Electronics of Radio, Part 1

Notes on David Rutledge’s book

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Basic concepts

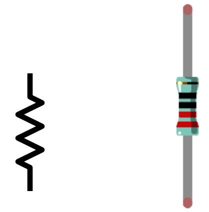
- Potential difference (V, ϕ): $\phi = \int_a^r E \cdot ds$, energy per charge, 1V = 1 J/C
- Kirhhoff loop: $\sum_{loop} V_i = 0$ (Conservation of energy)
- Kirhhoff node: $\sum_{node} I_i = 0$ (Conservation of charge)
- $\mathrm{V}(t) = V_p \cos(\omega t)\,, \omega = 2\pi f, \mathrm{I}(t) = I_p \cos(\omega t)\,, \omega = 2\pi f$
- Instantaneous power: $P(t) = V(t)I(t) = V_p I_p\, cos^2(\omega t)$
- Average power: $P_a = \int_0^{1/f} V(t)I(t)dt = \int_0^{2\pi/\omega} V_p I_p\, cos^2(\omega t)dt = \frac{V_p I_p}{2}$
- Band names:

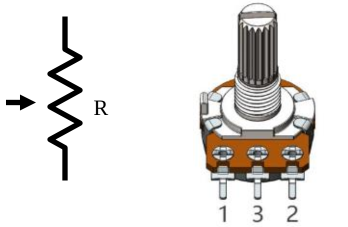
Name	Frequency
VLF	3-30kHz
LW	20-300kHz
MW	300kHz-3MHz
HF	3MHz-30MHz
VHF	30-300MHz

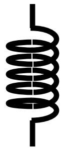
Name	Frequency
UHF	300MHz-1GHz
uW	1-30GHz
milliW	30-300GHz
submilliF	>300GHz

Resistors, capacitors, inductors

- Resistors
  - Analytic model: $IR = V$
  - Energy dissipated: $E = \int_{t_i}^{t_f} IV\,\mathrm{dt} = \int_{t_i}^{t_f} I^2 R dt$
- Capacitors
  - Analytic model: $CV = q, C\frac{dV}{dt} = i$
  - Capacitor Energy stored: $E = \int_{t_i}^{t_f} CV\frac{dV}{dt}\mathrm{dt} = \frac{1}{2}CV^2$
- Inductors
  - Analytic model: $V = L\frac{di}{dt}$
  - Inductor Energy stored: $E = \int_{t_i}^{t_f} IV\,\mathrm{dt} = \int_{t_i}^{t_f} LI\frac{dI}{dt}dt = \frac{1}{2}LI^2$
  - Open air: $L(H) = \mu_0 K n^2 \frac{A}{l}$, distances in meters, $\mu_0 = 4\pi \times 10^{-7}, K = 1$



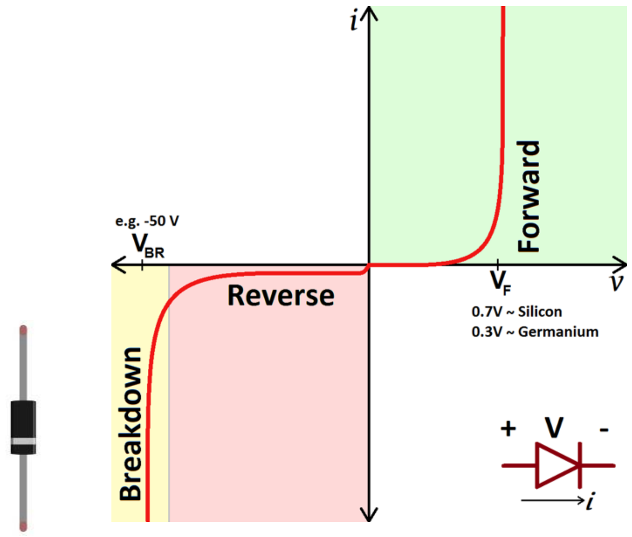




Credit: Make Electronics

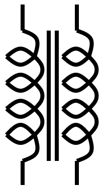
Diodes, transformers

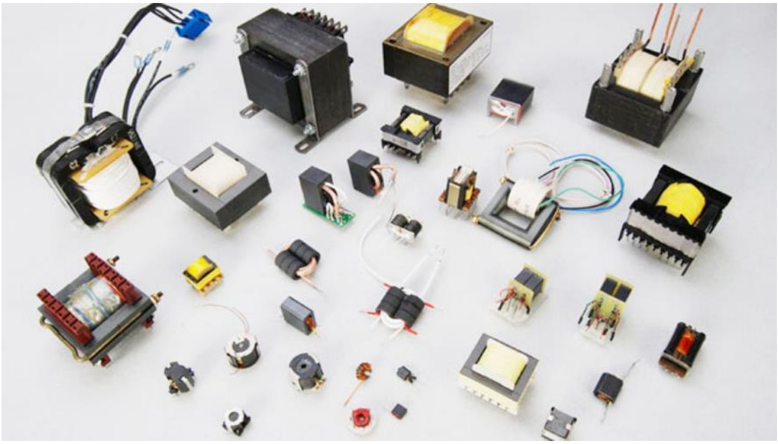
- Diodes
  - Devices that allow current to flow only in one direction
  - Silicon diodes, for example have, essentially infinite resistance if $V_{ac}<0$, that is if the cathode is at a higher potential than the anode and very low resistance if $V_{ac}>.7V$.
  - The cathode is usually labelled with a band
  - Schottky diode: A metal-semiconductor contact that has a lower forward voltage drop than a PN diode.
  - Zener diode: Provides a stable reference voltage.
  - Tunnel diode: Has negative resistance due to tunneling, a quantum mechanical effect.
  - PN junction diode: A diode made of two layers of semiconductor material.
  - PIN diode: For high frequency (low capacitance)



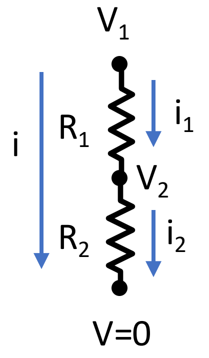
Credit: Make Electronics

- Transformers
  - AC only: $\frac{N_2}{N_1} = \frac{V_2}{V_1}$

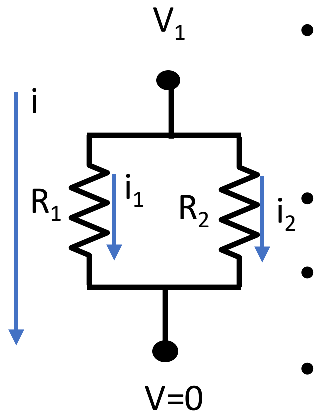




Simple circuit analysis with Kirchhoff

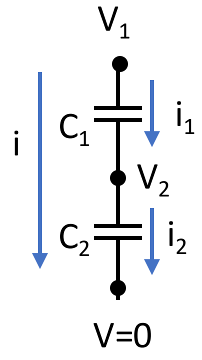


- R_{eq} is the equivalent resistance, replacing the top left circuit with a single resistance.
- By Kirchhoff's node rule, $i_1 = i_2 = i$, so
- $\frac{V_1 - V_2}{R_1} = \frac{V_2}{R_2} = \frac{V_1}{R_{eq}}$ thus $\frac{R_1}{R_{eq}} V_1 = V_1 - V_2$ and $\frac{R_2}{R_{eq}} V_1 = V_2$. Adding, we get $\frac{R_1}{R_{eq}} V_1 + \frac{R_2}{R_{eq}} V_1 = V_1$. Dividing by V_1 and solving, we get $R_1 + R_2 = R_{eq}$

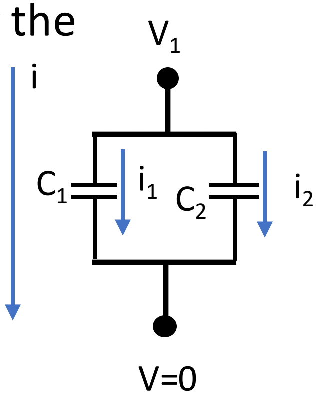


- Again let R_{eq} is the equivalent resistance, replacing the bottom left circuit with a single resistance.
- By Kirchhoff's node rule, $i_1 + i_2 = i$, so
- $\frac{V_1}{R_1} + \frac{V_1}{R_2} = \frac{V_1}{R_{eq}}$.
- Solving, we get. $\frac{R_1 R_2}{R_1 + R_2} = R_{eq}$

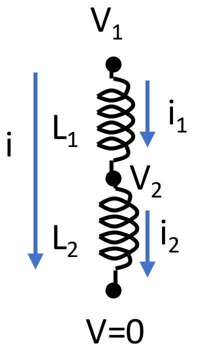
- C_{eq} is the equivalent capacitance, replacing the top right circuit with a single capacitor.
- By Kirchhoff's node rule, $i_1 = i_2 = i$, so
- $C_1 \frac{d\,(V_1 - V_2)}{dt} = C_2 \frac{d\,(V_2)}{dt} = C_{eq} \frac{dV_1}{dt}$
- $\frac{C_{eq}}{C_1} \frac{d\,(V_1)}{dt} = \frac{d\,(V_1 - V_2)}{dt}$ and $\frac{C_{eq}}{C_2} \frac{d\,(V_1)}{dt} = \frac{d\,(V_2)}{dt}$
- Adding and cancelling the $\frac{d\,(V_2)}{dt}$, we get
- $\frac{C_{eq}}{C_1} + \frac{C_{eq}}{C_2}$ =1 and solving, we get. $\frac{C_1 C_2}{C_1 + C_2} = C_{eq}$



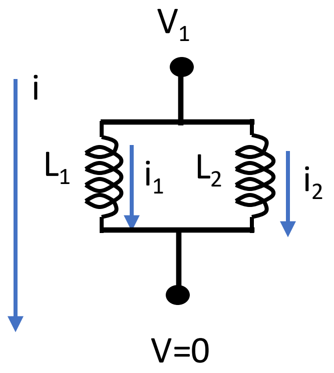
- C_{eq} is the equivalent capacitance, replacing the bottom right circuit with a single capacitor.
- By Kirchhoff's node rule, $i_1 + i_2 = i$
- $C_{eq} \frac{dV_1}{dt} = C_1 \frac{dV_1}{dt} + C_2 \frac{dV_1}{dt}$, so
- $C_{eq} = C_1 + C_2$



Simple circuit analysis with Kirchhoff

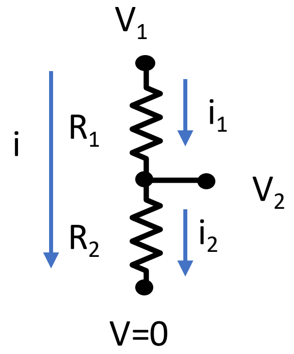


- Let L_{eq} be the equivalent inductance, replacing the top left circuit with a single inductor.
- By Kirchhoff's node rule, $i_1 = i_2 = i$, so
- $L_{eq}\frac{di}{dt} = V_1, L_1\frac{di_1}{dt} = V_1 - V_2, L_1\frac{di_2}{dt} = V_2$
- $V_1 = L_{eq}\frac{di}{dt} = L_1\frac{di}{dt} + L_2\frac{di}{dt}$ and
- $L_{eq} = L_1 + L_2$

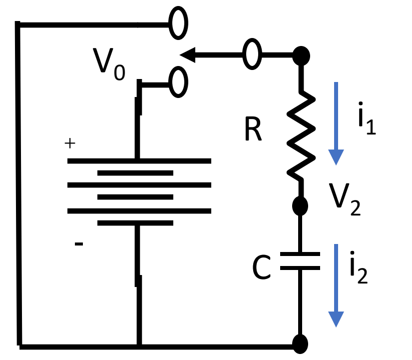


- Let L_{eq} be the equivalent inductance, replacing the bottom left circuit with a single inductor.
- $\frac{di}{dt} = \frac{V_1}{L_{eq}}, \frac{di_1}{dt} = \frac{V_1}{L_1}, \frac{di_2}{dt} = \frac{V_1}{L_2},$
- By Kirchhoff's node rule, $i_1 + i_2 = i$, so
- $\frac{V_1}{L_{eq}} = \frac{V_1}{L_1} + \frac{V_1}{L_2}$ and
- $L_{eq} = \frac{L_1 L_2}{L_1 + L_2}$

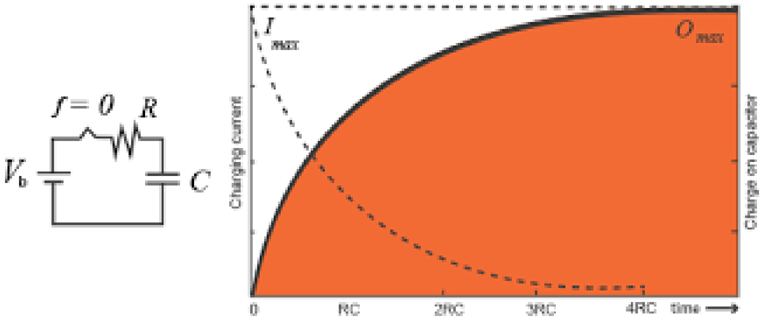
- The circuit on the right, is useful and is called a *voltage divider*.
- $i = i_1 = i_2$ so $\frac{V_1 - V_2}{R_1} = \frac{V_2}{R_2}, V_1 - V_2 = \frac{R_1}{R_2}V_2$
- Thus, $V_1 = (1 + \frac{R_1}{R_2})V_2$ and so
- $V_2 = \frac{R_2}{R_1 + R_2}V_1$

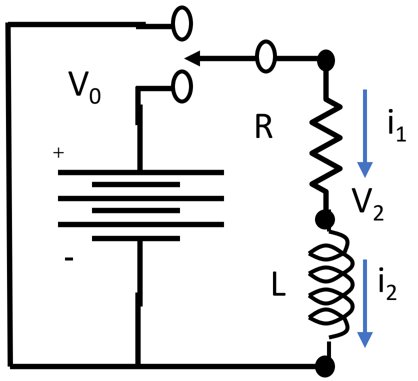


RC/RL circuit analysis with Kirchhoff

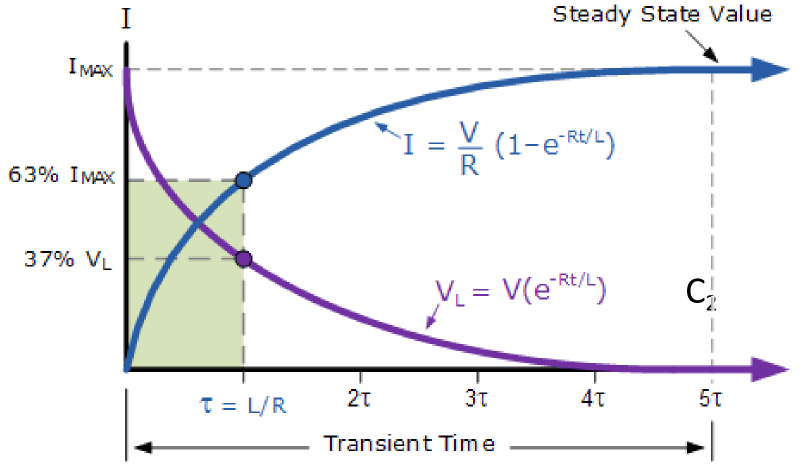


- RC behavior: charging
  - $V_0 - V_2 = i_1 R = V_R, i_1 = \frac{V_R}{R}$
  - $i_2 = C\frac{dV_2}{dt}, V_C = V_2$
  - $i_1 = i_2, V_C = V_0 - V_R$
  - $\frac{V_R}{R} = C\frac{dV_C}{dt}$, RC $\frac{dV_C}{dt} = V_0 - V_C$, or RC $\frac{dV_C}{dt} + V_C = V_0$
  - Solution is $V_C = V_0 - V_0 e^{-\frac{t}{RC}}$





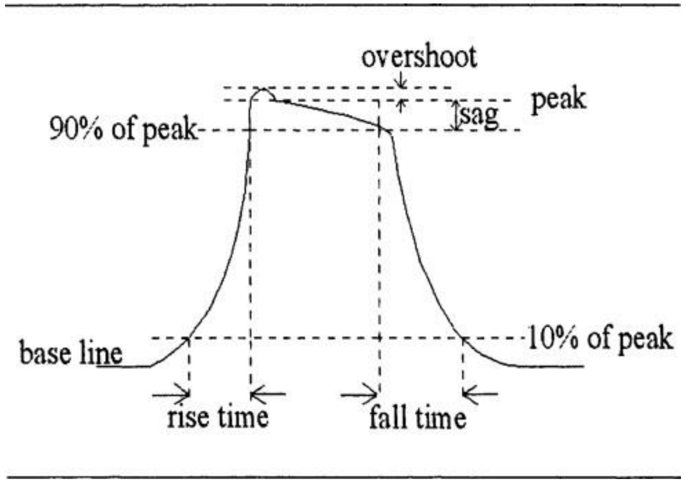
- RL behavior: charging
  - $V_0 - V_2 = i_1 R = V_R$
  - $V_L = V_2 = L\frac{di_2}{dt}$
  - $i_1 = i_2, V_R = V_0 - V_L$, so L $\frac{d}{dt}\frac{V_0 - V_L}{R} = V_L$
  - $\frac{L}{R}\frac{d\,V_L}{dt} + V_L = 0$
  - Solution is $V_L = V_0 e^{-\frac{Rt}{L}}$



