the Zener we notice that the maximum power dissipated is when there is no load, at about 0.16 W (which is much lower than the power drawn when the emitter follower was not in the circuit).

Exercise. This is the solution to exercise 2.8 in the book.

Solution. Consider the circuit that needs to be implemented:



In this case $V_E = 0$. The voltage drop across R_E is $V_{EE} = 15$V. So for quiescent current of 5mA choose $R_E = 3$ kΩ. We further choose R_B such that $R_B \ll \beta R_E$ or $R_B = \frac{\beta R_E}{10} = 30$ kΩ for $\beta = 100$.

The capacitor C forms a high pass filter with the impedance it sees as load. That load is R_B in parallel with the impedance looking into the transistor base βR_E (we assume the circuit load on transistor emitter is much larger than βR_E). Hence the capacitor sees a load:

$$Z_{C,load} = \frac{R_B \beta R_E}{R_B + \beta R_E} = 27.3 \text{ k}\Omega$$

We must choose:

$$\omega_{3\text{dB}} < 2\pi 20 = 125.66 \, \frac{\text{rad}}{\text{s}}$$

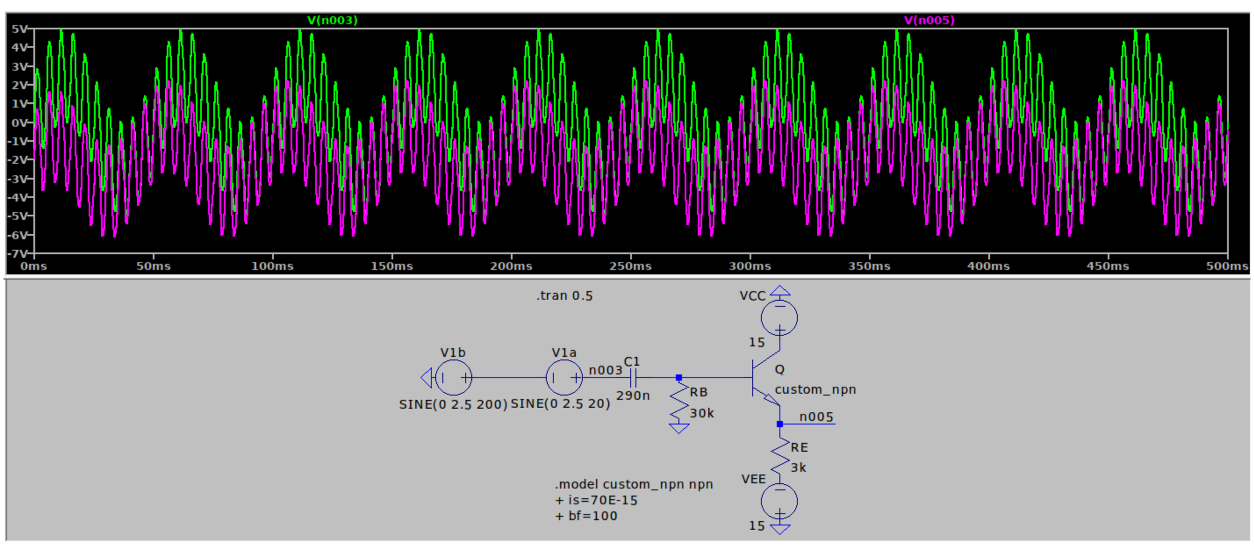
Hence:

$$\frac{1}{Z_{C,load} C} < 126.66$$

or

$$C > 0.29 \, \mu\text{F}$$

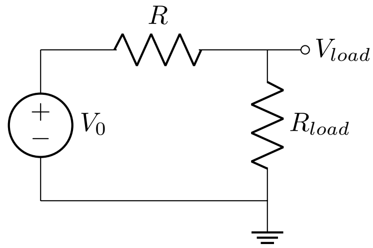
An LTSpice simulation is presented below (see schematic and plotting):



Notice that the 20 Hz signal passes through, as do higher frequency signals. There seems to be a negative DC bias, as the output voltage oscillates around a negative value.

Exercise. This is the solution to exercise 2.9 in the book.

Solution. We can consider a voltage source with a resistor in series and a variable load:



The problem requires that $0 \leq V_{load} \leq 10$ V. Since, when $V_{load} \ll V_{in}$, we have:

$$I = \frac{V_{load}}{R_{load}} = \frac{V_{in} - V_{load}}{R} \approx \frac{V_{in}}{R}$$

For 1% variation in current we must have:

$$0.99\frac{V_{in}}{R} \leq I \leq 1.01\frac{V_{in}}{R}$$

or

$$0.99\frac{V_{in}}{R} \leq \frac{V_{load}}{R_{load}} \leq 1.01\frac{V_{in}}{R}$$

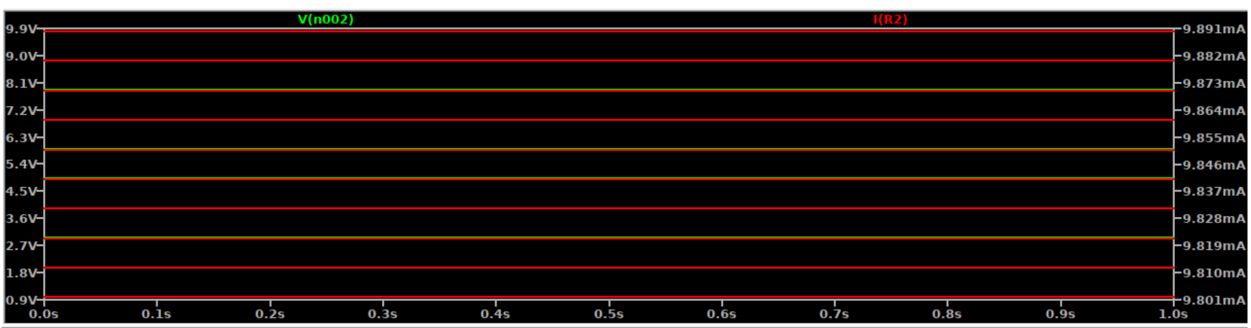
or

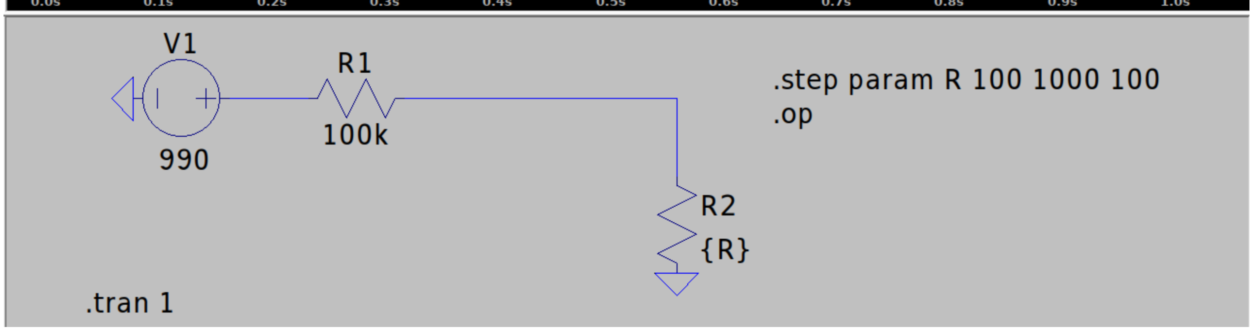
$$V_{in} \geq \frac{V_{load}R}{1.01R_{load}}$$

or

$$V_{in} \geq 9.901\frac{R}{R_{load}}$$

We can simulate the various scenarios using the following schematic and plotting:





Notice that V_{in} is very large to support a various range of loads. Namely, R has to be very large to support multiple R_{load} values, which in turn requires a very large V_{in} to achive a milliamp worth of current.