Voltage responses

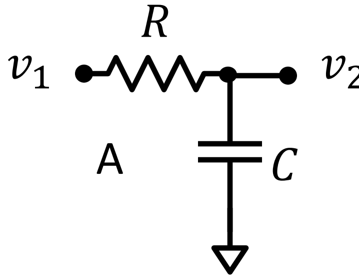
- Response to square wave with width T
  - A: $\tau = RC, \tau = T$
  - B: $\tau = RC, \tau = .1T$

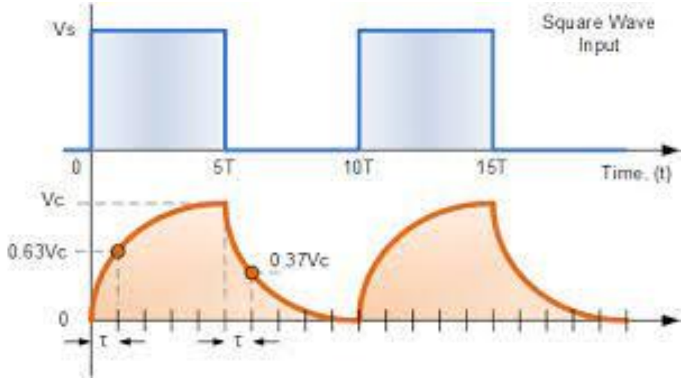
- Overshoot below



Low pass

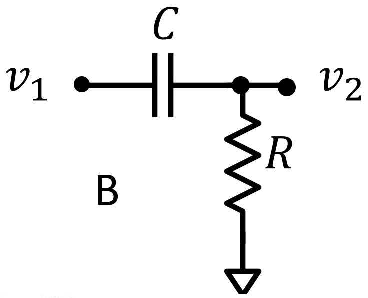
A

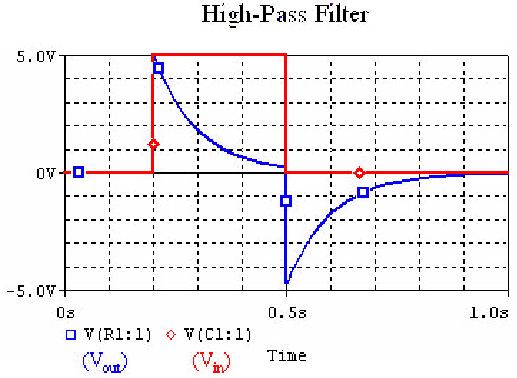




High pass

B





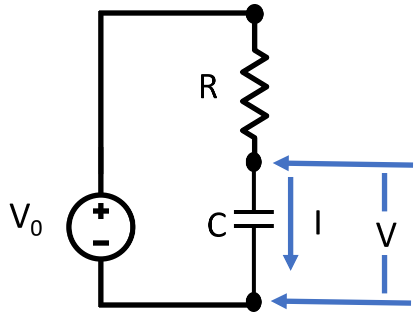
Phasors

- $V(t) = RI(t)$
- $V(t) = L\dot{I}(t)$
- $I(t) = C\dot{V}(t)$
- Suppose $V(t) = Acos(\omega t + \theta)$ and $\mathrm{I}(t) = Bcos(\omega t + \phi)$. If $\phi > \theta$, we say the current leads the voltage.
- $V(t) = Re(Ae^{j(\omega t+\theta)})$, and $\mathrm{I}(t) = Re(Be^{j(\omega t+\phi)})$
- Now define $\hat{V} = V = Ae^{j\theta}$ and $\hat{I} = Be^{j\phi}$, so $|V| = A$, $|I| = B$, $\angle V = \theta$, and $\angle I = \phi$. $\hat{V}$ and $\hat{I}$ are called phasors and do not include time. Note that $V(t) = Re(\hat{V}e^{j\omega t})$ and $\mathrm{I}(t) = Re\left(\hat{I}e^{j\omega t}\right)$.
- Note that $I = CVj\omega$, for a capacitor and $V = LIj\omega$, for an inductor
- $\hat{V} = Z\hat{I}, Z = R + jX$
- $\hat{I} = Y\hat{V}, Y = G + jB$

Circuit analysis and impedance

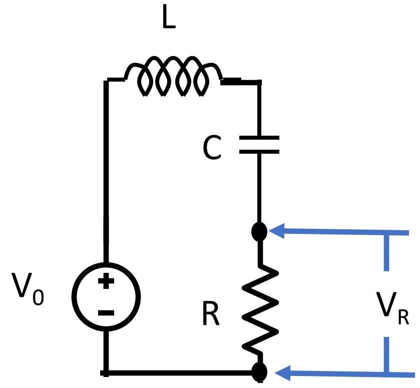
- Impedance unifies the “simple” ohms law with capacitance and inductance.
- $Z = R$, for resistors, $Z = j\omega L$, for inductors and $Z = \frac{1}{j\omega C}$, for capacitors.
- In general, $Z = R + jX$ and all the ohm like laws hold for resistors, capacitors and inductors .
    - $Z_{eq} = Z_1 + Z_2$ for two components with impedance Z_1, Z_2 connected in series
    - $Z_{eq} = \frac{Z_1 Z_2}{Z_1 + Z_2}$ for two components with impedance Z_1, Z_2 connected in parallel
- For example, for a resistor and capacitor in series has impedance $Z = R + \frac{1}{j\omega C}$

Phasors, impedance and power



- For the circuit on the right, $Z = R + \frac{1}{j\omega C}$ is the impedance for the resistor and capacitor in series.
- The phasor $I = \frac{V_0}{Z}$ and the phasor $V = \frac{I}{j\omega C} = \frac{V_0}{1+j\omega RC}$
- Further, $|I| = \frac{V_0}{|Z|}$, $\angle I = \angle\frac{V_0}{|Z|}$ and $|V| = \frac{|I|}{|j\omega C|} = |\frac{V_0}{1+j\omega RC}|$
- For phasors V, I, define the complex power as $P_{av} = \frac{V\bar{I}}{2} = Z\frac{I\bar{I}}{2} = \mathrm{R}\frac{|I|^2}{2} + j\mathrm{X}\frac{|I|^2}{2}$; the first term is the real power, the second is called the *reactive* power.
- The average power is $P_a = Re(P) = Re(\frac{V\bar{I}}{2})$. We define the reactive power as $P_r = Im(P)$.
- $P_r = \omega(E_L - E_C)$, where E_L and E_C are respectively, the energy stored in the inductor and capacitor respectively.

Q and phasors



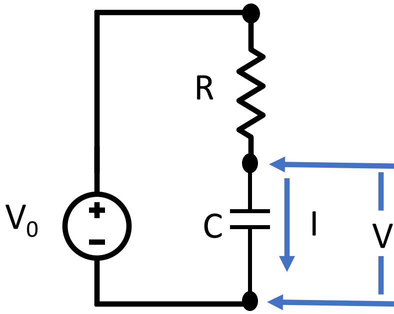
- Consider the series resonance on the right. $Z_{LCR} = R + j\left(\omega L - \frac{1}{\omega C}\right)$
- The phasor, $I = \frac{V_0}{Z_{LCR}}$, and the phasor $V_R = \frac{V_0}{Z_{LCR}} Z_R$, where $Z_R = R$.
- So $V_R = \frac{RC\omega V_0}{RC\omega + j(LC\omega^2 - 1)}$.
- $|V_R|$ is maximum when $\omega^2 LC = 1$. Put $\omega_0 = \frac{1}{\sqrt{LC}}$. When $\omega = \omega_0$, $|V_R| = V_R = V_0$.
- $|V_R| = \frac{V_0}{\sqrt{2}}$, when $X = R$. Note that the power through R when $X = R$ is half the power through R when $X = 0$ or $\omega = \omega_0$.
- Let the frequencies where $R = \pm X$ be denoted $\omega_u$and ω_l, where $\omega_u > \omega_l$.
- We define $Q = \frac{X}{R} = \frac{\omega_0 L}{R} = \frac{1}{\omega_0 CR}$.
- Solving for $\omega_u$and ω_l, we get $\frac{L\omega_u}{\omega_0} - \frac{\omega_0}{C\omega_u} = R$ and $\frac{L\omega_l}{\omega_0} - \frac{\omega_0}{C\omega_l} = -R$, or, in terms of Q,
- $\frac{\omega_u}{\omega_0} - \frac{\omega_0}{\omega_u} = \frac{1}{Q}$ and $\frac{\omega_l}{\omega_0} - \frac{\omega_0}{\omega_l} = -\frac{1}{Q}$. In fact, $\omega_0 = \sqrt{\omega_u \omega_l}$, and so $\frac{\omega_u}{\omega_0} - \frac{\omega_l}{\omega_0} = \frac{1}{Q}$.
- Thus $Q = \frac{\omega_0}{\omega_u - \omega_l} = \frac{\omega_0}{\Delta\omega}$
- From the definition of P_a, earlier, $Q = \omega_0 \frac{E}{P_a}$, where E is the total energy stored in *L* and *C*, which is in turn the peak E_L and peak E_C at resonance.

Resonance and Q

- Series Resonance
  - At ω_u and ω_l, $X = \pm R$ [ω_u is upper 3dB cutoff and ω_l is lower 3dB cutoff]
  - $\omega_u L - \frac{1}{\omega_u C} = R$, $\omega_l L - \frac{1}{\omega_l C} = -R$
  - Define $Q = \frac{X}{R}$
  - $\frac{\omega_u}{\omega_0} - \frac{\omega_0}{\omega_u} = \frac{R}{\omega_0 L} = \frac{1}{Q}$ and $\frac{\omega_l}{\omega_0} - \frac{\omega_0}{\omega_l} = -\frac{R}{\omega_0 L} = -\frac{1}{Q}$
  - $\frac{\omega_u}{\omega_0} - \frac{\omega_0}{\omega_u} = \frac{\omega_0}{\omega_l} - \frac{\omega_l}{\omega_0}$, so ${\omega_0}^2 = \omega_u \omega_l$ and $\frac{\omega_u - \omega_l}{\omega_0} = \frac{1}{Q}$
- Parallel Resonance
  - $\frac{\omega_u}{\omega_0} - \frac{\omega_0}{\omega_u} = \frac{G}{\omega_0 C} = \frac{1}{Q_p}$ and $\frac{\omega_l}{\omega_0} - \frac{\omega_0}{\omega_l} = -\frac{G}{\omega_0 C} = -\frac{1}{Q_p}$

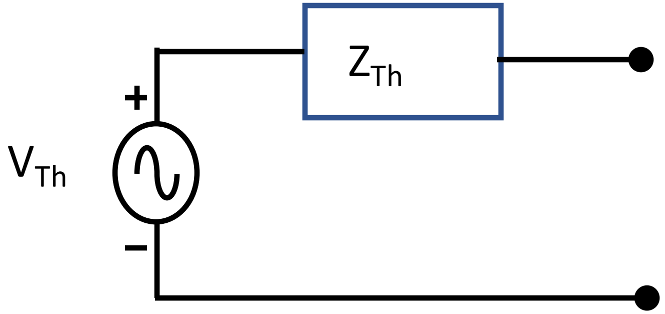
Phasors, impedance and power

- For the circuit on the right, $Z = R + \frac{1}{j\omega C}$ is the impedance for the resistor and capacitor in series.
- The phasor $I = \frac{V_0}{Z}$ and the phasor $V = \frac{I}{j\omega C} = \frac{V_0}{1+j\omega RC}$
- Further, $|I| = \frac{V_0}{|Z|}$, $\angle I = \angle \frac{V_0}{|Z|}$ and $|V| = \frac{|I|}{|j\omega C|} = |\frac{V_0}{1+j\omega RC}|$
- Complex power: $P = \frac{V\bar{I}}{2} = Z\frac{|I|^2}{2} = P_a + jP_r = R\frac{|I|^2}{2} + \mathrm{j}X\frac{|I|^2}{2}$
  - P_a is power delivered to resistor, P_r is power stored in inductor
  - For phasors V, I, define the complex power as $P_{av} = \frac{V\bar{I}}{2} = Z\frac{I\bar{I}}{2} = \mathrm{R}\frac{|I|^2}{2} + j\mathrm{X}\frac{|I|^2}{2}$; the first term is the real power, the second is called the *reactive* power.
- $P_r = \omega(E_L - E_C)$, where $E_L$and E_C are respectively, the energy stored in the inductor and capacitor respectively.
- $P_r = \frac{\omega L|I|^2}{2} - \frac{\omega C|V_c|^2}{2} = \omega(E_L - E_C)$
- $Q = \omega\frac{L|I|^2}{R|I|^2} = \omega\frac{L}{R} = \omega\frac{E_L}{P_a}$

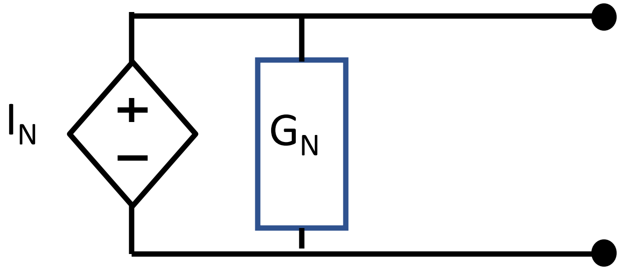


Thevenin and Norton

- **Thevenin:** Any combination of *linear* sources and passive elements terminating in two terminals is equivalent to a pure voltage source in series with an impedance

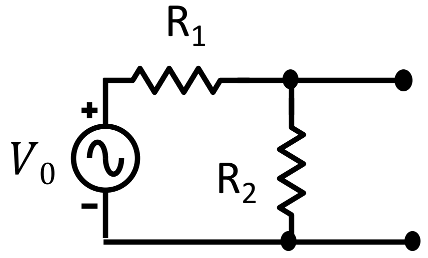


- **Norton:** Any combination of *linear* sources and passive elements terminating in two terminals is equivalent to a pure current source in parallel with a conductance

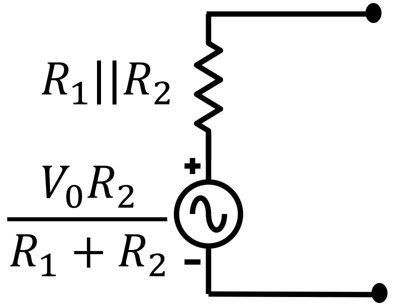


- Above, $G_N = \frac{1}{Z_{Th}}$
- Similar theorems for *linear* two terminal input and output devices (with transfer function)

Thevenin and Norton

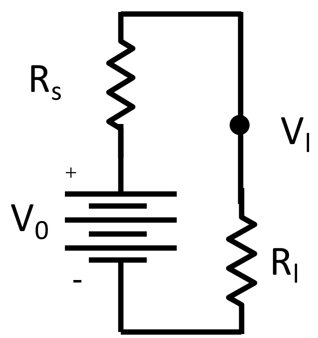


is Thevinn equivalent to



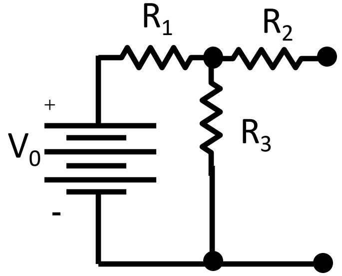
- We can use lookback resistance to calculate the Thevenin equivalent resistance and ideal source.
- To find the lookback resistance, short the source and apply the usual laws.
  - Here $R_s = R_1||R_2$
- To find the new ideal source, notice R_1 and R_2 form a voltage divider.
  - The new source voltage is $\frac{V_0 R_2}{R_1+R_2}$
- In general, a Norton equivalent with parameters (i_N, Z_N) is the same as a Thevenin equivalent with parameters (V_{Th}, Z_{Th}) with $Z_{Th} = Z_N$ and $V_{Th} = i_N Z_N$

Exercise 1: Resistors

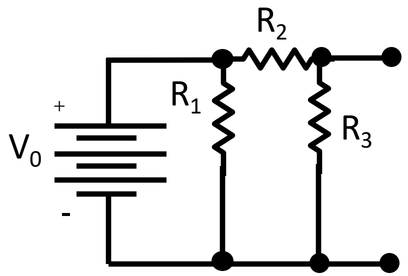


A

1. Consider (A). Find the formula for power in the load. Find the R_l that maximizes the power to the load.
   - $V_l = \frac{R_l}{R_s+R_l}V_0$, $I_l = \frac{V_0}{R_s+R_l}$.
   - $P_l = V_l I_l = \frac{R_l}{(R_s+R_l)^2}{V_0}^2$, which is maximum when $R_l = R_s$
2. Find the Thevenin and Norton parameters for (B).
   - $V_{Th} = \frac{R_3}{R_1+R_3}V_0$
   - $R_{Th} = R_2 + R_1||R_3$
3. Find the Thevenin and Norton parameters for (C).
   - $V_{Th} = \frac{R_3}{R_2+R_3}V_0$
   - $R_{Th} = R_2||R_3$