Exercise. This is the solution to exercise 2.10 in the book.

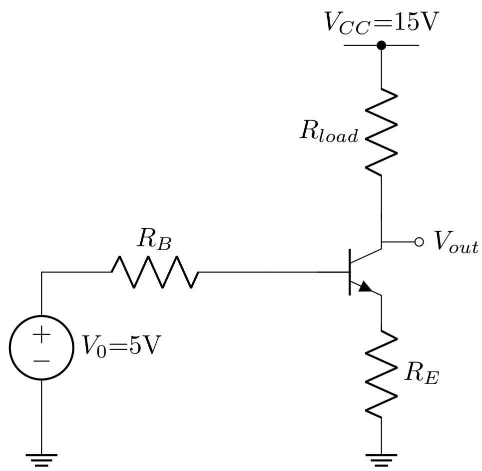
Solution. We have the following power ratio:

$$\frac{P_R}{P_{load}} = \frac{I^2R}{I^2R_{load}} = \frac{R}{R_{load}}$$

Since in this case $R \ll R_{load}$ this shows that most of the power is dissipated in the series resistor rather than the load. This is obviously not very efficient. For example for the circuit shown in exercise 2.9 when $R = 100\text{k}\Omega$ and $R_{load} = 1\text{k}\Omega$ the power dissipated in R is 9.6W while the power dissipated in the load is 96mW.

Exercise. This is the solution to exercise 2.11 in the book.

Solution. Consider the following circuit.



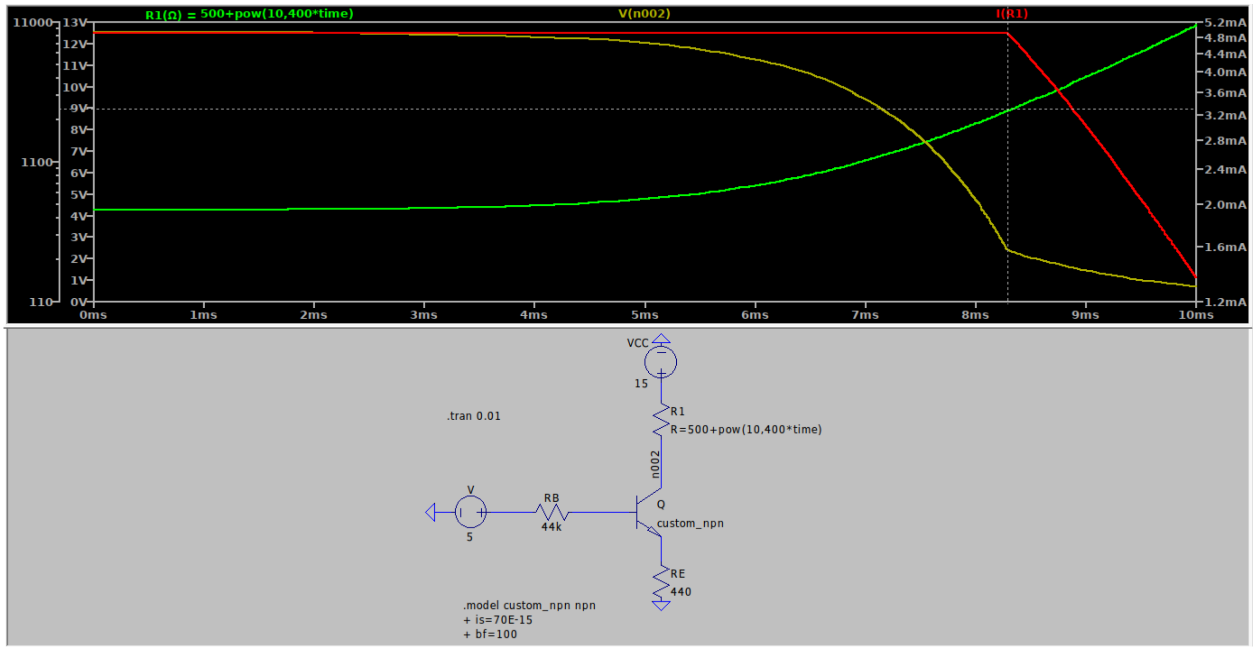
The current thru the collector of the *npn* transistor is $I_C = 5\text{mA}$. We'll use a transistor with $\beta = 100$ hence $I_B = 50\mu\text{A}$ and thus $I_E = 5\text{mA}$. Assuming a 0.6V forward bias drop on the BE diode of the transistor we choose:

$$R_B = \frac{5 - V_B}{I_B} \qquad R_E = \frac{V_B - 0.6}{I_E}$$

If we further choose $R_E = 100R_B$ we get $V_B = 2.8\text{V}$ and $V_E = 2.2\text{V}$ and $R_B = 44\text{k}\Omega$ and $R_E = 440\Omega$. The maximum collector voltage is $V_C = V_E + 0.2$ or $V_C = 2.4\text{V}$. Hence the minimum tolerated $V_C = 2.4$ and the output compliance is $15 - 2.4 = 12.6\text{V}$. Hence the maximum load is:

$$R_{load,max} = \frac{12.6V}{5 \cdot 10^{-3}\text{A}} = 2.52\text{k}\Omega$$

We can simulate the circuit with a variable load using the following LTSpice schematic and plotting.



Notice that the current is constant at 5mA until the load surpasses the determined max value when the current starts to drop and due to collector voltage not being able to further drop, and hence the transistor enters saturation.

Exercise. This is the solution to exercise 2.12 in the book.

Solution. In the first case (grounded emitter) the base input to the BJT is such that the output is a sine wave with amplitude 0.1V or 1V. Hence the maximum variation of gain is:

$$\frac{\Delta G}{G} \approx \frac{\Delta V_{out}}{V_{drop}}$$

and is obtained by considering the maximum deviation of V_{out} from the quiescent level V_{drop}. For 0.1V amplitude we get:

$$\frac{\Delta G}{G} \approx \frac{0.1}{5} = 0.02$$

However, as stated in the text the maximum deviation has both a linear and non-linear components and the non-linear distortion has magnitude 1/3 of the maximum deviation. Hence distortion is 0.02/3=0.0067 or some 0.7%. For 1V amplitude

$$\frac{\Delta G}{G} \approx \frac{1}{5} = 0.2$$

and hence the distortion is some 6.7%.

In the second case the emitter is connected to the ground through a resistor R_E. Then the gain is:

$$G = -\frac{R_C}{r_e + R_E}$$

Using:

$$r_e = \frac{V_T}{I_C}$$

then

$$G = -\frac{R_C I_C}{V_T + I_C R_E}$$

Differentiating (the variable is I_C):

$$\mathrm{d}G = -\frac{R_C \mathrm{d}I_C}{V_T + I_C R_E} + \frac{R_C I_C R_E \mathrm{d}I_C}{(V_T + I_C R_E)^2}$$

or

$$\mathrm{d}G = -\frac{R_C V_T \mathrm{d}I_C}{(V_T + I_C R_E)^2}$$

hence:

$$\frac{\mathrm{d}G}{G} = \frac{R_C V_T \mathrm{d}I_C}{(V_T + I_C R_E) R_C I_C}$$

or

$$\frac{\mathrm{d}G}{G} = \frac{V_T \mathrm{d}V_{drop}}{(V_T + I_C R_E) V_{drop}}$$

or

$$\frac{\mathrm{d}G}{G} = \frac{r_e \mathrm{d}V_{drop}}{(r_e + R_E) V_{drop}}$$

Thus indeed by adding an emitter resistor the gain distortion is reduced by a factor of $r_e/(r_e + R_E) = V_T/(V_T + I_C R_E)$.