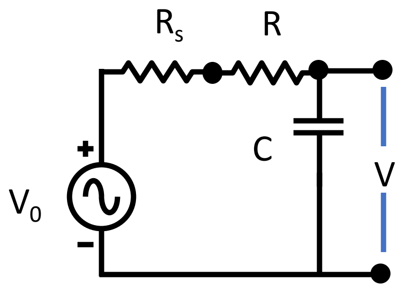
B

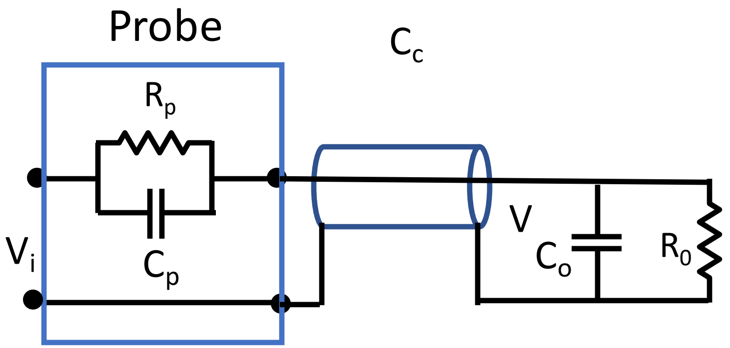


C

Exercise 3: Capacitors

1. In the circuit on the right, V_0 is a 2 volt pp ideal square wave source of frequency 20Hz, $R_S = 50\Omega$, $R = 300k\Omega$ and $C = 10\ nF$. Period is $50\ millisec$
2. What is the voltage, V, at the output?
   - About a volt at peak
3. Let t_2 be the time to discharge to 0V. Calculate τ and t_2.
   - $\tau = RC = \ 3 \times 10^5 \times 10^{-8}\ sec = 3\ millisec$
   - $t_2 = \tau \ln(2) \approx 2ms$ (t_2 is the time to decay to $\frac{1}{2}$ its value)
4. Capacitance on the scope prevents the delay from being 0. Measure the new t_2 with these changes. Given
   - $C_c = 100pf/m, C_o = 50pF, C_p = 10pF, R_0 = 10^6\Omega$
5. Calculate the new t_2.
   - $\tau = 6\mu$-sec





Signals and modulation

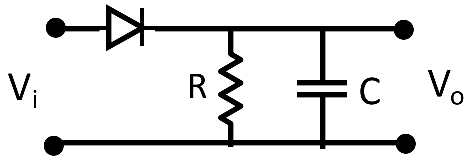
- Gain: $G = \frac{P_o}{P_i}$, Loss: $L = \frac{P_o}{P_{max}}$, Rejection: $\mathrm{R} = \frac{P_{max}}{P_{pb}}$. Gain (G) expressed in decibels: $G = 10\ log_{10}(^{P_{out}}/_{P_{in}})$
- $P_s = \int \sigma\, E \cdot E + \epsilon \frac{E \cdot E}{2} + \frac{H \cdot H}{2\mu}\ \mathrm{dV} + \int E \times H\, dA$
- Mixer: $V(t) = \cos(\omega_1 t)\cos(\omega_2 t) = \frac{1}{2}[\cos(\omega_+ t) + \cos(\omega_- t)]$, $\omega_+ = \omega_1 + \omega_2$, $\omega_- = \omega_1 - \omega_2$
- Modulation

Name	Equation
AM	$V(t) = (1 + am(t))V_c\cos(\omega_c t)$
FM	$V(t) = V_c\cos([\omega_c + am(t)]t)$
Angle	$V(t) = V_c\cos\left(\omega_c t + \phi(t)\right), \phi(t) = am(t)$. [Like FM]
FSK	$V(t) = V_c\cos(\omega_1 t)$, if 1 $V(t) = V_c\cos(\omega_0 t)$, if 0
PSK	$V(t) = +V_p\cos(\omega t)$, if 1 $V(t) = -V_p\cos(\omega t)$, if 0

Exercise 4: Diode detectors

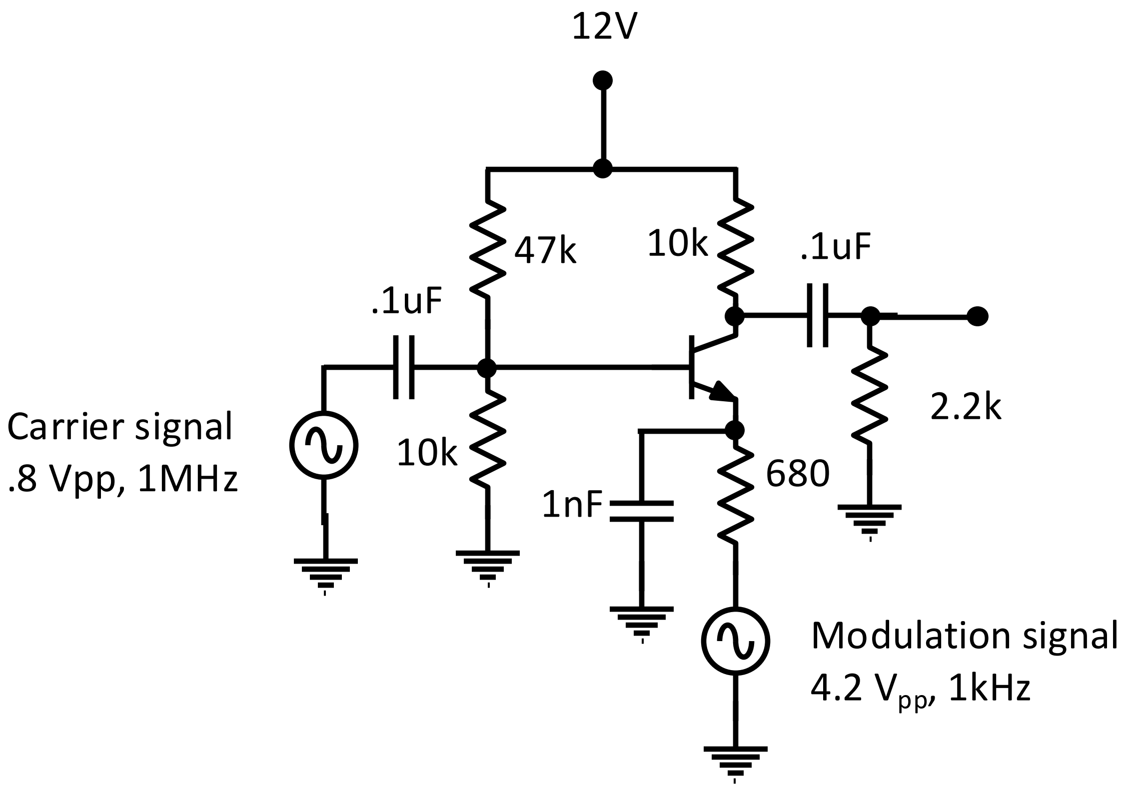
- For AM, $V(t) = V_c \cos(\omega_c t) + a(t)\cos(\omega_c t)$, Define the modulation depth $m = \frac{a_p}{V_c}$
- In circuit on the right, $R = 10k\Omega$, $C = 10\ nF$
- Set function generator for $f_c = 1MHz$, $V_{c,pp} = 5V$, $f_m = 1kHz$, $m = .7$

1. Calculate τ for the RC circuit. $\tau = 10^4 \times 10^{-8}\ \mathrm{sec} = .1ms$.
   - T_m is period of modulating signal. $T_m = 10^{-3}sec = 1ms$. So $\tau \ll T_m$
   - T_c is period of modulating signal. $T_c = 10^{-6}sec = 1\mu s$. $\tau \gg T_c$
   - As you change f_m does the frequency of V_o track it? (It better)
2. Compare the max voltage of the AM signal to the max of V_0.
   - $V_0, p \approx .8V, V_{i,p} \approx 1.4V$
3. What happens when we make $m = 1.0$

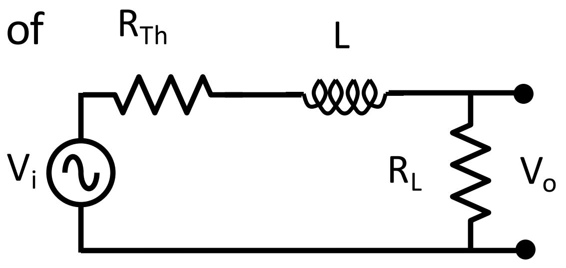


AM Modulator for previous exercise

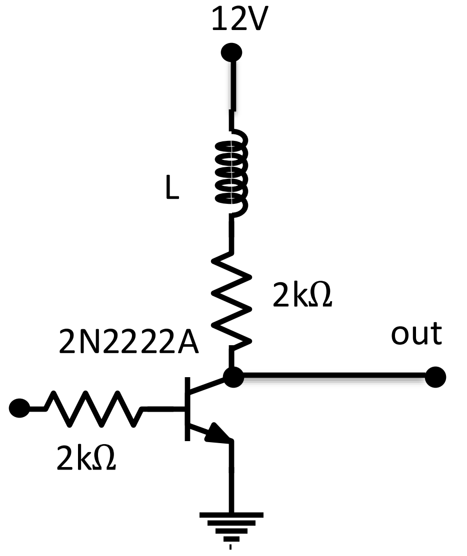
- I didn't have a signal generator that produced an AM signal, so I used the modulator on the right with the indicated inputs to produce the AM needed for the detector in the previous exercise.



Exercise 5: Inductors



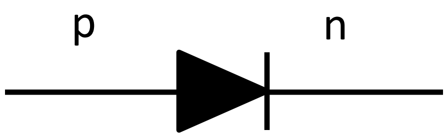
- Set function generator for square wave with 5V V_{pp}, a Thevenin equivalent source resistance of $R_{Th} = 50\Omega$ and frequency $1kHz$. Connect a load, $R_L = 100\Omega$ load, L=1mH
  1. Observe square wave with rounded corners, measure the time, t_2 to decay to 0
     - About $20\mu sec$
  2. In the top circuit, calculate inductor current and the expected delay, t_2
     - $Z_{eq} = 150 + jL\omega$, $\omega = 2\pi \times 10^3$, $V_i = Re(V_{i,p}e^{j\omega t})$
     - As phasors, $iZ_{eq} = V_i$, $|i|\sqrt{150^2 + (\omega L)^2} = V_{i,p}$, $|i| = \frac{V_{i,p}}{\sqrt{150^2+(2\pi)^2}}$, $\theta = \angle i = \arctan(-\frac{2\pi}{150})$, $\theta \approx -2.4\, rad = -15°$
     - $V_o = Re\left(\frac{100V_{i,p}}{\sqrt{150^2+(2\pi)^2}} e^{j(\omega t+\theta)}\right)$, $|V_o| = 1.6V$,
     - $\tau_{RL} = \frac{10^{-3}}{100} sec \approx 10\ \mu sec$
  3. In the second circuit, use 2 scope channels: one at input, one at output.
     - 1μ sec $rise\ time$. Ringing at 10MHz. $\frac{1}{\sqrt{LC}} = 62.8 \times 10^6$. $C = \frac{10^3}{(62.8\times10^6)^2} \approx\ .25pF$
- Note: I made the pull-up 100K.



Diodes and bipolar small signal models

- Diode model:
  - $i_d = i_s(\exp\left(\frac{eV_d}{kT}\right) - 1), \frac{e}{kT} = 40V^{-1}$ at room temperature
  - T is the junction temperature
  - $i_s \approx 10^{-12} amp$
  - V_d is voltage across diode

- $r_d = \frac{e}{kTi_d}$
- When i_d is a few nano-Amps, $r_d \approx 5\Omega$
- When i_d is a few μA, $r_d \approx 10^4\Omega$



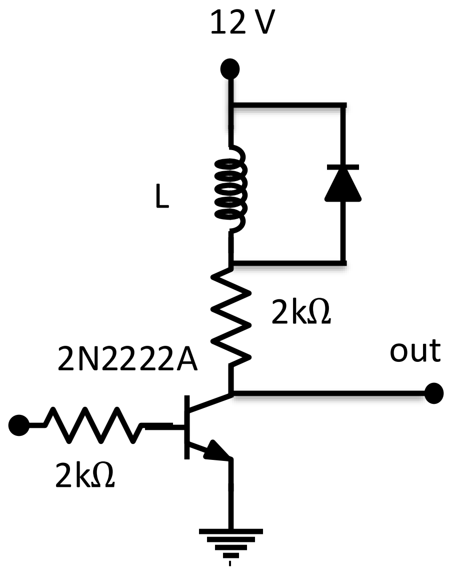
- Transition from p to n in $1\ \mu m$
- Doping provides
  - n side has excess free electrons
  - p side has excess holes

Exercise 6: Diodes and snubbers

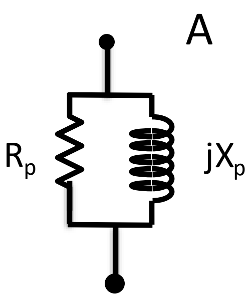
- Add indicated snubber diode.

1. Swing up is nearly immediate with snubber
2. Ringing at 10MHz. $\frac{1}{\sqrt{LC}} = 62.8 \times 10^6$. $C = \frac{10^3}{(62.8 \times 10^6)^2} \approx .25pF$
3. What is its effect on ringing?
   - Ringing is uniform at 5 MHz
4. Diode should be on when transistor is off.

- Note: I made the pull-up 100K.

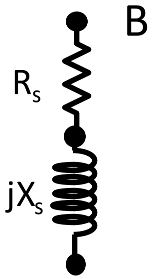


Exercise 7: Parallel to series conversion

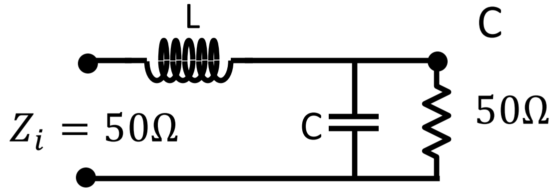


- For series: $Z_s = R_s + j\omega L$, $Q_s = \frac{X_s}{R_s}$
- For parallel: $\frac{1}{Z_p} = \frac{1}{R_p} + \frac{1}{j\omega L}$, so $Z_p = \frac{j\omega L R_p}{R_p + j\omega L}$ and $Q_p = \frac{R_p}{X_p}$
- If $Q_p = Q_S$, $X_p X_s = R_p R_S$

1. If circuits (A) and (B) have the same impedance, what is the relationship between R_p, X_p and R_s, X_s?
   - $\frac{1}{Z_p} = \frac{1}{R_p} + \frac{1}{jX_s}$, $Z_p = \frac{jR_pX_p}{R_p + jX_p} = \frac{{X_p}^2 R_p + j{R_p}^2 X_p}{{R_p}^2 + {X_p}^2}$, $Z_s = R_s + jX_s$
   - $R_s = \frac{{X_p}^2 R_p}{{R_p}^2 + {X_p}^2}$, $X_s = \frac{{R_p}^2 X_p}{{R_p}^2 + {X_p}^2}$, $R_s = X_p \frac{X_p R_p}{{R_p}^2 + {X_p}^2}$, $X_s = R_p \frac{X_p R_p}{{R_p}^2 + {X_p}^2}$, set $\rho = \frac{X_p R_p}{{R_p}^2 + {X_p}^2}$ for later reference
   - This shows the Q's must be equal as stated above.



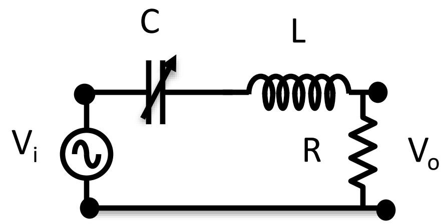
2. Find a formula for X_s, for large $Q = Q_p = Q_s$ and small $Q = Q_p = Q_s$
3. Use circuit (C) to transfer a 50Ω load (circuit C) to a 5Ω load. What is X_C at 7 MHz? What are C and L in that model?
   - Use the parallel to series conversion to make a series equivalent circuit consisting of C and the 50Ω with $R_s = 5\Omega$



Exercise 8: Series resonance

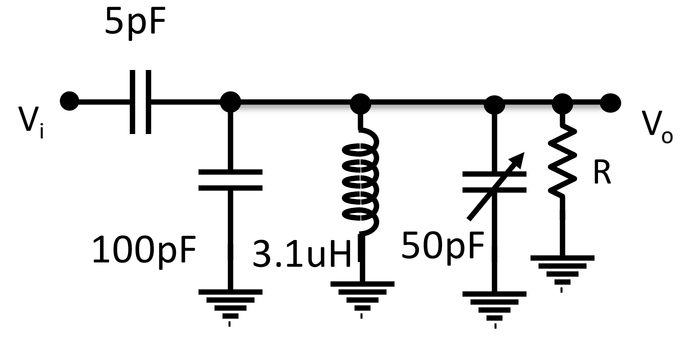
- For the circuit on the right, C$= 8 - 50pf$, $L = 15\mu H$ forming a bandpass filter. $R = 100\Omega$
- If $C = 34pf$, the resonant frequency is $\omega = \frac{1}{\sqrt{35\times10^{-12}\times15\times10^{-6}}} = \frac{10^9}{\sqrt{525}} \approx 44.2$, so the resonant frequency is $\frac{44.2}{2\pi} \approx 7.07MHz$

1. Tune the resonant frequency to $7MHz$ and find f_u, f_l and Δf and thus Q.
   - $f_u = 7.67MHz, f_l = 6.47MHz, Q = \frac{f}{\Delta f} = \frac{7}{1.2} = 5.8$
2. Compute what these values should be
   - $Z_{eq} = R + j(\omega L - \frac{1}{\omega C})$
   - As phasors, $i = |i|e^{j\theta}, |i| = \frac{V_{i,0}}{\sqrt{R^2+(\omega L-\frac{1}{\omega C})^2}}, \theta = -arctan(\frac{\omega L-\frac{1}{\omega C}}{R})$
   - $V_0 = iR$, Power through R at ω is $P(\omega) = |i^2|R = \frac{{V_{i,0}}^2 R}{R^2+(\omega L-\frac{1}{\omega C})^2}$. At resonance, $P(\omega_r) = \frac{{V_{i,0}}^2}{R}$. To find half power, set $\frac{1}{2} = (\frac{{V_{i,0}}^2 R}{R^2+(\omega L-\frac{1}{\omega C})^2})/(\frac{{V_{i,0}}^2}{R})$, or $\mathrm{R} = \omega L - \frac{1}{\omega C}$.
   - Solving gives $f_u = 7.67MHz$, $f_l = 6.53MHz$, $Q = 6.1$
   - General formulas: $\omega_u = \frac{R}{2L} + \sqrt{(\frac{R}{2L})^2+\frac{1}{LC}}$, $\omega_l = -\frac{R}{2L} + \sqrt{(\frac{R}{2L})^2+\frac{1}{LC}}$



Exercise 9: Parallel resonance

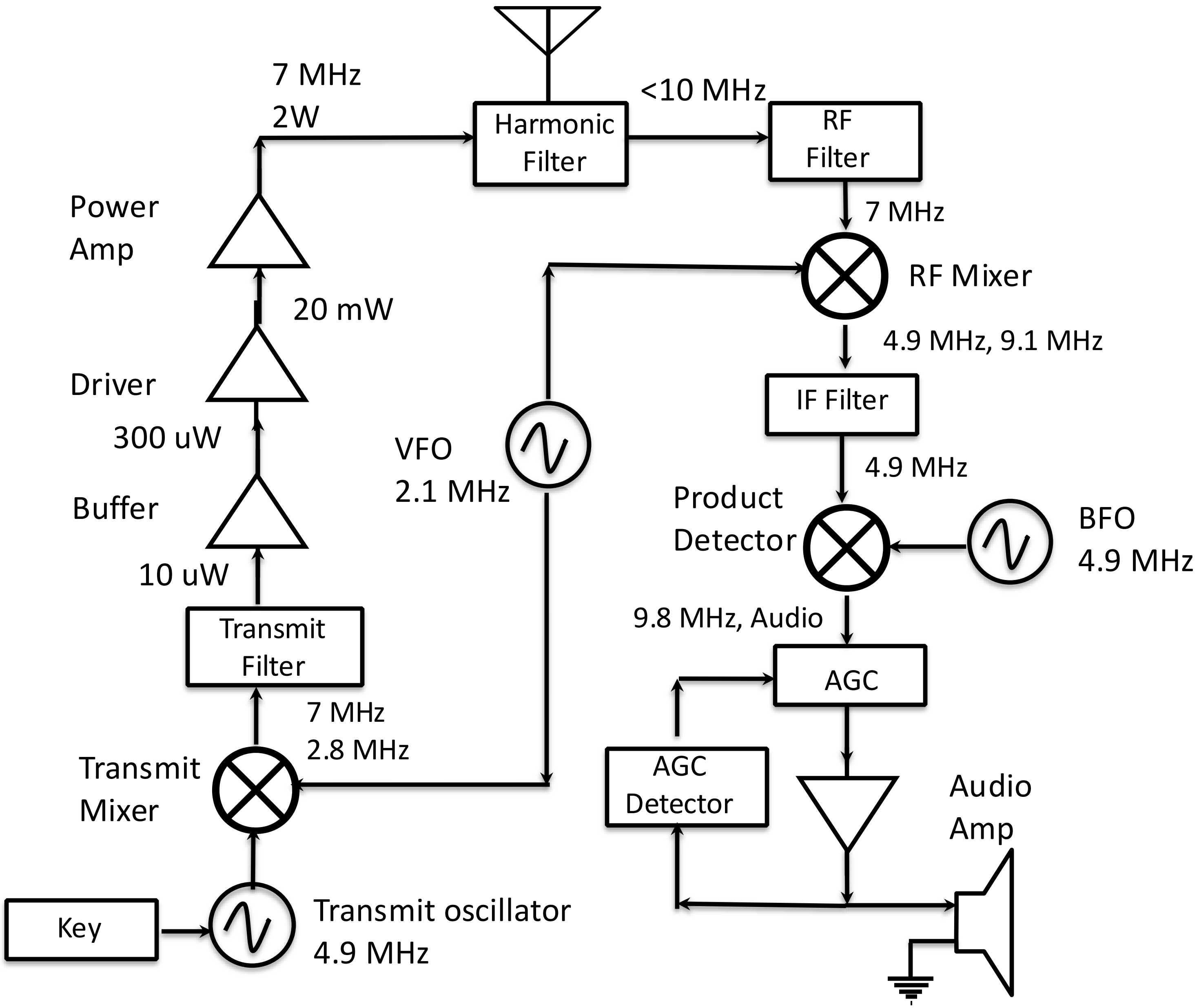
- $L = A_l N^2, A_l = 4 \frac{nH}{turn^2}$ for T37-2 core so for 28 turns, $L = 3.1\mu H$



1. Find the resonant frequency, the frequencies corresponding to a 3db falloff, the bandwidth and the Q of this circuit. This circuit is in the transmit oscillator.
   - At tuned resonance (7MHz), effective capacitance is about 167 pF
   - $Q_p = \omega_0 RC$
   - For $R = 1500\Omega$, network: $Q = 1500 \times 44 \times 10^6 \times 1.67 \times 10^{-10} = 11$
   - $BW = \frac{f_r}{Q} = \frac{7MHZ}{11} = .636MHz.\ f_u = f_r + \frac{BW}{2} = 7.318MHz, f_l = f_r - \frac{BW}{2} = 6.682MHz.$ This is 3dB cutoff.

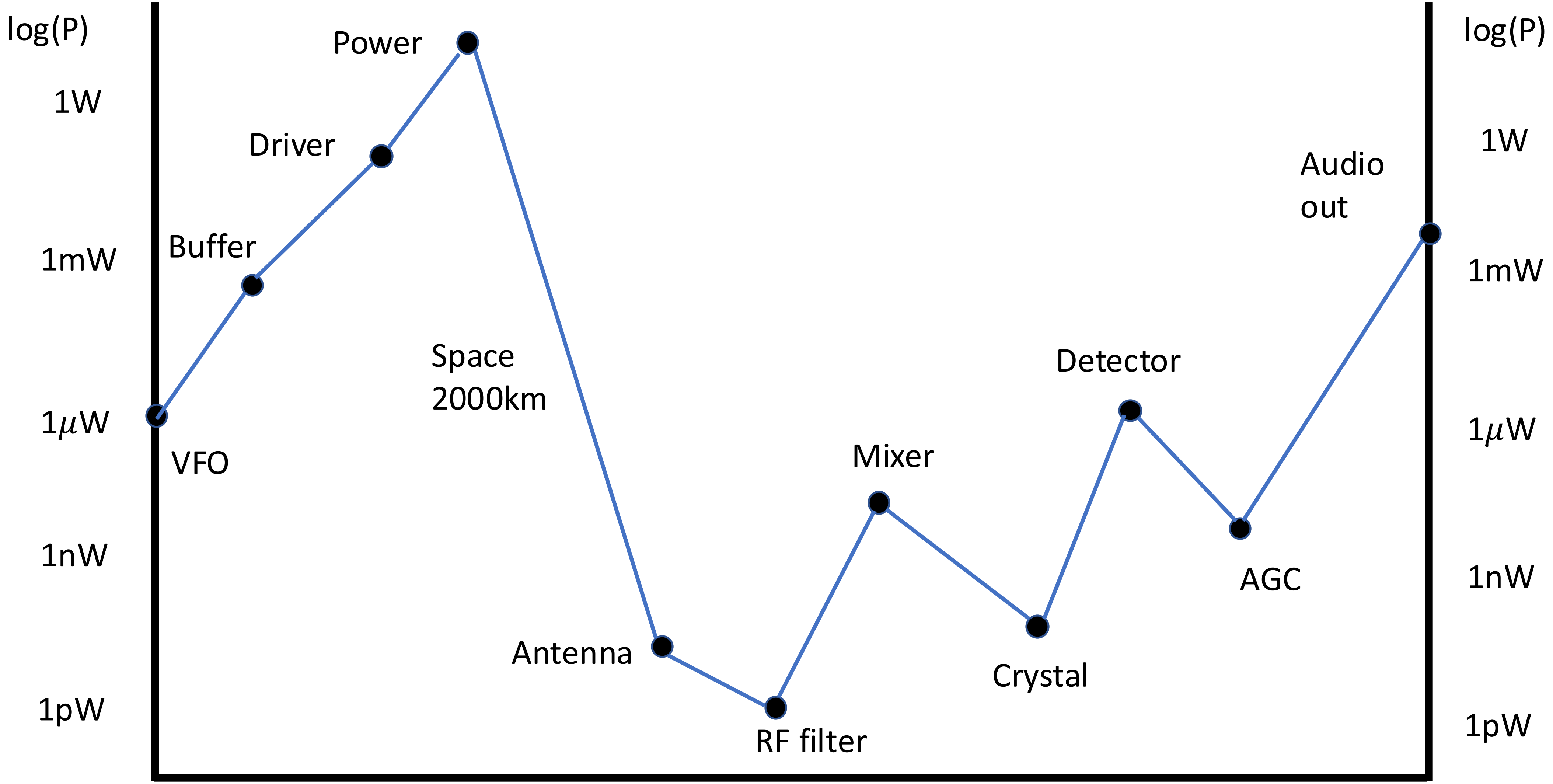
- General formulas: $BW = \frac{f_r}{Q}, f_u = f_r + \frac{BW}{2}, f_l = f_r - \frac{BW}{2}$

Norcal 40A

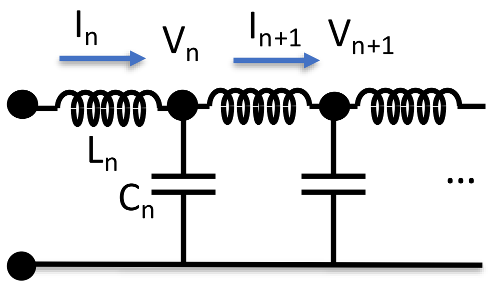


7 MHz
2W
Harmonic Filter
<10 MHz
RF Filter
7 MHz
RF Mixer
Power Amp
20 mW
4.9 MHz, 9.1 MHz
Driver
IF Filter
300 uW
VFO
2.1 MHz
4.9 MHz
Product Detector
BFO
4.9 MHz
Buffer
10 uW
9.8 MHz, Audio
Transmit Filter
AGC
7 MHz
2.8 MHz
Transmit Mixer
AGC Detector
Audio Amp
Key
Transmit oscillator
4.9 MHz

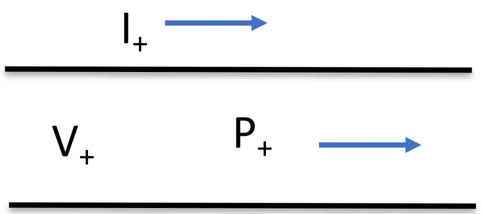
NorCal power levels

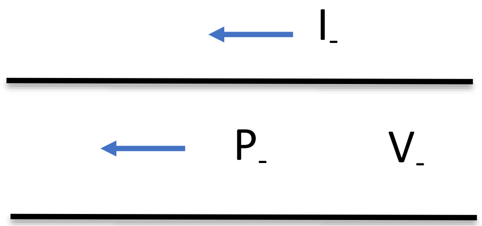


Transmission Lines

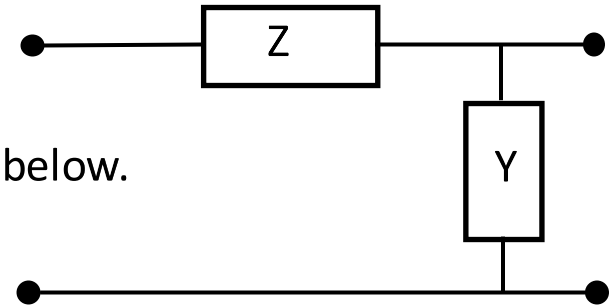


- $V_{n+1} - V_n = -L_l \frac{\partial I_{n+1}}{\partial t}$, $L = \frac{L_l}{l}$, so $\frac{\partial V}{\partial z} = -L \frac{\partial I}{\partial t}$
- $I_{n+1} - I_n = -C_l \frac{\partial V_n}{\partial t}$, $C = \frac{C_l}{l}$, so $\frac{\partial I}{\partial z} = -C \frac{\partial V}{\partial t}$
- Thus, $\frac{\partial^2 V}{\partial z^2} = LC \frac{\partial^2 V}{\partial t^2}$ and $\frac{\partial^2 I}{\partial z^2} = LC \frac{\partial^2 I}{\partial t^2}$, whose solution is is $V(z \pm vt)$
- $v = \frac{1}{\sqrt{LC}}$ for the forward wave .
- $\frac{\partial V}{\partial z} = -L \frac{\partial I}{\partial t}$ and $\frac{\partial I}{\partial z} = -C \frac{\partial V}{\partial t}$ implies
- $V' = vLI'$ and $\frac{V}{I} = \sqrt{\frac{L}{C}}$, so $Z_0 = \sqrt{\frac{L}{C}}$, where Z_0 is the forward impedance
- $Z_0 = \frac{V_+}{I_+}$, $-Z_0 = \frac{V_-}{I_-}$, $V = V_+ + V_-$, $-Z_0$ is the backwards looking impedance
- $P_+(t) = \frac{{V_+}^2}{Z_0}$, $P_-(t) = -\frac{{V_-}^2}{Z_0}$ (the negative sign implies energy flows to the left)





General transmission line

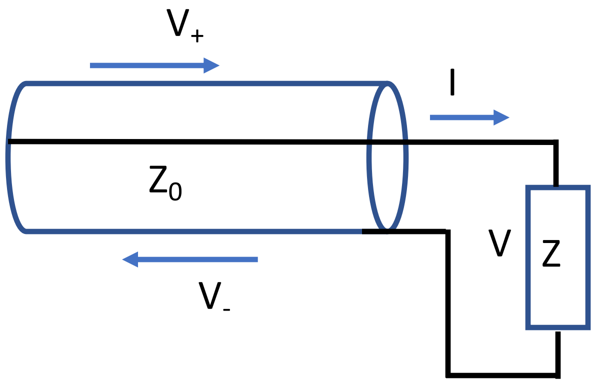


- $V(z - vt) = Acos(\omega t - \beta z), v = \frac{\omega}{\beta}$. The phasor is $\hat{V} = Ae^{-j\beta z}$, although we drop the cap below.
- Now the forward and backward voltage phasors are $V_+ = Ae^{-j\beta z}, V_- = Ae^{j\beta z}$
- The complex power is $P_{av} = \frac{V\bar{I}}{2}, P_+ = \frac{V_+\overline{I_+}}{2} = \frac{|V_+|^2}{2Z_0}, P_- = \frac{V_-\overline{I_-}}{2} = -\frac{|V_-|^2}{2Z_0}$, with $\frac{V}{I} = Z_0$
- Suppose over a transmission line, Z is the distributed impedance/m, Y is the distributed admittance/m and suppose the forward wave is $Ae^{j(\omega t - kz)}$, with phasor is $V = Ae^{-jkz}$.
- Let $Z = \frac{V}{I}$ then $\frac{dV}{dz} = -ZI, \frac{dI}{dz} = -YV$.
- Put $jk = \alpha + \beta j$ (to account for attenuation), then $jk = \sqrt{ZY}$ and the forward phasor becomes $e^{(-\alpha z - j\beta z)}$. $\alpha_{dB/m}$ is a transmission loss. $\alpha_{dB/m} = 8.686\alpha_{nepers/m}$.
- By differentiating , we get $jkV = ZI, -jkI = YV$. Solutions are $jk = \sqrt{ZY},\ Z_0 = \frac{V}{I} = \sqrt{\frac{Z}{Y}}$, all complex
- So, if $Z = R + j\omega L, Y = j\omega C + G$ for the transmission line, then $jk = \sqrt{(j\omega L + R)(j\omega C + G)}$ and $Z_o = \sqrt{\frac{(j\omega L+R)}{(j\omega C+G)}}$ (positive real root). (See the coax slide)

Transmission Lines - dispersion

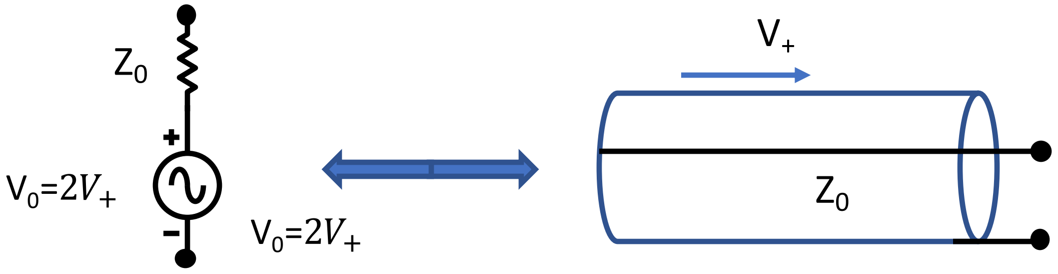
- α and v can vary with frequency; this is dispersion.
- Heaviside: Adjust parameters so $\frac{R}{L} = \frac{G}{C}$, then α doesn't depend on v and we get:
  - $jk = j\omega\sqrt{LC}(1 + \frac{R}{j\omega L})$ and $v = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}}$, $\alpha = \sqrt{RG}$
  - We also get $Z_0 = \sqrt{\frac{L}{C}}$ as with a lossless line.
  - If $\omega L \gg R$
    - $G = 0$ and $Z_0 = \sqrt{\frac{(j\omega L+R)}{j\omega C}} \approx \sqrt{\frac{L}{C}}$
  - If $\mathrm{R} \gg \omega L$
    - $jk = \sqrt{\frac{(j\omega L+R)}{j\omega C}} \approx \sqrt{j\omega RC}$, and $\alpha = \sqrt{\frac{\omega RC}{2}}$, $\alpha = \sqrt{\frac{2\omega}{RC}}$
- For first transatlantic cable, $L = 460\frac{nH}{m}, C = 75\frac{pF}{m}, f = 12Hz, R = 7\frac{m\Omega}{m}, l = 3600\ km, \alpha = \sqrt{\frac{\omega RC}{2}}$ $=4.4 \times 10^{-3}\ {}^{nepers}/_{m}, \alpha l = 140dB$
  - $\alpha l \approx 140dB$ and highly dispersive