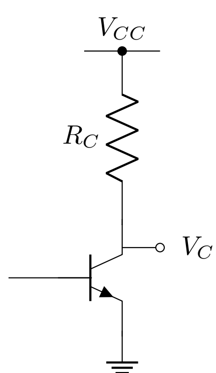
Using $V_T = 25$ mV and $I_C R_E = 250$ mV (the 0.25V drop across the emitter resistor at quiescent current):

$$f_G = \frac{V_T}{V_T + I_C R_E} = \frac{25}{250 + 25} = 0.091$$

Hence the distortion is reduced to 0.0067·0.091=0.061% for 0.1V amplitude output and 0.067·0.091=0.607% for 1V amplitude case.

Exercise. This is the solution to exercise 2.13 in the book.

Solution. The circuit is:



The collector voltage drop at the initial temperature is:

$$V_{drop,0} = I_{C,0} R_C = 0.5 V_{CC}$$

Since collector current increses by 9%/°C then when temperature increases by 8°C:

$$I_{C,1} = (1.09)^8 I_{C,0} = 1.99 I_{C,0}$$

Therefore:

$$V_{drop,1} = I_{C,1} R_C = 0.996 V_{CC}$$

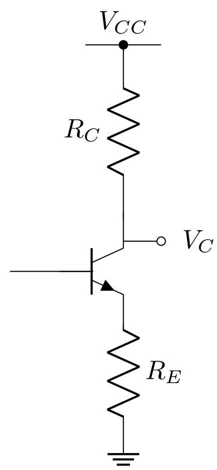
Therefore:

$$V_{CE} = 4 \cdot 10^{-3} V_{CC}$$

and thus the transistor is in saturation.

Exercise. This is the solution to exercise 2.14 in the book.

Solution. Consider the circuit:



Let's first understand how some of the values depicted in Figure 2.49 of the text are derived. The requirements are $V_{CC} = 20$V, $I_{C,0} = 1$mA, and $V_C = 10$V and a gain of 50. This obviously means that:

$$R_C = \frac{V_{CC} - V_C}{I_{C,0}} = 10 \text{ k}\Omega$$

and since:

$$|G| = \frac{R_C}{r_e + R_E} = 50$$

and $r_e = 25/I_C(\text{mA}) = 25\ \Omega$:

$$R_E = \frac{R_C}{|G|} - r_e = 200 - 25 = 175\ \Omega$$

At the initial temperature $I_{C,0} = 1$ mA and $V_E = I_E R_E \approx I_{C,0} R_E = 0.175$ V. Therefore $V_B = 0.775$ V.

Since the base-emitter voltage drop reduces by 2.1mV/°C then, when temperature increases by 20 °C $V_{BE} = 0.6 - 0.042 = 0.558$V. As the base is held at 0.775 V by the resistor divider then emitter voltage must increase to $V_E = 0.775 - 0.558 = 0.217$ V. This means that emitter current must be $I_E \approx I_{C,1} = \frac{V_E}{R_E} = 1.24$ mA.

Therefore:

$$\frac{I_{C,1}}{I_{C,0}} = 1.24$$

or collector current increased by nearly 25 %.

Notice that in exercise 2.13 we've used an estimate of a 9%/°C increase in collector current for the grounded emitter amplifier. Using this estimate in this case we would get $I_{C,1} = (1.09)^2 I_{C,0} = 5.6 I_{C,0}$. But such an estimate is incorrect because the transistor would saturate well before that collector current is reached. Moreover, as noted in the text, R_E provides feedback with respect to the base-emitter voltage drop, and hence the collector current change is determined by V_{BE} change.

Exercise. This is the solution to exercise 2.15 in the book.

Solution. For the grounded emitter amplifier:

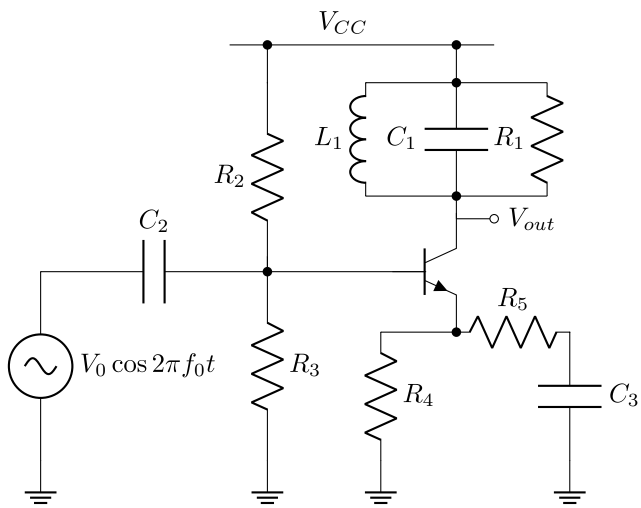
$$G_V = -\frac{R_C}{r_e} = -\frac{I_C R_C}{V_T} = -\frac{0.5 V_{CC}}{V_T}$$

Since $V_T = \frac{kT}{q} = 0.025$V at about 300K. Therefore:

$$G_V = -20 V_{CC}$$

Exercise. This is the solution to exercise 2.16 in the book.

Solution. The circuit looks like below:



We need to calculate the capacitor value in in the LC circuit. From section 1.7.14 we have:

$$Q = \omega RC = \frac{R}{\omega L}$$

or

$$Q = \sqrt{\frac{R^2 C}{L}}$$

Using the provided values for $R = 6.2$ kΩ and $L = 1$mH we get:

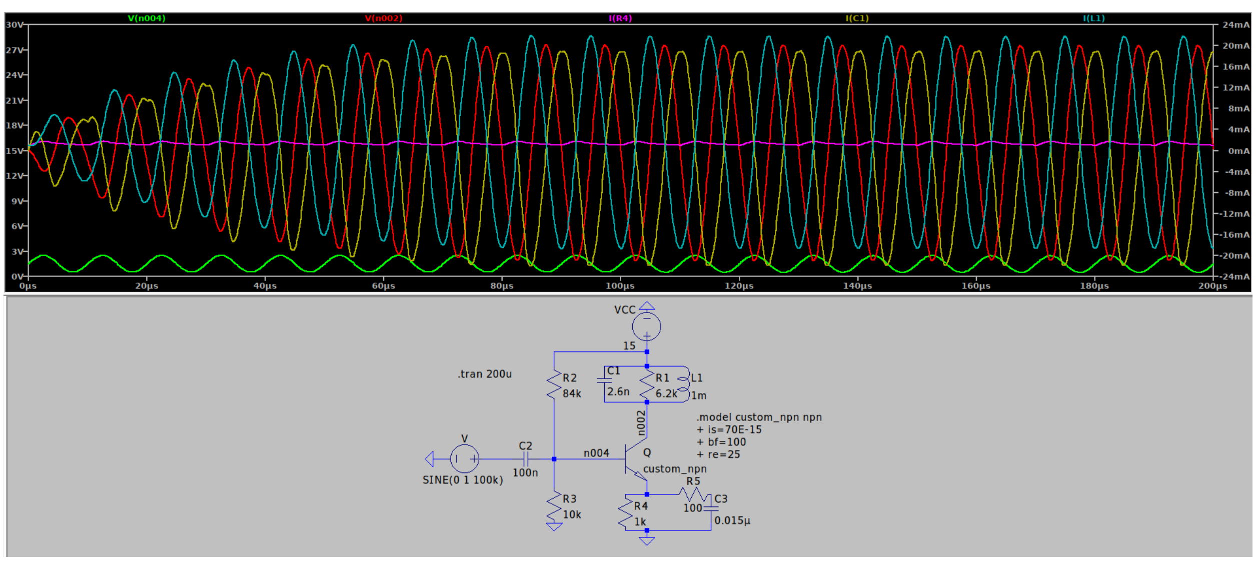
$$C = 2.6 \text{ nF}$$

To get the 1mA quiescent current we use a similar approach to Figures 2.50/2.51 in the text except adjust the bypassing capacitor for 100 kHz operation. Hence the bypass capacitor will be 5000× smaller or:

$$C_{bypass} \approx 15 \text{ nF}$$

We use the suggested modification of Figure 2.51 in the text and run the DC path thru a 1k emitter resistor and the signal path thru a resistor and bypass capacitor.

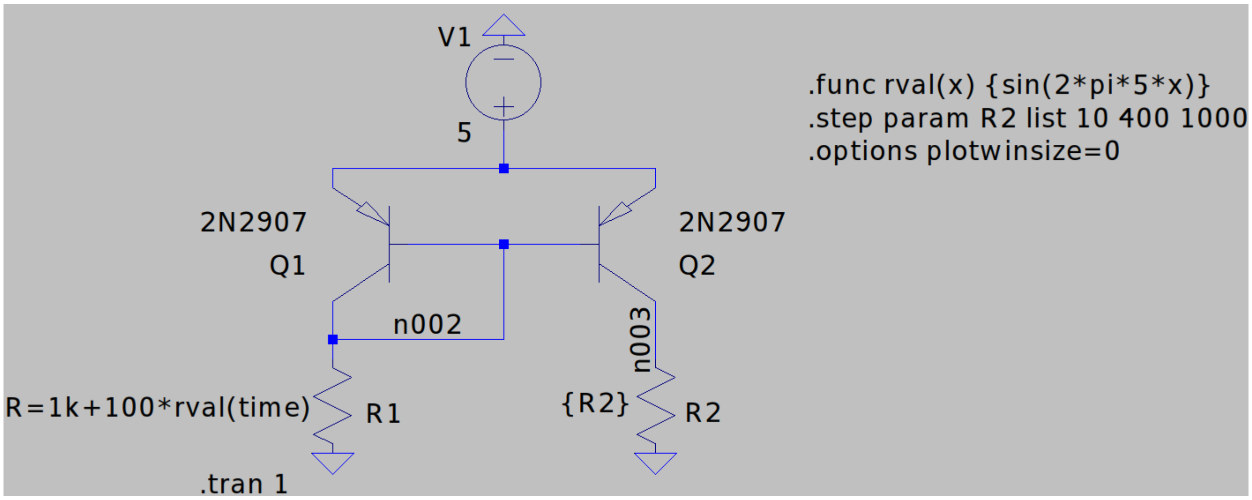
We can simulate the circuit with the following LTSpice schematic and plotting.



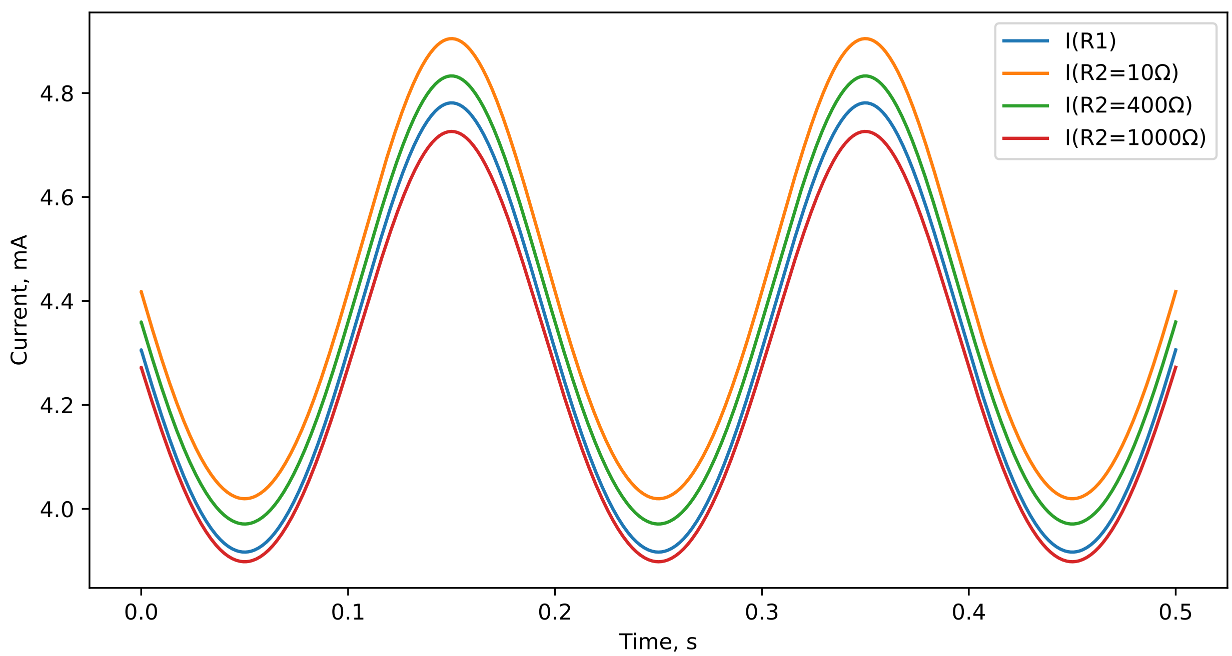
Notice that the transistor base oscillates at 100 kHz around the bias voltage of 1.6V and the collector around V_{CC} of 15V. One can easily verify that increasing the value of resistor R_5 decreases the gain as the impedance of the emitter side increases.

Exercise. This is the solution to exercise 2.17 in the book.