Transmission Lines-reflections



(A)

- Now we look at the end of the transmission line and define $\rho = \frac{V_-}{V_+}$, and $\rho_i = \frac{i_-}{i_+} = -\rho$
- $V = V_+ + V_- = (1+\rho)V_+$
- $\tau = \frac{V}{V_+} = 1 + \rho = \frac{2Z}{Z+Z_0}$, $V = 2V_+$
- Consider the circuit in the upper right (A): $V = V_+ + V_-, I = I_+ + I_-,\ Z = \frac{V}{I}$
- $Z = \frac{V}{I} = \frac{V_+ + V_-}{I_+ + I_-}, Z_0 = \frac{V_+}{I_+}$
- $\frac{Z}{Z_0} = \frac{1+\rho}{1-\rho}, \rho = \frac{Z-Z_0}{Z+Z_0},\ \rho_{open-circuit} = 1.$

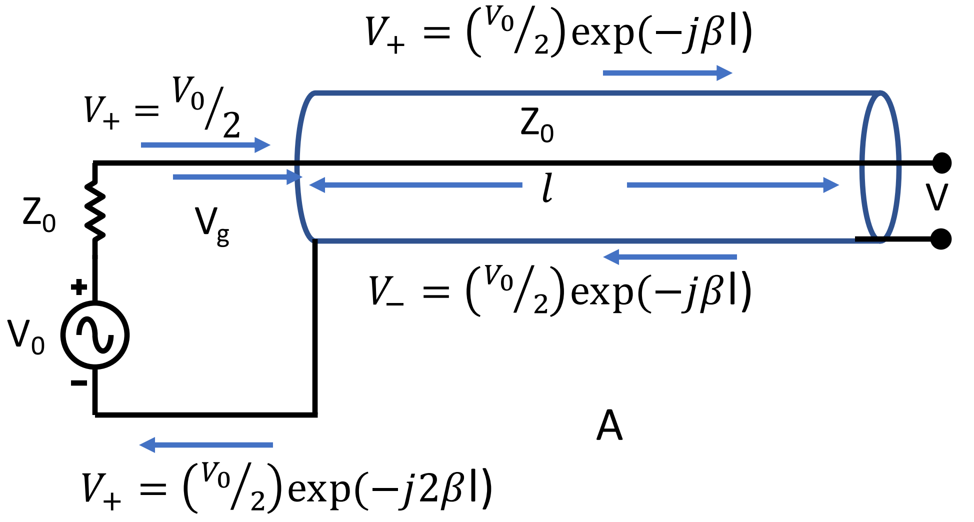


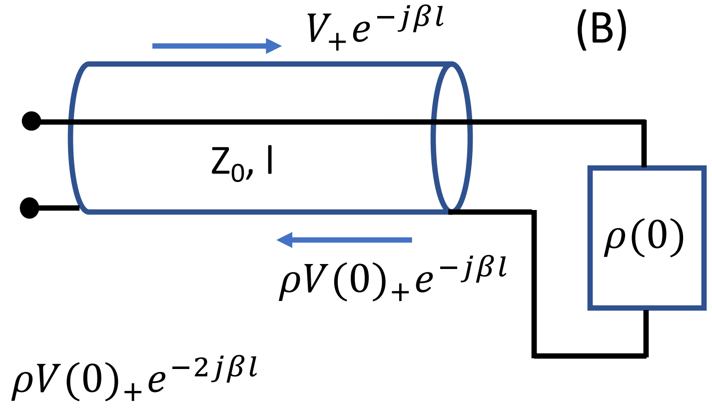
(B)

- For (B):
- Lookback resistance is $R_s = Z_0$, short circuit for (B) is $i_s = \frac{V_0}{Z_0}$
- Thevenin equivalent for open circuit is (B)
- $P_+ = \frac{{V_+}^2}{2Z_0} = \frac{{V_0}^2}{8R_s}$. This is the total available power.

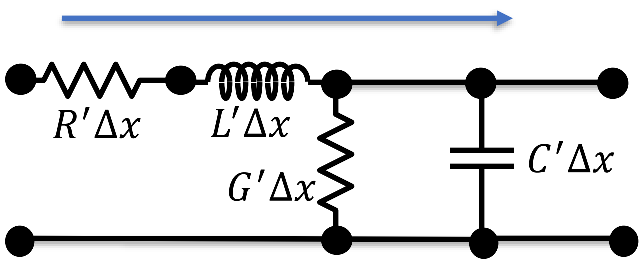
Transmission Lines – resonance and Q

- For (A) on right, $V_+ = \frac{V_0}{2}, V = V_+ + V_- = V_0 e^{-j\beta l}, V_- = \frac{V_0}{2} e^{-2j\beta l}$
  - $V_g = V_0 e^{-j\beta l}\cos(\beta l) = \frac{V_0}{2}(1 + e^{-2j\beta l}), V_g(\frac{\lambda}{4}) = 0$
  - $I_g = \frac{V_+}{Z_0} - \frac{V_-}{Z_0} = jI_s e^{-j\beta l}\sin(\beta l).$
  - $X = \frac{V_g}{jI_g} = \frac{Z_0}{\tan(\beta l)}$
- $Q = \omega\frac{E}{P_a}, E = \frac{lP_+}{v}, P_a = P_+ - P_+ e^{-2\alpha l}\rho(0) \approx 2\alpha l P_+, Q = \frac{\beta}{2\alpha}$
- In (B) to the right, the coefficient of reflection is $\rho(0)$ and the generator absorbs the reverse wave. $V = V_+ + V_- = V_0 e^{-j\beta l}$.
  - $V_f = \rho(0)V_+ e^{-j\beta l}, V_r = \rho(0)V_+ e^{-2j\beta l}$
  - $\rho(l) = \frac{V_-}{V_+} = \rho(0)e^{-2j\beta l}$ is the reflection coefficient at generator.
  - $\rho\left(\frac{\lambda}{2}\right) = \rho(0), \rho\left(\frac{\lambda}{4}\right) = -\rho(0)$
  - $\frac{Z(\lambda/4)}{Z_0} = \frac{Z_0}{Z(0)}, z = \frac{Z}{Z_0}, y = \frac{1}{z}, z\left(\frac{\lambda}{4}\right) = -\frac{1}{z(0)}$
  - Normalized: $Z(\lambda/4) = \frac{1}{z(0)}$
  - $Z_0 = \sqrt{Z(\lambda/4)Z(0)}, Z_0 = \sqrt{R_S R_L}$





Modelling Coaxial Cables

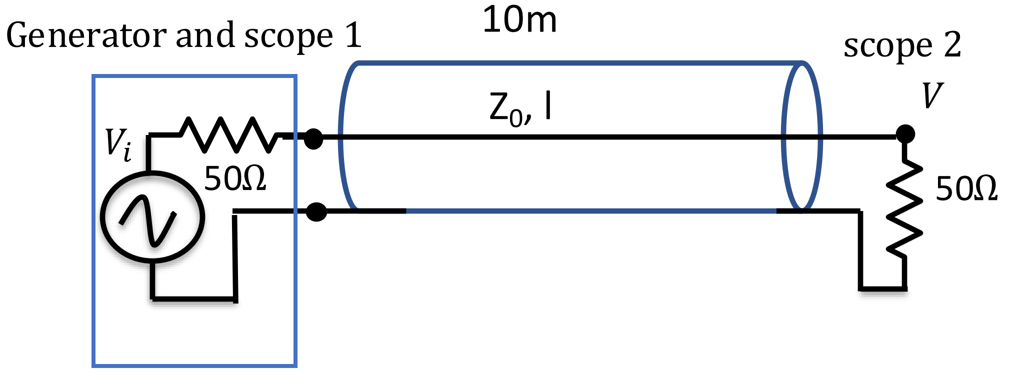


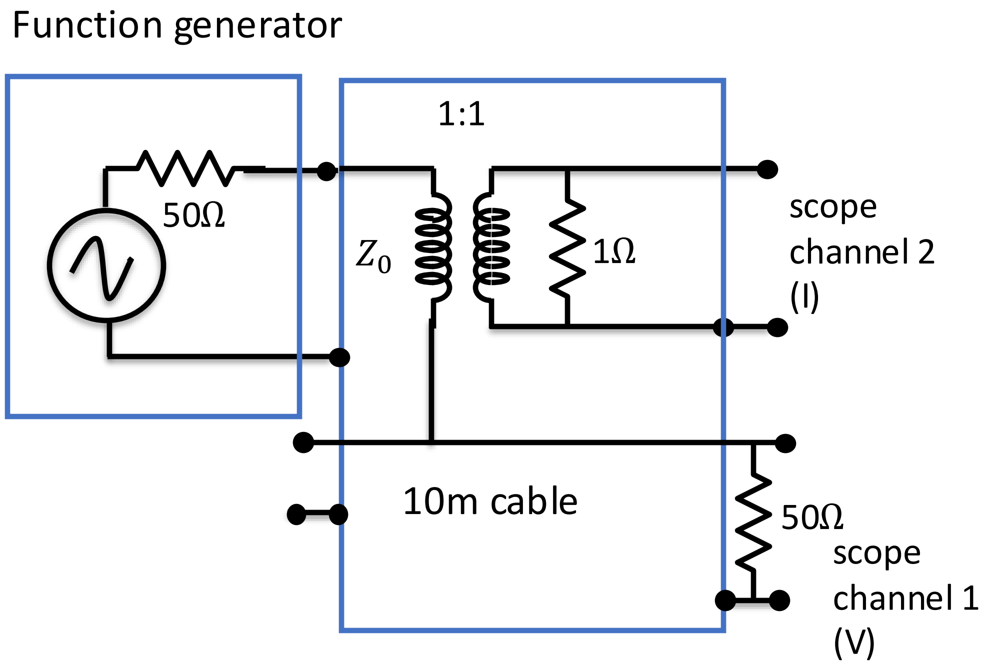
- A small segment of coax (of length Δx) can be modelled by the circuit to the right.
  - R' and L' are characteristic of the two parallel conductors.
  - G' and C$'$ are characteristic of the dielectric between the two conductors and the distance between them.
- The characteristic impedance, Z_0, is the impedance of an infinitely long coax. For an RG-58 coax it's about 53Ω.
- For a coax consisting of a series of segments on the right, $Z_0 = \sqrt{\frac{R'+j\omega L'}{G'+j\omega C'}}$
- In the last few slides, we have simplified the model considerably.
- Notice that at high frequency, $Z_0 \approx \sqrt{\frac{L'}{C'}}$
- In an earlier slide (with a simpler circuit model), we showed that the propagation speed of a signal of high frequency is $v = \frac{1}{\sqrt{C'L'}}$
- Since $Z_0\sqrt{C'} = \sqrt{L'}$, $Z_0 C' = \sqrt{C'L'}$.
- An RG-58 cable, has $C' = 82\ {}^{pf}/_{m}$ and thus, $v = \frac{1}{53 \times 82 \times 10^{-12}}\, {}^{m}/_{s} \approx .75c$

Exercise 10: Coax

- We'll measure the velocity of waves in RG58U by connecting one channel of the scope to the input and one to the output.

1. Measure the velocity, v, in 10m coax at 7MHz. Try different frequencies. Use $50ns$, $5V_{pp}$using square waves at 20kHz. Ans: about $\frac{2}{3}c$
2. Do the same with an antenna.
3. Calculate Z_0 with 50Ω termination for the circuit on the right.
4. Remove the 50Ω and measure the V and use it and Z_0 to calculate L, and C for the coax
   - Measured speed is $v = 2 \times 10^8\ m/s$. $Z_0 = 50\Omega$. For high impedance, $Z_0 = \sqrt{\frac{L}{C}}$ and $v = \frac{1}{\sqrt{LC}}$. So, $Z_0{}^2C = L$ and $v^2 = \frac{1}{LC}$, so $Z_0{}^2C^2v^2$=1. $C = \frac{1}{Z_0 v} = 10^{-10}F$. $2500 \times 10^{-10}F =$ $\mathrm{L} = 250\mathrm{nH}$, which is what we use in the next problem.

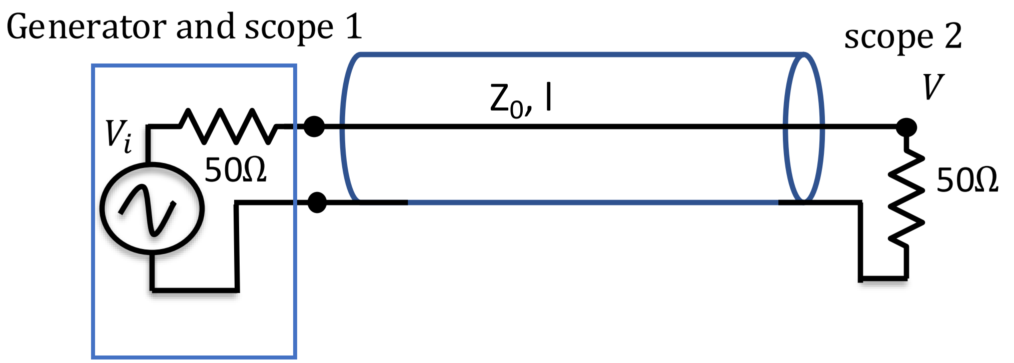




Exercise 11: Waves

- Suppose we want to send voice over 100km of coax, $Z_L = 50\Omega, l = 100km$.

1. Measure the SWR which is the ratio of the maximum to minimum output
   - $V = V_+ + V_-, \rho = \frac{Z - Z_0}{Z + Z_0}, Z = 50\Omega,$ we get Z_0 from the previous exercise.
   - $|V_{max}| = |V_f| + |V_r| = (1+\rho)|V_f|, |V_{min}| = (1-\rho)|V_f| = |V_f|.\ SWR = \frac{V_{max}}{V_{min}} = \frac{1+\rho}{1-\rho},$
2. If $L = 250\frac{nH}{m}, C = 100pf/m$ and the distributed resistance at voice is $50\ m\Omega/m$, calculate total dB loss at 500, 1000 and 2000Hz using the high frequency approximation.
   - $Z(f) = j\omega L + R = 50 \times 10^{-3} + j \cdot 2\pi f \cdot 250 \times 10^{-9}$
   - $Y(f) = j\omega C + \frac{1}{R} = \frac{1}{50 \times 10^{-3}} + j\ 2\pi f \cdot 10^{-10}$
   - $Z_0(f) = \sqrt{\frac{Z(f)}{Y(f)}}$
   - $Z_0(500) = 400\Omega, Z_0(1000) = 282\Omega, Z_0(500) = 200\Omega,$
   - High resistance approximation: $\alpha(f) = \sqrt{\frac{\omega RC}{2}},$
   - $\alpha(500) = 8.8 \times 10^{-5}, \alpha(1000) = 12.6 \times 10^{-5}$
   - $\alpha(2000) = 17.6 \times 10^{-5} \times 10^5$
   - For $100km$, loss is $\alpha \times 10^5$



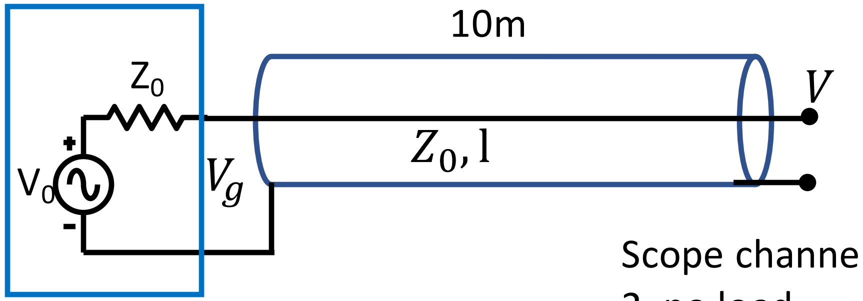
Exercise 11: Waves

3. Add a 100mH inductor every 1km. Now what's the loss?
   - $Z(f) = j\omega L + R = 50 \times 10^{-3} + j \cdot 2\pi f \cdot 10^{-4}$, $Z_0(f) = \sqrt{\frac{Z(f)}{Y(f)}}$
   - $Z_0(f) = \sqrt{\frac{Z(f)}{Y(f)}}$
   - $Z_0(500) = 318\Omega$, $Z_0(1000) = 317\Omega$, $Z_0(2000) = 316\Omega$
   - High reactance approximation: $\alpha(f) = \frac{R}{2Z_0}$, $Z_0 = \sqrt{\frac{L}{C}} = 1000\Omega$
   - $\alpha(f) = \frac{R}{2Z_0(f)}$, $\alpha(500) = \alpha(1000) = \alpha(2000) = \frac{5\times10^{-2}}{2000} = 5.5 \times 10^{-5}\, nepers/m$
   - For $100km$, loss is $\alpha \times 10^5$=5.5 or $5.5 \times 8.868 \approx 49dB$