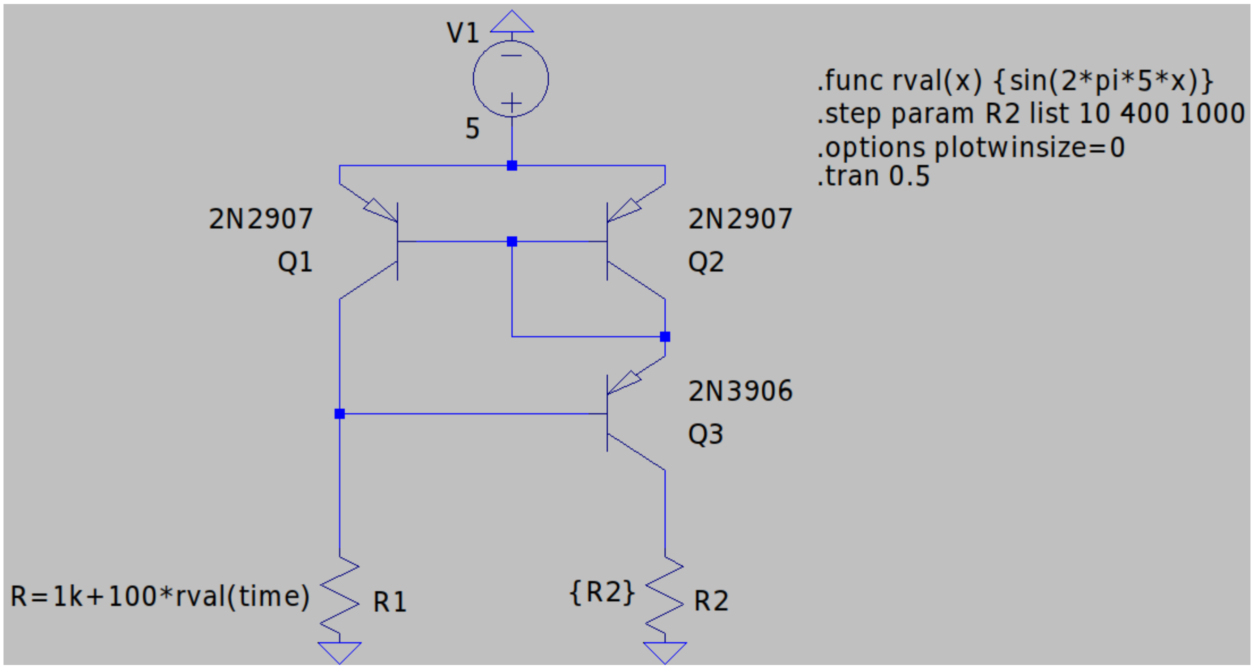
Solution. We can use LTSpice to simulate the circuits. Consider the simple current mirroring circuit below:



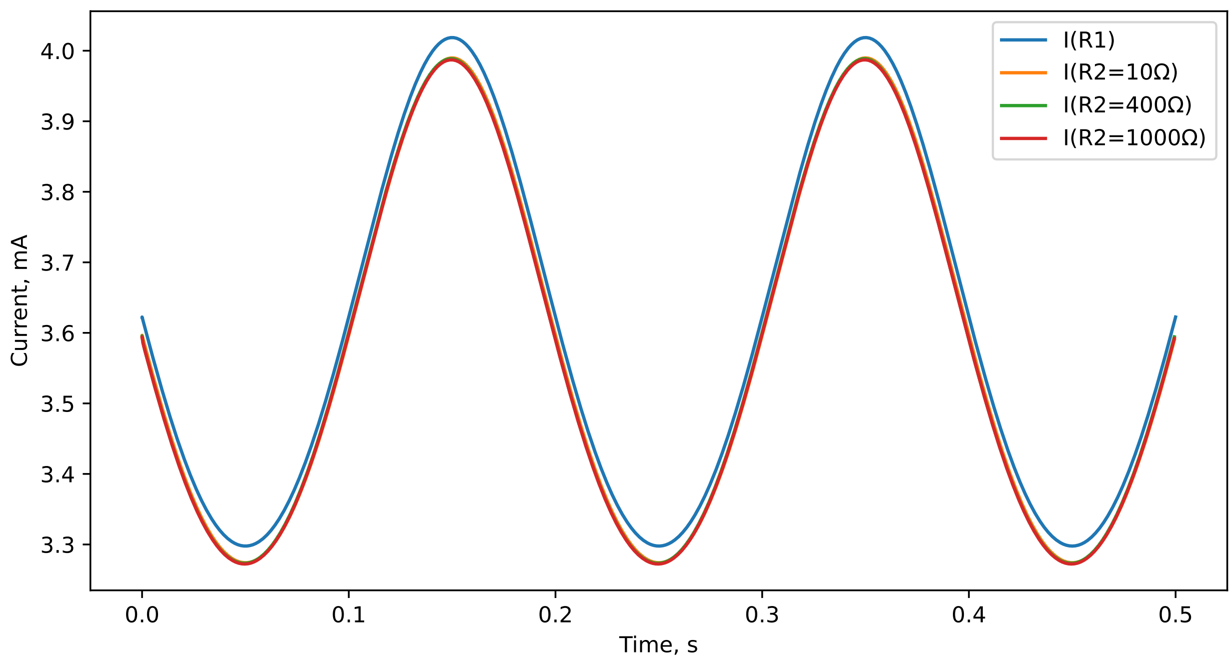
We can use the following schematic (.cir for batch) simulation. We can run the LTSpice simulation using the following Python script and use LTSpice Python raw data access from ltspy3.py.



Now consider the improved current mirror with an additional *pnp* but with Q3 of different type from Q1/Q2.



We can use the following schematic (.cir for batch) simulation. We can run the LTSpice simulation using the following Python script.



Notice that indeed in this case the current mirroring performs better.

Exercise. This is the solution to exercise 2.18 in the book.

Solution.

Exercise. This is the solution to exercise 2.19 in the book.

Solution.

Exercise. This is the solution to exercise 2.20 in the book.

Solution.

Exercise. This is the solution to exercise 2.21 in the book.

Solution.

Exercise. This is the solution to exercise 2.22 in the book.

Solution.

Exercise. This is the solution to exercise 2.23 in the book.

Solution.

Exercise. This is the solution to exercise 2.24 in the book.

Solution. (a) We can calculate

$$V_B = 10\text{V}\frac{1.6\text{k}}{1.6\text{k} + 8.2\text{k}} = 1.632\text{V}$$

With $V_{BE} = 0.6$V then $V_E = 1.032$V and thus $I_E = 0.688$mA. For *npn* with large β then $I_C = I_{load} \approx I_E = 0.688$mA.

Since $V_E = 1.032$V then before the transistor enters saturation $V_{CE} > 0.1$V. Hence the minimum $V_C = 1.132$V. Therefore the output compliance is:

$$V = V_{CC} - V_{C,min} = 10\text{V} - 1.132\text{V} = 8.87\text{V}$$

(b) If we consider the nonsaturated region of the *npn* with $I_C = \beta I_B$ and $R_1 = 8.2\text{k}\Omega$, $R_2 = 1.6\text{k}\Omega$ and $R_E = 1.5\text{k}\Omega$ one can derive the following expression:

$$I_B = \frac{V_{CC}R_2 - V_{BE}(R_1 + R_2)}{R_1R_2 + R_E(\beta + 1)(R_1 + R_2)}$$

and also:

$$V_B = \frac{V_{CC}R_2 - I_BR_1R_2}{R_1 + R_2}$$

Hence with $V_{BE} = 0.6$V and $V_{CC} = 10$V we get for $\beta = 50$:

$$I_B = 13.266\mu\text{A} \qquad I_C = \beta I_B = 0.6633\text{mA}$$

and for $\beta = 100$:

$$I_B = 6.756\mu\text{A} \qquad I_C = \beta I_B = 0.6756\text{mA}$$

Thus as the *npn* β decreases the load current decreases within the output compliance range.

(c) Notice that $V_{CE} \approx 0.1$V close to saturation at the limit of output compliance and $V_{CE} \approx 8.868$V when the load impedance is 0Ω (and $V_C = V_{CC} = 10$V). Hence, over the compliance range V_{BE} will decrease by at most 0.8868mV from 0.6V down to 0.5991132V.

Hence using the expression above for I_B and I_C we can calculate for a *npn* with $\beta = 50$ that $I_B = 13.277\mu$A and $I_C = 0.6638$mA. For $\beta = 100$ then $I_B = 6.7623\mu$A and $I_C = 0.6762$mA.

Thus, as load impedance decreases and V_{CE} increases the current through the load slightly increases.

(d) In the nonsaturated region of the *npn* with $I_C = \beta I_B$ we use:

$$I_B = \frac{V_{CC}R_2 - V_{BE}(R_1 + R_2)}{R_1R_2 + R_E(\beta + 1)(R_1 + R_2)}$$

and the fact that V_{BE} depends on temperature to determine the temperature dependance of I_C when β is constant.

$$I_C = \frac{\beta(V_{CC}R_2 - V_{BE}(R_1+R_2))}{R_1R_2 + R_E(\beta+1)(R_1+R_2)}$$

or

$$\frac{\mathrm{d}I_C}{\mathrm{d}T} = -\frac{\mathrm{d}V_{BE}}{\mathrm{d}T}\frac{\beta(R_1+R_2)}{R_1R_2 + R_E(\beta+1)(R_1+R_2)}$$

For $\beta = 100$ and $\frac{\mathrm{d}V_{BE}}{\mathrm{d}T} = -2.1\mathrm{mV}/°\mathrm{C}$ (see section 2.3.2.C in the text which discusses the dependance of V_{BE} on temperature to include effect of V_T and $I_S(T)$):

$$\frac{\mathrm{d}I_C}{\mathrm{d}T} = 1.374\mu\mathrm{A}/°\mathrm{C}$$

Similarly considering that $\frac{\mathrm{d}\ln\beta}{\mathrm{d}T} = \ln(1+0.004) = 0.004$ and with $\beta = 100$:

$$\begin{aligned}\frac{\mathrm{d}I_C}{\mathrm{d}T} \quad = \quad & -\frac{\mathrm{d}V_{BE}}{\mathrm{d}T}\frac{\beta(R_1+R_2)}{R_1R_2 + R_E(\beta+1)(R_1+R_2)} \\ & -\frac{\beta^2(V_{CC}R_2 - V_{BE}(R_1+R_2))R_E(R_1+R_2)}{(R_1R_2 + R_E(\beta+1)(R_1+R_2))^2}\frac{\mathrm{d}\ln\beta}{\mathrm{d}T} \\ & +\frac{\beta(V_{CC}R_2 - V_{BE}(R_1+R_2))}{R_1R_2 + R_E(\beta+1)(R_1+R_2)}\frac{\mathrm{d}\ln\beta}{\mathrm{d}T}\end{aligned}$$

or using numerical values:

$$\frac{\mathrm{d}I_C}{\mathrm{d}T} = 1.424\mu\mathrm{A}/°\mathrm{C}$$

Exercise. This is the solution to exercise 2.25 in the book.

Solution.

Exercise. This is the solution to exercise 2.26 in the book.

Solution.

Exercise. This is the solution to exercise 2.27 in the book.

Solution.

Exercise. This is the solution to exercise 2.28 in the book.

Solution.

Exercise. This is the solution to exercise 2.29 in the book.

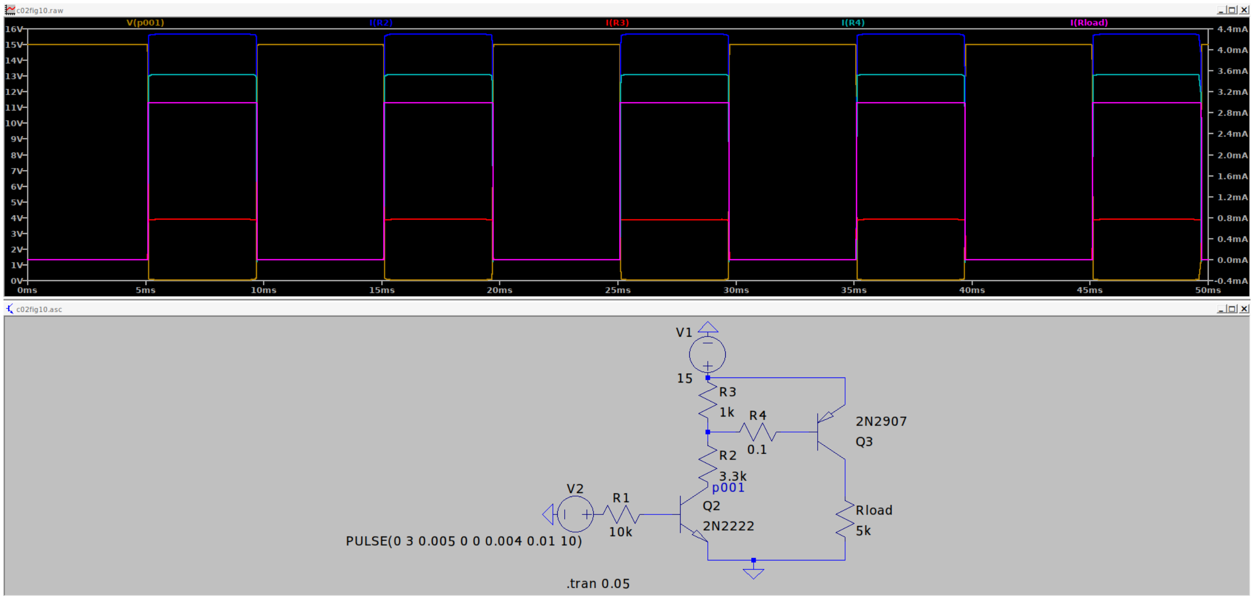
Solution.

Exercise. This is the solution to exercise 2.30 in the book.

Solution.

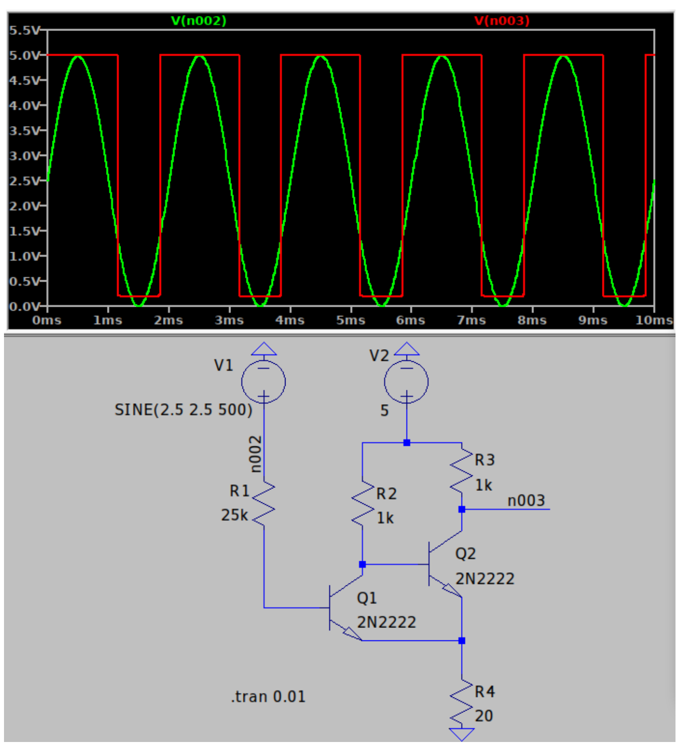
Notes. Extra derivations for chapter 2.

There is a discussion around figure 2.10 in the text that is investigated below. See the LTSpice schematic and plotting.