Exercise 12: Resonance

$RG58U$ has a capacitance of about $100\frac{pF}{m}$. Let α be the attenuation constant and β be the phase

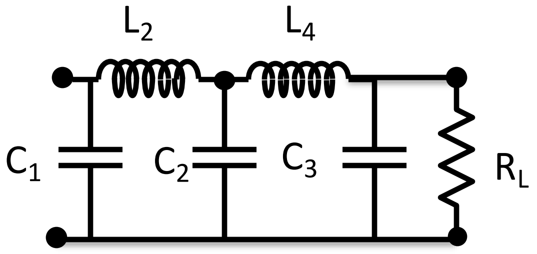
1. Derive an expression for $|\frac{V_g}{V}|$ and assuming α is small by finding the first resonance where V_g is minimum.
   - $V_g = V_0 e^{-j\beta l}\cos(\beta l), V = V_0 \exp(-j\beta l)$. $\left|\frac{V_g}{V}\right| = \cos(\beta l)$. So, at $l = \frac{\lambda}{2}, \left|V_g\right| = |V|$
2. Find α and the wave velocity by finding the resonant frequency (without the load, $1V_{pp}$) and noting the time delay with a scope on the input and output. Use $|\frac{V_g}{V}|$ to calculate α.
   - $|V|$ will be maximum at resonant frequency with unterminated line.
   - $|V_g(l)|$ is minimum when $l = \frac{\lambda_r}{4}$ and $\beta l = \frac{\pi}{4}$. This gives β.
   - At $l = \frac{\lambda}{2}, \left|\frac{V_g}{V}\right| = e^{-\alpha(\lambda/2)}$
3. Use this to calculate the velocity, v. How large is the frequency shift caused?
   - $v = \frac{\omega_r}{\beta}$. [$v$ should be about $2 \times 10^8\ m/s$]
4. Find, as usual, f_u, f_l, and Q.
   - $Q = \frac{\beta}{2\alpha}$
   - $Q = \frac{f_r}{BW}$, so $BW = \frac{f_r}{Q}$. $f_u = f_r + \frac{BW}{2}$, and $f_l = f_r - \frac{BW}{.2}$



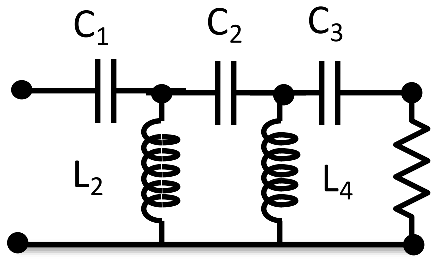
Scope channel 1 and function generator

Scope channel 2, no load

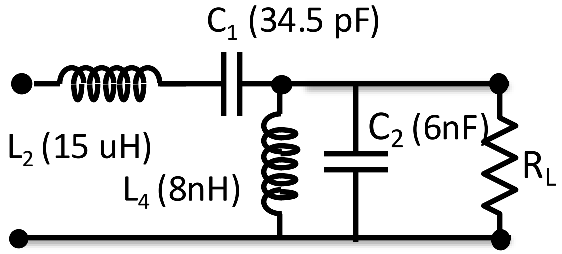
Filters



Low pass

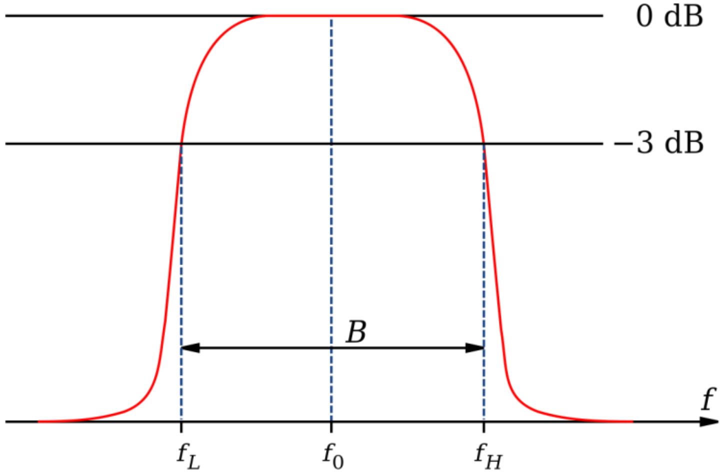


High pass

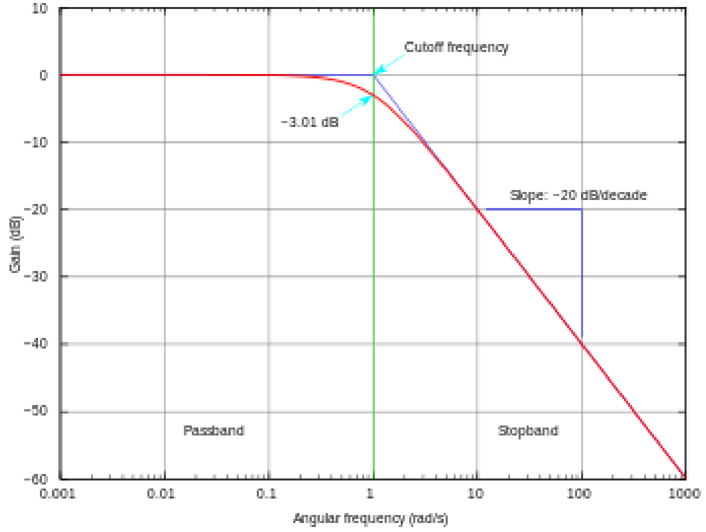


7 MHz bandpass

- Circuits on the left are called ladder filters.
- Low pass (Butterworth equivalent): Tabled values for inductors and capacitors based on frequency and dB drop-off.
- Can convert low pass into bandpass.
- For low pass to high pass
- Butterworth: $L = \frac{P_i}{P} = 1 + (\frac{f}{f_c})^{2n}$, f_c is 3dB cutoff
- Chebyshev: $L = 1 + \alpha {C_n}^2 (\frac{f}{f_c})^{2n}$, f_c is 3dB cutoff
- Normalized reactance's: $a_i = \sin(\frac{(2i-1)\pi}{2n})$
- Ripple loss: $1 + \alpha = 10^{L_r/_{10}}$
- $\beta = \sinh(\frac{tanh^{-1}(1/\sqrt{1+\alpha})}{n})$, $c_i = \frac{a_i a_{i-1}}{c_{i-1}(\beta^2 + sin^2((i-1)\pi/n))}$
- Example: cutoff at 10MHz, 4th order, 50ohm output, 3dB cutoff, $L(20MHz) = 6n = 24dB$, $a_1 = 0.765$, $X_1 = x_1 Z_0 = 38\Omega$, $L_1 = \frac{X_1}{\omega_c} = 610nH$, $b_2 = a_2 = 1.848$, $B_2 = b_2/Z_0$, $C_2 = B_2/\omega_c$
- Yuck!

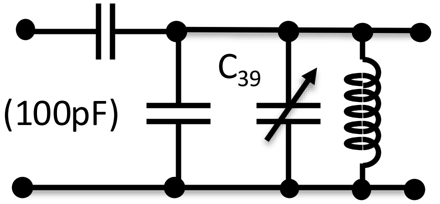


Bandpass - WIkipedia



Lowpass - WIkipedia

Norcal transmit bandpass filter



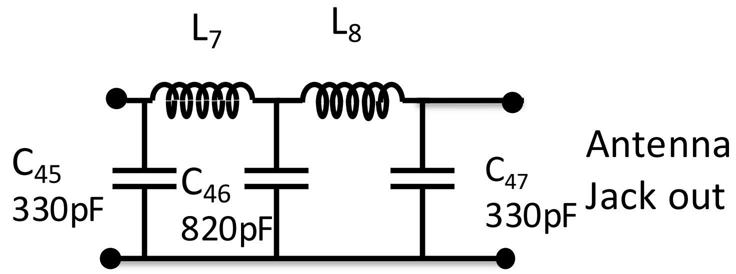
- $C_{39} = 50pF,$
- L_6 is 36 turns #28 on T37-2 which has $A_l = 4\frac{nH}{turn^2}, L_6 = A_l \cdot 36^2 = 3.1\mu H$
- $Z_2 = -\frac{j}{(C_{38}+C_{39})\omega_o},\ Z_3 = jL_6\omega_o, Z_1 = \frac{j}{C_{37}\omega_o}$
- $Z_{2,3-eq} = \frac{jL_6\omega_0}{L_6(C_{38}+C_{39})\omega_0{}^2-1}$ L6
- Resonance is when $Z_{2,3-eq} \to \infty, {\omega_o}^2 = \frac{1}{(C_{38}+C_{30})L_6} \approx \frac{10^{18}}{465}$, when almost all the voltage drop is across $Z_{2,3-eq}$ $\omega_o = \frac{10^9}{\sqrt{465}} \approx 50.8 \times 10^6, f_0 = \frac{\omega_o}{2\pi} \approx 7.1\ MHz$
- Q of filter is: $Q_s = \frac{X_s}{R_s}$. R_s comes from the other components and must be measured
- Note that $Z_{2,3-eq}$ is small for the other modulation product

Exercise 13: Norcal Harmonic Filter

- L_7, L_8 use T37-2 core, 18 turns, 1.3uH. Use 50Ω termination and set function generator at $10V_{pp}$.

1. Compute and compare loss at 7MHz and 14MHz.
2. From $A_l = 5nH/turn^2$, calculate L_7 and L_8.
3. What is the spur strength at 7, 14 and 28MHz? Measure and calculate.

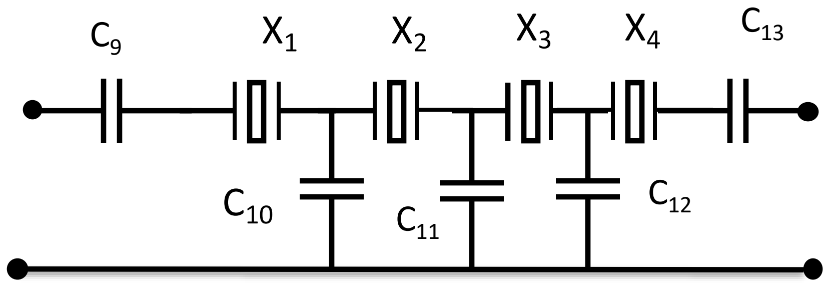
- Need Puff (a simulator) to get losses. However, answer is there is a 6dB drop-off at every frequency doubling

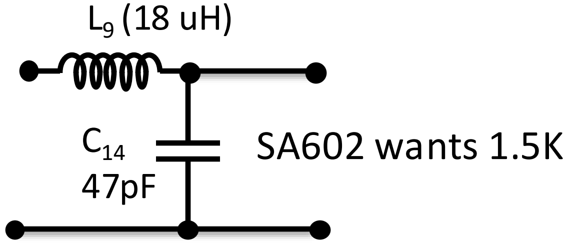


Exercise 14: Norcal IF Cohn Filter

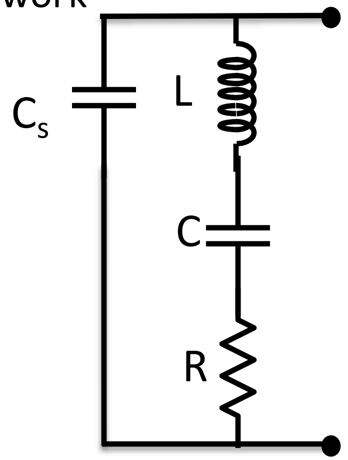
- X_1 through X_4 are 4.91 MHz
- C_{10}, C_{11}, C_{12} are 270 pF
- Set function generator to 50mV_{pp} from function generator
- Calculate R and X for filter

1. Measure the resonant frequency of one of the crystals
   - Duh
2. Calculate the parameters of the crystal. Omitting C_s
   - $f_r = \frac{1}{2\pi\sqrt{LC}}$ and $Q = \frac{1}{R}\sqrt{\frac{L}{C}}$. We can measure f_r and find Q using the 3dB bandwidth. R is the resistance at resonance.
   - $Q \approx 80$
   - $25\Omega < R < 100\Omega$
   - If $R = 50$, $\mathrm{C} = 8.1pF$, $L = 130\mu H$





Matching network



Equivalent circuit for crystal

Transformers

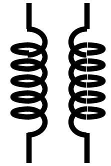
- For solenoid, $\oint B \cdot ds = \ \mu_0 n I$ inside
- $LI = \Phi_B$. Since there are n turns in the solenoid, over the solenoid, $LI = \mu_0 n^2 I$, so $L = \mu_0 n^2$.
- This is the source of $L = A_l n^2$
- $V_s = \frac{N_s}{N_p} V_p$
- $Z_p = (\frac{N_p}{N_s})^2 Z_s$

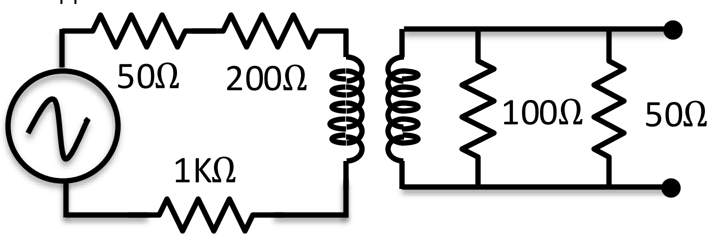
Exercise 15: Norcal Driver Transformers (1)

- T_1 uses FT 37−43. $L(\mu H) = \frac{A_L t^2}{1000}$, $A_L = 350.$ $f_r = 7 \times 10^6 MHz$, $n_p = 14$, $n_s = 4$, $\omega_r = 2\pi \times 7 \times 10^6 MHz = 4.4 \times 10^7$

1. Measure the output V_{out}.
2. Calculate V_{out}
   - $L_p = 68.6\mu H,\ L_s = 5.6\mu H$
   - $Z_{eq,in}(\omega) = 1250 + j(\omega L_p)$, $Z_{eq,in}(\omega_r) = 1250 + 3016j,\ |Z_{eq,in}(\omega_r)| = 3264$
   - $Z_{eq,out}(\omega) = 33 + j\omega L_s, Z_{eq,out}(\omega_r) = 33 + j246,\ |Z_{eq,out}(\omega_r)| = 248$
   - $V_{t,in} = \frac{3016}{3264} V_{in}$
   - $V_{out} = V_{t,out} = \frac{n_s}{n_p} V_{t,in} = .29 V_{t,in} = .29 \times \frac{3016}{3264} \times 5 = 1.3V$
   - $i_p(\omega) = \frac{V_{in}}{|Z_{eq,in}|} e^{j\theta_p(\omega)}$, $\theta_p(\omega) = \arctan\left(\frac{\omega L_p}{1250}\right)$; $i_s(\omega) = \frac{V_{out}}{|Z_{eq,out}|} e^{j\theta_s(\omega)}$, $\theta_s(\omega) = \arctan\left(\frac{\omega L_s}{33}\right)$.
   - $P_{in,a} = Re\left(\frac{V_{in}\overline{I_{in}}}{2}\right) = Re\left(\frac{{V_{in}}^2}{2|Z_{eq,in}(\omega)|} e^{j\theta_p(\omega)}\right)$
   - $P_{out,a} = Re\left(\frac{V_{out}\overline{I_{out}}}{2}\right) = Re\left(\frac{{V_{out}}^2}{2|Z_{eq,out}(\omega)|} e^{j\theta_s(\omega)}\right)$

T_1, 14:4





Exercise 15: Norcal Driver Transformers (2)

- $\cos\left(\theta_s(\omega_r)\right) = .13, \cos\left(\theta_p(\omega_r)\right) = .38,$
- $\frac{P_{out,a}(\omega_r)}{P_{in,a}(\omega_r)} = (\frac{V_{out}}{V_{in}})^2 \frac{|Z_{eq,in}(\omega_r)|}{|Z_{eq,out}(\omega_r)|} \frac{\cos(\theta_s(\omega_r))}{\cos(\theta_p(\omega_r))} = (\frac{1.3}{5})^2 \times \frac{3264}{248} \times \frac{.13}{.38} = .3$