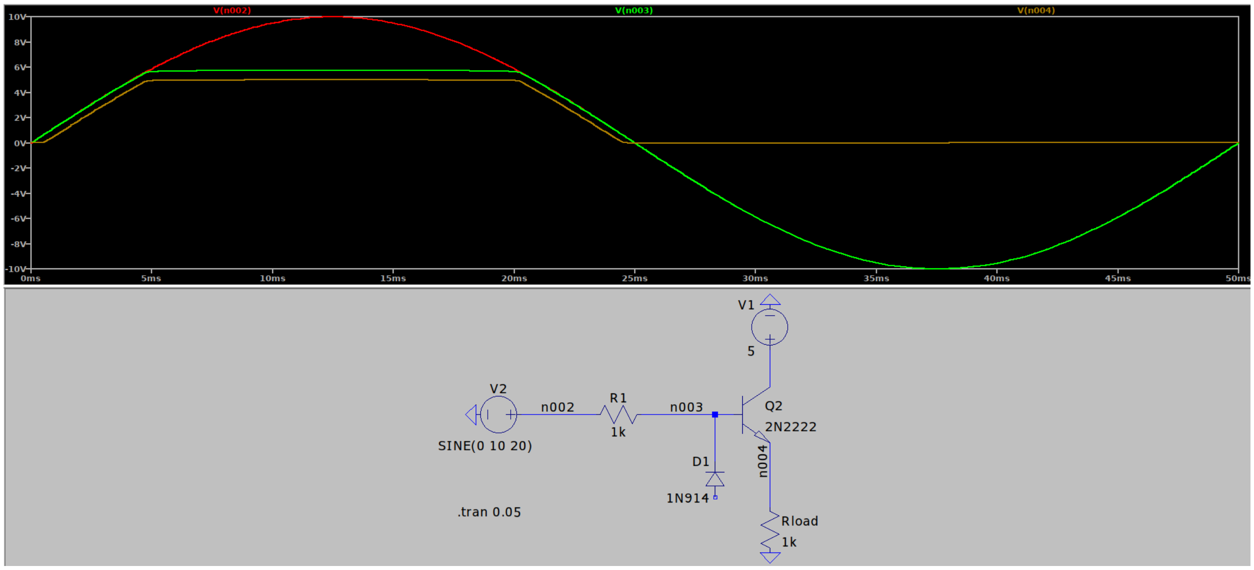
Notice that when the V2 pin is on at 3V Q2 becomes saturated and hence the voltage at collector $V_{C,Q2} = 0$. This turns on Q3, whose base is now forward biased to about 14.4V. Therefore the current through R3 is about 0.6mA and the current thru R2 is $14.4/3.3 = 4.4$mA. The base current $I_{B,Q3} = 3.7$mA. The collector at Q3 sits at $V_{C,Q3} = 15$V and hence the current thru the 5kΩ load is 3mA. Indeed, as the text suggests the voltage divider would sit at 11.5V were it not for the forward biased BE diode of Q3 as $11.5 = 15\frac{3.3}{3.3+1}$.

We can try to simulate the Schmitt trigger mentioned in figure 2.13 to understand how it can make the output of the pulse generator turn off sharply (schematic, plotting).



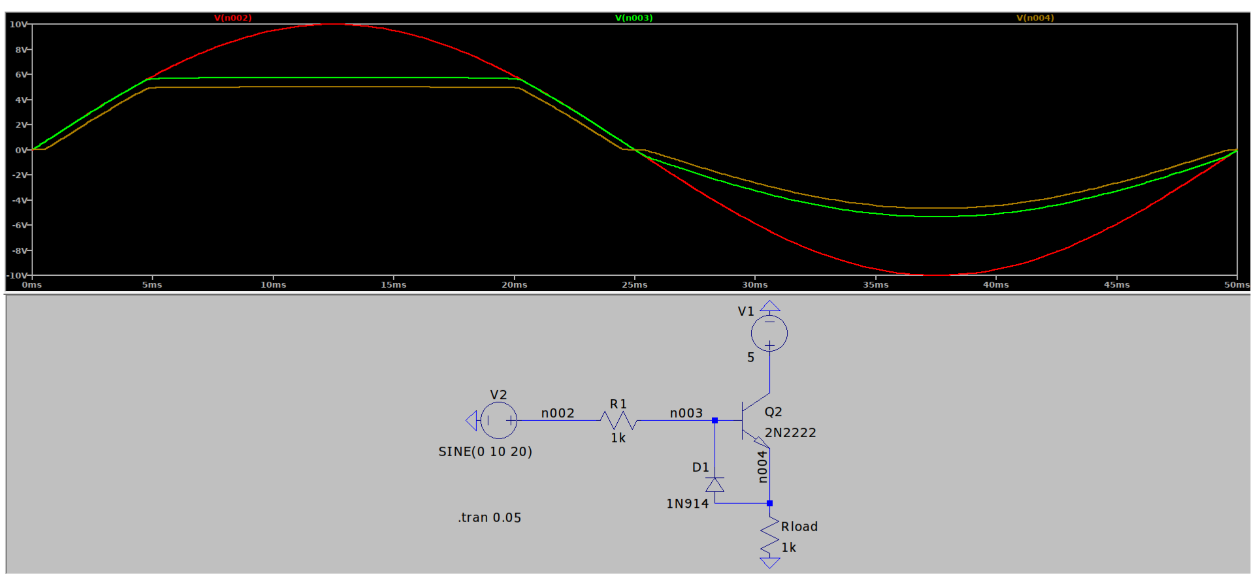
When the base voltage of Q_1 drops below the value which leads to the Q_1 collector voltage to be larger than the base bias voltage of Q_2 then Q_2 becomes active and collector voltage on Q_3 drops to a few hundreds of mV above the emitter of Q_2, which owing to resistor R_4 and the maximum I_C current of Q_2 will only sit a few tens of mV above ground. Hence, the output essentially turns off.

There's a bit of discussion around figure 2.19 in the text about protective diode for base-emitter breakdown of a transistor. We can simulate the conditions that would trigger the base-emitter breakdown as shown below:



As the voltage increases the transistor becomes active once base voltage goes above BE diode forward bias. The transistor saturates when the emitter voltage reaches $V_{CC} = 5$V. As the voltage reduces and then becomes negative the transistor desaturates, then stays active for a while, then turns off. It stays off while V_B is negative. However, it is during this period that the BE diode becomes reverse biased and can experience breakdown.

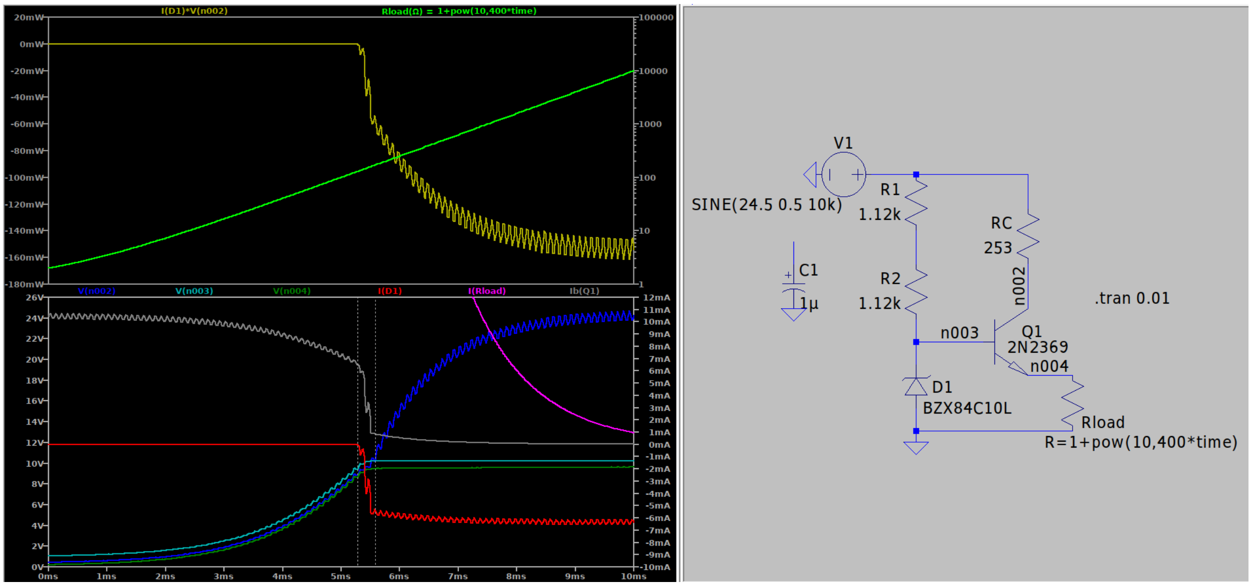
Now we can add the protective diode and observe the behavior:



In this case, instead of BE diode becoming reverse biased and reaching breakdown, the protective diode becomes forward biased and the base voltage V_B of Q1 is determined by the voltage divider created by R1 and Rload and the forward bias drop of D1.

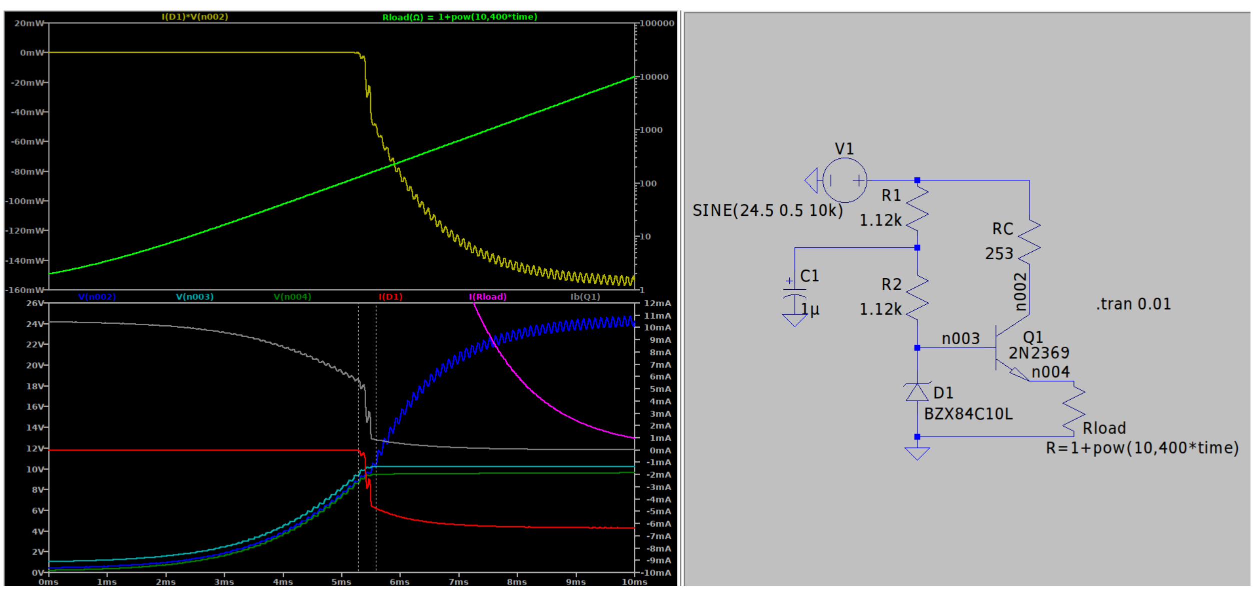
Simulate with the following schematic and plotting to get the results above.

We can notice the ripple in the Zener current in the plot in exercise 2.7 for $I(D1)$ (red line). It was suggested that a circuit like that shown in figure 2.22 of the text eliminates the ripple in the Zener current/voltage. Here's a replot of the situation discussed in exercise 2.7:



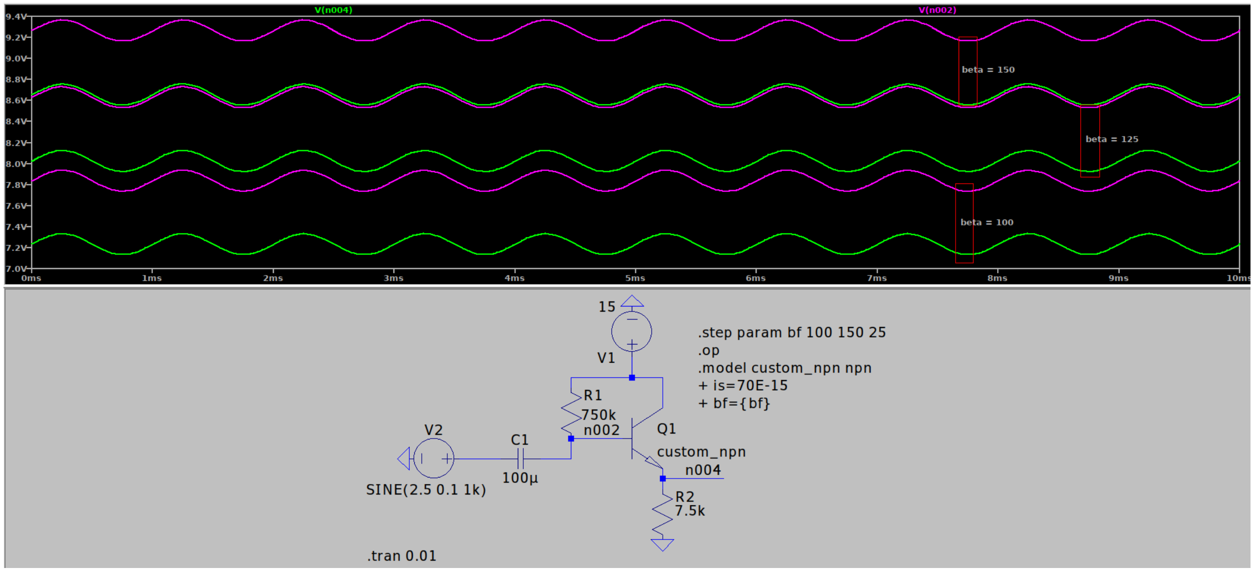
which more clearly shows the ripple in the diode current. The ripples are at 10kHz as the source oscillates at that frequency. Hence we must choose $RC \gg 1.6 \cdot 10^{-5}$s. For example if we choose the capacitor for 10nF the ripple will still be present but if we choose it as 1μF the ripple dissapears.

Then we can connect such a capacitor and observe the effect:



Simulate with the following schematic and plotting to get the results above.

Let's evaluate a bit the design detailed in figure 2.28. Indeed, if $\beta = 100$ then the output is around 7V, but as β changes both the base voltage and the output voltage change away from around 7V. Notice that there is no path for the current sourced through R1 to sink except for the base to emitter current of the transistor.



Simulate with the following schematic and plotting to get the results above.