3. Measure the 3dB cutoff, f_c.

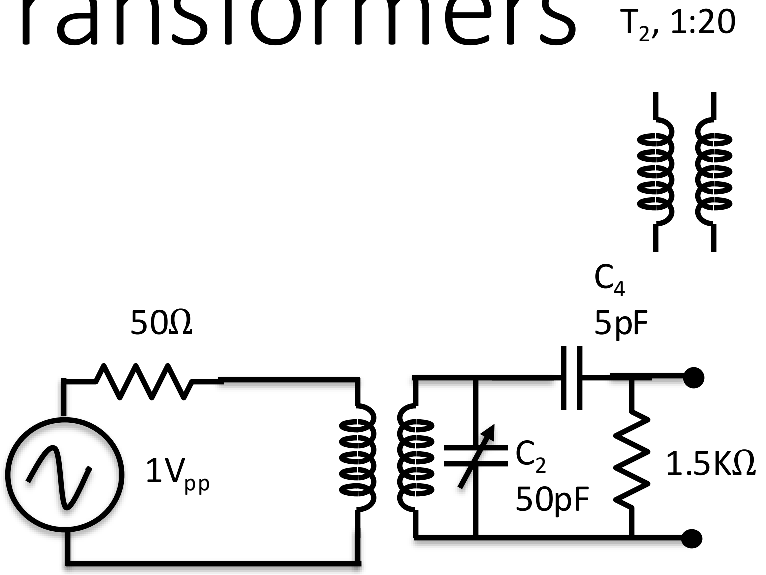
- $\frac{P_{out,a}(\omega)}{P_{in,a}(\omega)} = (\frac{V_{out}}{V_{in}})^2 \frac{|Z_{eq,in}(\omega)|}{|Z_{eq,out}(\omega)|} \frac{\cos(\theta_s(\omega))}{\cos(\theta_p(\omega))} = .15$

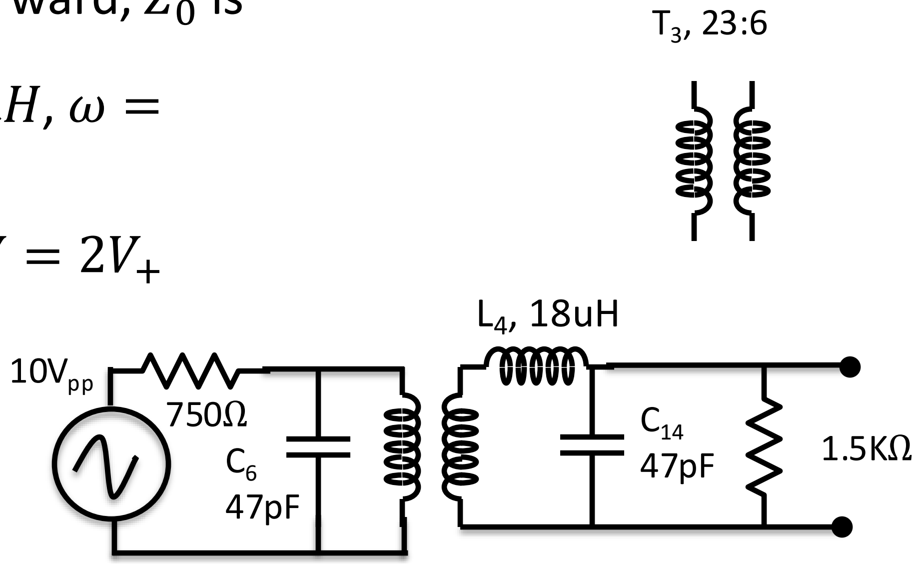
Exercise 16: Norcal Tuned Transformers

- $T_{2,}$ T_3 are IF matchers using FT 37−61. $A_L = 55\,nH/turn^2$.

1. Measure 3dB bandwidth
2. Find P/P_+

- T_2
  - $P_a = Re\left(\frac{V\bar{I}}{2}\right), V = V_+ + V_-, \rho = \frac{V_+}{V_-} = \frac{Z-Z_0}{Z+Z_0}$, Z is look forward, Z_0 is look back resistance. $P_+ = \frac{{V_+}^2}{2Z_0}$. $L_{in} = 55nH, L_{out} = 22\mu H, \omega = 4.4 \times 10^7$
  - $Z = 50 + 2.4j, Z_0 = 203 - 4030j, \rho = 1$, so $V_+ = V_-, V = 2V_+$
  - $\frac{P}{P_+} = \frac{1}{4}$
- Similar calculation for T_3.





Acoustics

- $P = P_0 + P_e, \rho = \rho_0 + \rho_e$

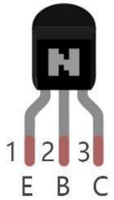
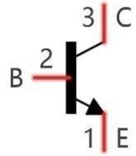
1. Gas moves and changes density: Displacement of undisturbed air is x. At time t, it's at $x + \chi(x,t)$, so $\rho_0 \Delta \mathrm{x} = \rho(\mathrm{x} + \Delta \mathrm{x} + \chi(x + \Delta x, t) - x - \chi(x,t))$, or $\rho_0 \Delta \mathrm{x} = \rho(\frac{\partial \chi(x,t)}{\partial x} \Delta x + \Delta x)$. So, $\rho_e = -\rho_0 \frac{\partial \chi}{\partial x}$.
2. Change in density causes change in pressure: $P = f(\rho)$, $P_0 + P_e = f(\rho) = f(\rho_0) + \rho_e f'(\rho_0), f'(\rho_0) = \kappa = (\frac{dP}{d\rho})_0, \text{or } P_e = \kappa \rho_e.$
3. Pressure differences cause motion: $P(x,t) - P(x + \Delta x, t) = -\frac{\partial P_e}{\partial x} \Delta x$, Newton's law gives $\rho_0 \frac{\partial^2 \chi}{\partial t^2} = -\frac{\partial P_e}{\partial x} = -\kappa \frac{\partial \rho_e}{\partial x}$.

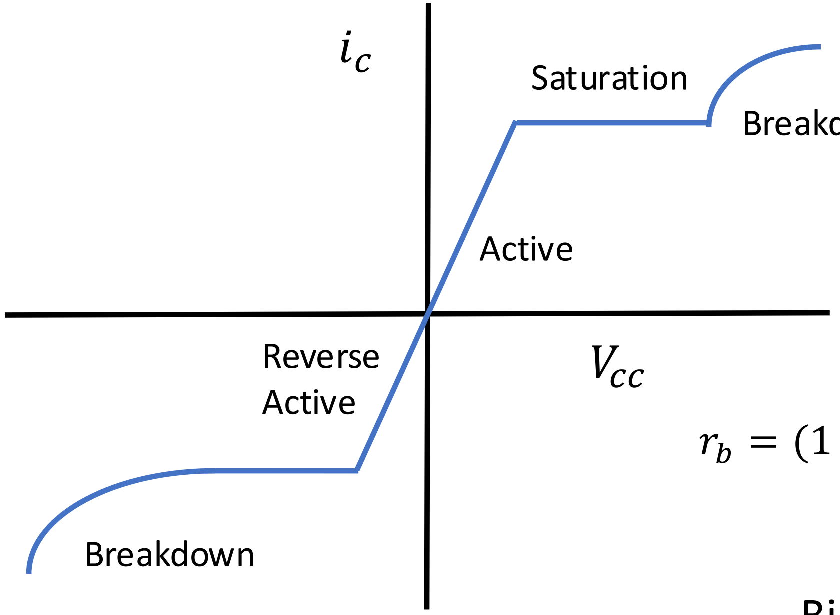
- Substituting (1) into (3) gives $\frac{\partial^2 \chi}{\partial t^2} = \kappa \frac{\partial^2 \chi}{\partial x^2}$, put $\kappa = \frac{1}{{c_s}^2}$.
- Solution is $\chi(x,t) = f(x - vt)$ [Different f than above].
- To find, $\kappa = (\frac{dP}{d\rho})_0$, note that the flow is adiabatic so $PV^\gamma = C'$ and ρ varies inversely as V, so $P = \rho^\gamma C$, and finally, using $PV = Nkt, \kappa = (\frac{dP}{d\rho})_0 = \frac{\gamma k T}{n}$
- $L_p = 20 \log(\frac{P}{P_0}), P_0 = 20\ \mu Pa.$

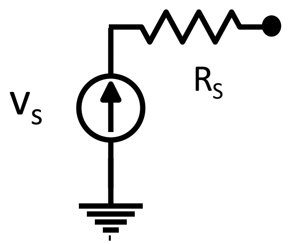
Sound	L_p	Power density
rustling leaves	10dB	1pW/m^2
broadcast studio	20dB	1pW/m^2
classroom	50dB	10nW/m^2
heavy truck	90dB	1nW/m^2
Shout at 1m	100dB	10mW/m^2
jackhammer	110db	100mW/m^2
jet takeoff at 50m	120dB	1W/m^2

Bipolar Transistors - I

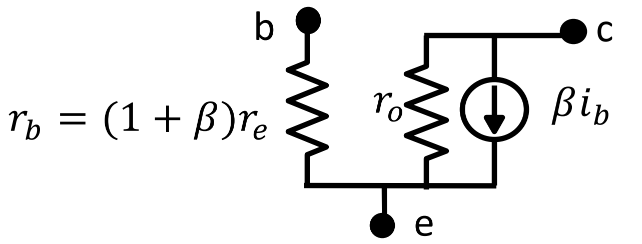
- NPN Model
  - $V_f \approx .7V, V_s \approx .2V$
  - Conducts when $V_{be} > V_f$
  - Saturated when $V_{ce} = 0$
  - $i_c = \beta i_b$
  - $i_c = \alpha i_e$
  - $\beta = \frac{\alpha}{1-\alpha}$ [$= h_{fe}$, small signal]
  - $\beta \approx 100, \beta_r \approx 10$
  - $r_e i_e = 25mV, r_b = (1+\beta) r_e, r_e \approx 33\Omega$
  - $i_b = \frac{v_{be}}{(1+\beta) r_e}$
  - $g_m v_{be} = g_m r_b i_b$
- Switch
  - $G_s = \frac{i_b}{15mV}$
  - $R_S = 2\Omega$



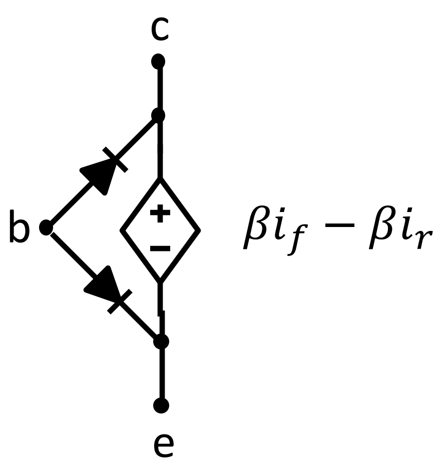


Bipolar source model



Bipolar equivalent circuit

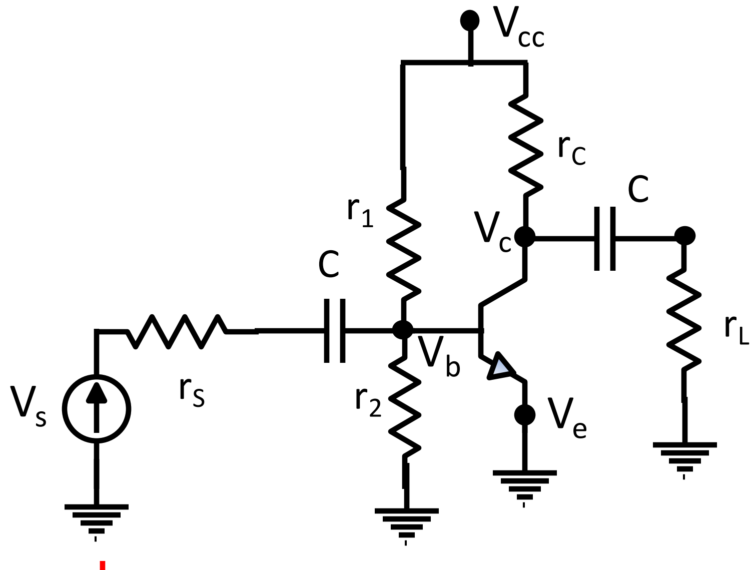
	V_{be}	V_{bc}	V_{ce}	i_c
active	V_f	$< V_f$	$>V_s$	βi_b
rev saturated	$<V_f$	V_f	$<-V_s$	$-(\beta_r + 1) i_b$
saturated	V_f	V_f	$V_s > V_{ce} \succ -V_s$	$>-(\beta_r + 1) i_b < \beta i_b$
cutoff	$< V_f$	$< V_f$	*	0



Bipolar model

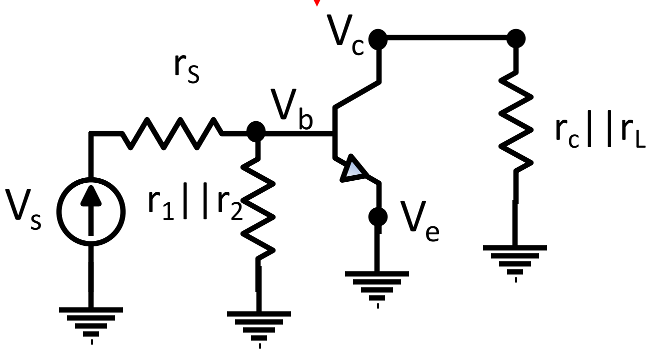
Bipolar Transistors - II

- NPN Mode
  - $V_f = .7\text{V}$
  - $\beta = g_m r_\pi$
  - $g_b = \frac{i_b}{V_t}, V_t \approx 25mV, g_m = \frac{i_c}{V_t}$
- DC
  - $\frac{V_{cc}-2V_f}{R_C} < i_c, \beta i_b = i_c$
  - $V_c = V_{cc} - i_c R_C$
  - $\frac{V_{cc}-V_b}{R_B} = i_b$
- Small signal
  1. Convert to AC only and simplify
  2. Thevenize circuit
  3. Replace transistor with model

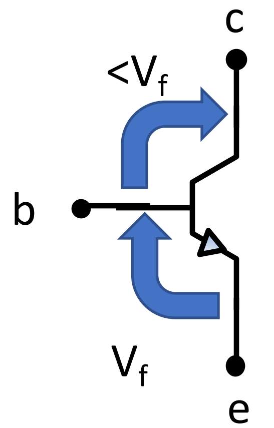


Small signal - original

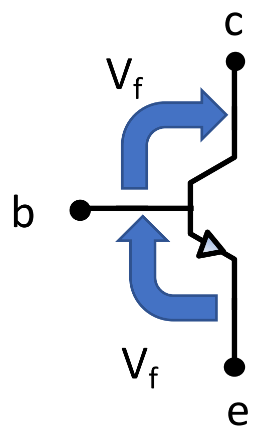
Step 1



Small signal



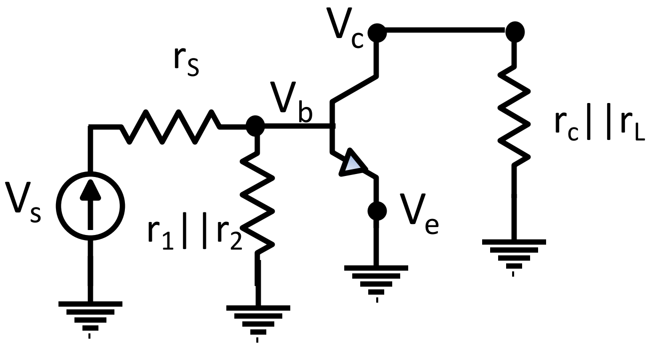
Active



Saturated

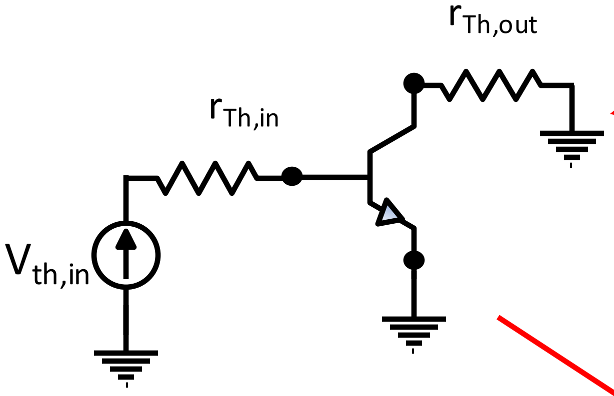
Bipolar Transistors - III

- Small signal
  1. Convert to AC only and simplify
  2. Thevienize
     - $V_{th,in} = V_s \frac{r_1||r_2}{r_1||r_2+r_s}$
     - $r_{th,in} = r_s||r_1||r_2$
     - $r_{th,out} = r_C||r_L$
  3. Replace transistor with model
     - $\frac{V_{be}}{V_{Th,in}} = \frac{r_\pi}{r_\pi + r_{Th,in}}$
     - $\frac{V_{out}}{V_{be}} = -g_m\, r_0||r_c||r_L$
     - r_0 is the transistor model resistance between b and c
     - $A_{gail} = \frac{V_{out}}{V_s}$



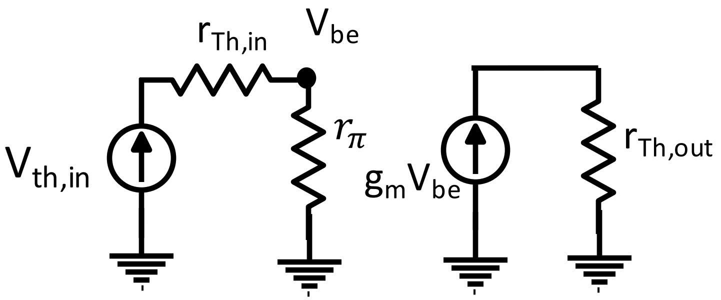
Small signal

Step 2



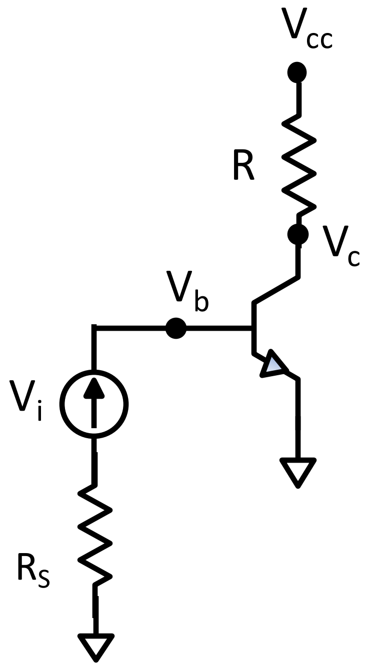
Small signal

Step 3



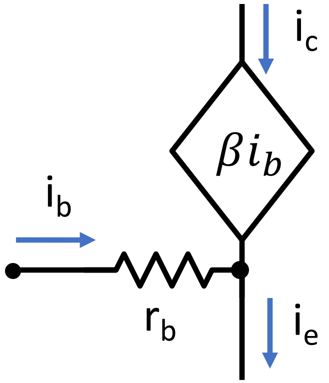
Small signal

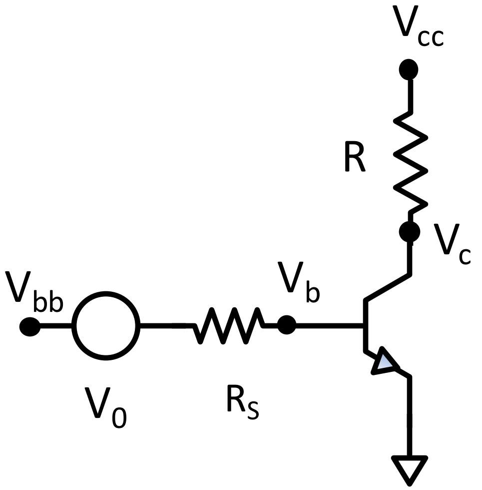
Bipolar transistors - IV



- At saturation, $v_{bc} < V_f$, so there is conduction from the collector to the base.
- $i_b = i_{bs}\exp(\frac{V_b}{V_t})$, V_t is the thermal voltage, $V_t = 25mV$, i_{bs} is the base saturation current.
- $i_c = i_{cs}\exp\left(\frac{V_c}{V_t}\right)$. Note $i_{cs} = \beta i_{bs}$. Both increase rapidly with temperature

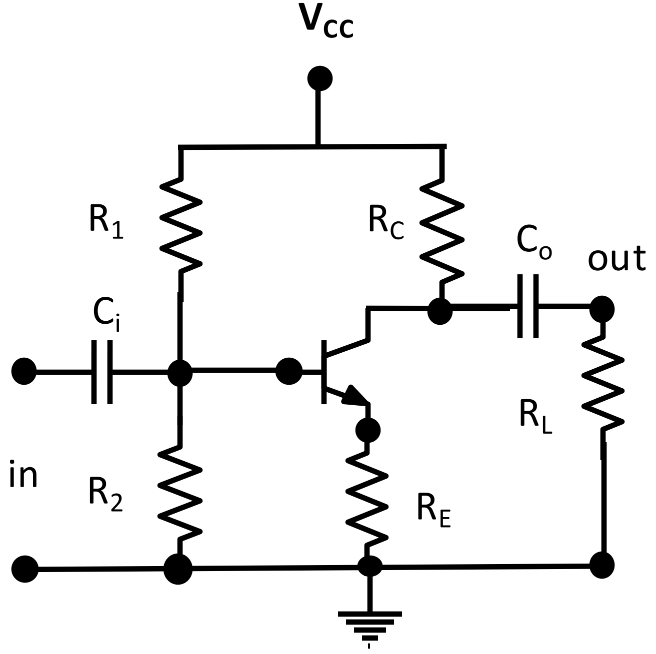
- Base resistance
  - $g_b = \frac{i_b}{V_t} = \frac{di_b}{dV_b}$
  - $r_b = \frac{25mV}{i_b}$
  - $g_m = \frac{i_c}{V_t} = \frac{di_c}{dV_b}$
  - $i_b = r_b V_b$



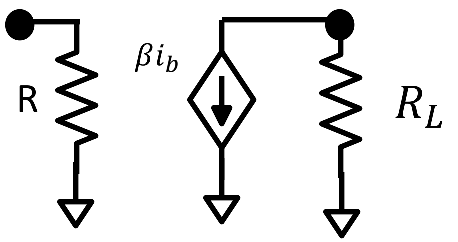


More on bipolar transistor model

- Saturated when $V_{ce} = 0$
- $v_{R_2} = v_{be} + (i_c + i_b)R_E$
- $i_{R_2} = \frac{v_{R_2}}{R_2}$
- $i_{R_1} = i_{R_2} + i_b$
- $v_{cc} = R_1 i_b + (R_1 + R_2) i_2$
- For $R_B = R_1 || R_2$, $v_{cc}R_B - v_{be}R_1 = R_1R_2i_b + (i_c + i_b)R_1R_E$, $i_c = \beta i_b$
- $i_c = \frac{v_{cc}\frac{R_B}{R_1} - v_{be}}{R_E + \frac{(R_C + R_E)}{\beta}}$
- If $R_E \gg \frac{(R_C+R_E)}{\beta}$, $\frac{\partial i_c}{\partial v_{be}} = -\frac{1}{R_E}$, be acts like diode so $i_c = i_s \beta \exp(\frac{V_{be}}{V_T})$.
- Want $V_E \approx 2v$



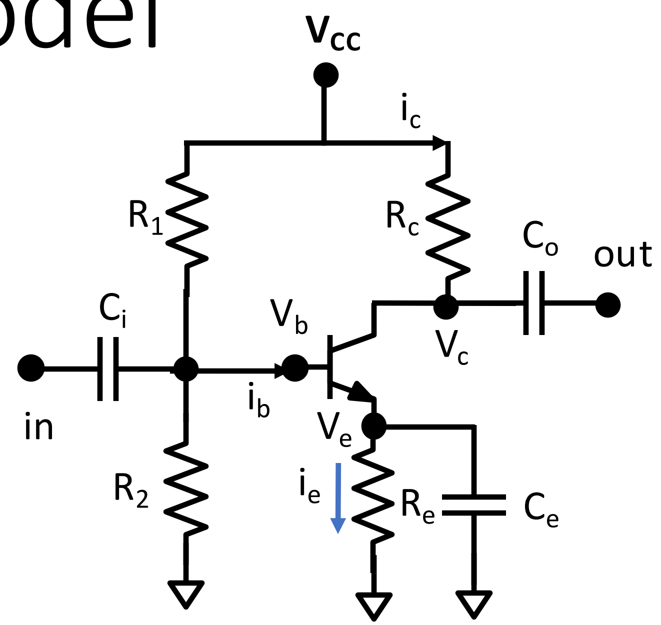
Small signal equivalent



$$R = R_1 || R_2 || r_b$$

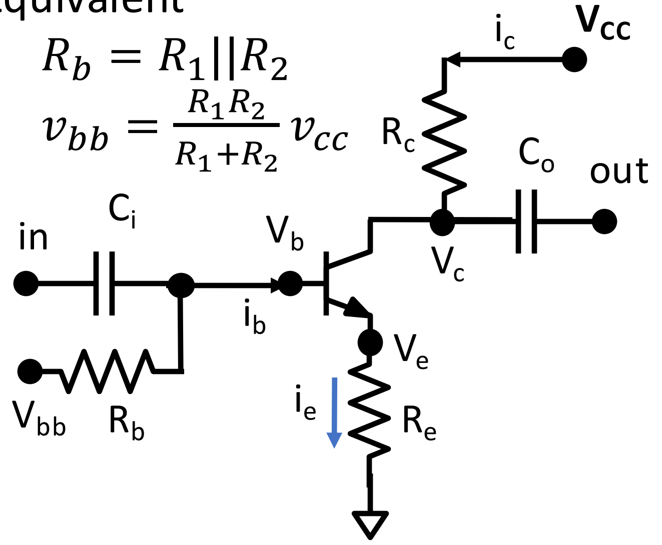
$$r_e = \frac{V_T}{i_c}, \qquad r_b = r_e(\beta + 1)$$

More on bipolar transistor model

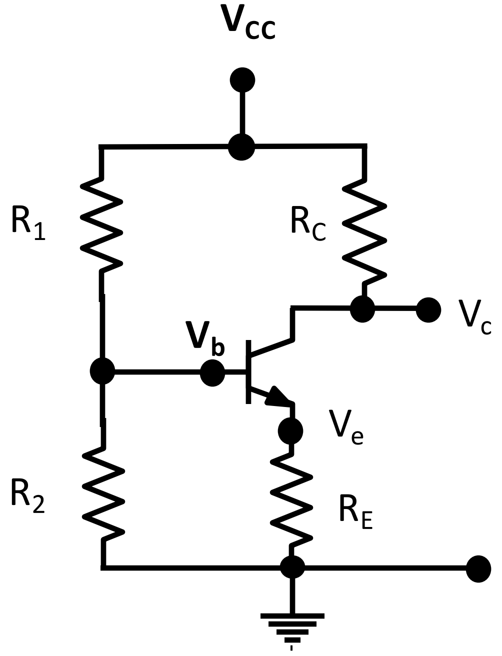


- $v_{be} = v_b - v_e, v_{ce} = v_c - v_e, v_{bb} = \frac{R_2}{R_1+R_2} v_{cc}, v_f \approx$ .6 (for Si)
- $v_b = v_e + v_f, i_b = \frac{v_{bb}-v_b}{R_b}$
- $i_c = \beta i_b + i_{ceo}, \frac{\partial i_b}{\partial v_{be}} = \frac{1}{r_d}$
- $\frac{\partial i_c}{\partial v_{be}} = g_s = -\frac{1}{R_B}$, and $\frac{\partial i_c}{\partial \beta}$ measure stability
- $i_b = \frac{v_{bb}-v_b}{R_b}$
- $i_c = \frac{\beta(v_{bb}-v_{be})}{R_b+(1+\beta)R_e} + \frac{(R_b-R_e)}{R_b+(1+\beta)R_e} i_{ceo}, i_{ceo}$ is leakage current
- So, if $\beta R_e \gg R_b + R_e, i_c = \frac{v_{bb}-v_{be}}{R_e} = \frac{v_{cc}-v_c}{R_c}$
- $Z_{in} = R_1||R_2||(\beta+1)R_e, Z_{out} = R_c$
- $v_{bb} - v_{be} = \frac{R_e}{R_c}(v_{cc} - v_c)$
- Typical for Si: $v_{be} = v_f \approx .6V, \beta \approx 200, v_{cc} = 9V, v_{bb} = 3V, R_b = 10^4\Omega, R_c = 10^3\Omega, R_e = 270\Omega$
  - $3 - .6 = \frac{270}{1000}(9 - v_c)$, so $v_c = .2V$
  - For voltage divider: $R_1 = 2 \times 10^4\Omega, R_2 = 10^4\Omega,$

Equivalent
- $R_b = R_1||R_2$
- $v_{bb} = \frac{R_1R_2}{R_1+R_2} v_{cc}$



Transistor experiment

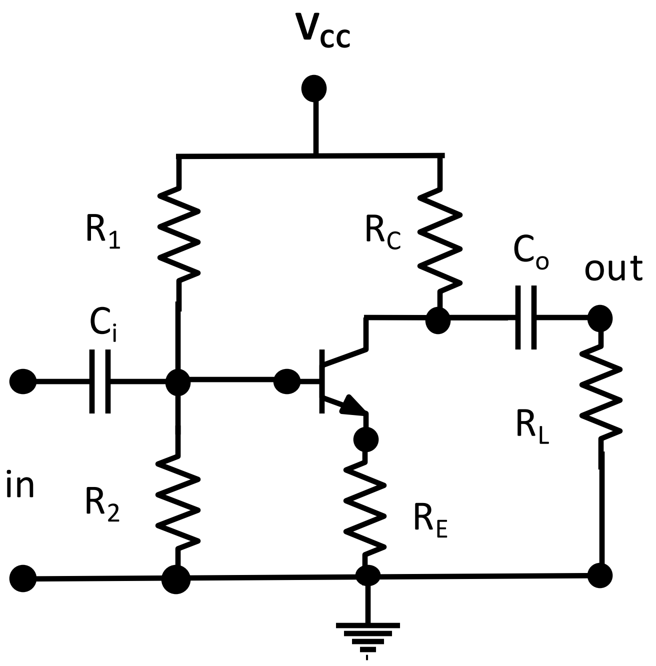


Experiment A

- $V_{cc} = 9V$, $R_1 = 22.8k\Omega$, $R_2 = 7.2k\Omega$, $R_c = 1k\Omega$, $R_E = 220k\Omega$. 2n3904 transistor, $\beta = 150$.
- With no transistor, R_2 adjusted so $V_b = 2.36V$. $V_b = 2.24V$, $V_e = 1.54V$, $V_c = 1.89V$. $i_c = 7mA$, $i_b = 46\mu$A.

- Experiment B
  - Again, $V_{cc} = 9V$, $R_1 = 20k\Omega$, $R_2 = 10k\Omega$, $R_c = 1k\Omega$, $R_E = 220\Omega$. 2n3904 transistor, $\beta = 150$. With no transistor, R_2 adjusted so $V_b = 5.8V$. Put transistor in and $V_b = 2.4V$.
  - With transistor, $V_b = 2.4V$, $V_e = 1.7V$, $V_c = 1.74V$. $i_c = 7mA$, $i_b = 46\mu$A.
  - Analyze these with our transistor model.
  - Now use the Thevenin equivalents to analyze them.

Turn the transistor experiment into a CE amplifier

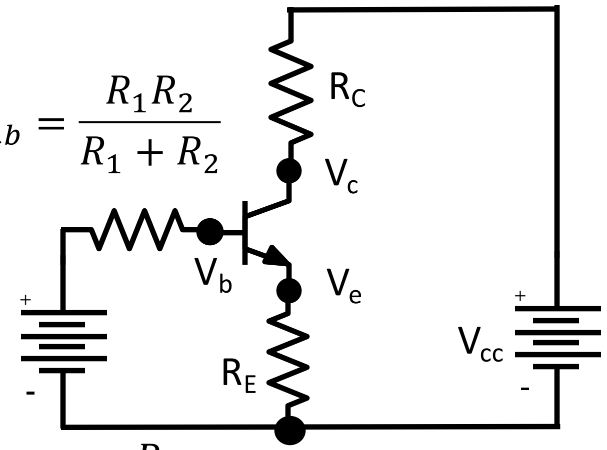


- Add C_i and C_o. Component values are:
  - $C_i = C_o = 1\mu F$
  - $R_1 = 20k\Omega, R_2 = 10k\Omega$
  - $R_C = 1k\Omega, R_E = 220\Omega$
  - $V_{cc} = 9V$

1. Use a function generator to generate a $V_{pp} = 800mV$, 10kHz.
2. The input impedance is $Z_{in} = R_1||R_2||(\beta + 1)R_E$, and the output impedance is $Z_{out} = R_C$. Add a load R_L whose value is Z_{out}.
3. Now connect a scope to the output and measure the gain. Calculate what it should be an compare them. How do the input and output waveforms compare?

Transistor experiment - Thevenin equivalent DC

$$R_{thb} = \frac{R_1 R_2}{R_1 + R_2}$$



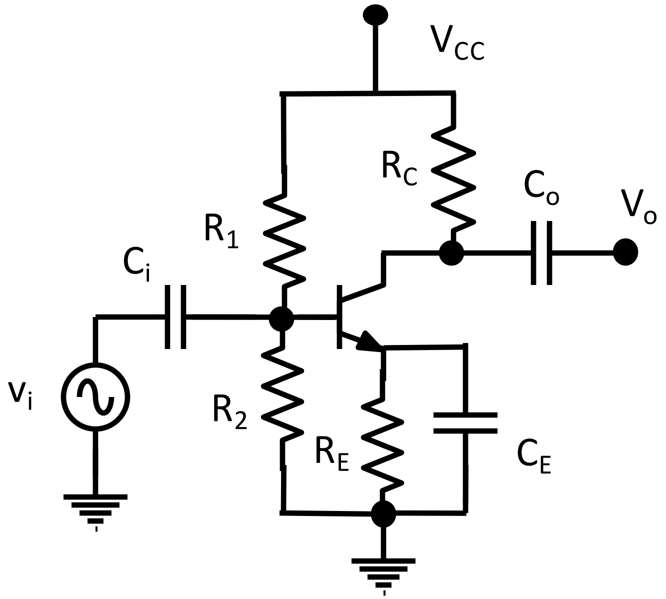
$$V_{thb} = \frac{R_1}{R_1 + R_2} V_{cc}$$

- In Experiment A
  - $R_1 = 22.8k\Omega, R_2 = 7.2k\Omega, V_{thb} = 2.16V, R_{th} = 5.5k\Omega.$
  - If $r_e \approx 33\Omega, r_b \approx 5k\Omega,\ i_b = \frac{2.16-1.54}{11500} = 53\mu A$, which is close.

- In Experiment B
  - $R_1 = 22k\Omega, R_2 = 8k\Omega, V_{thb} = 2.4V, R_{th} = 5.9k\Omega.$
  - If $r_e \approx 33\Omega, r_b \approx 5k\Omega,\ i_b = \frac{2.4-1.7}{10900} = 64\mu A$, which is also close, but a little high.

- Turn this into a CE amplifier by adding 1uF input and output capacitors. Measure and calculate the voltages and gains.

BJT common emitter amplifier



Common emitter amp

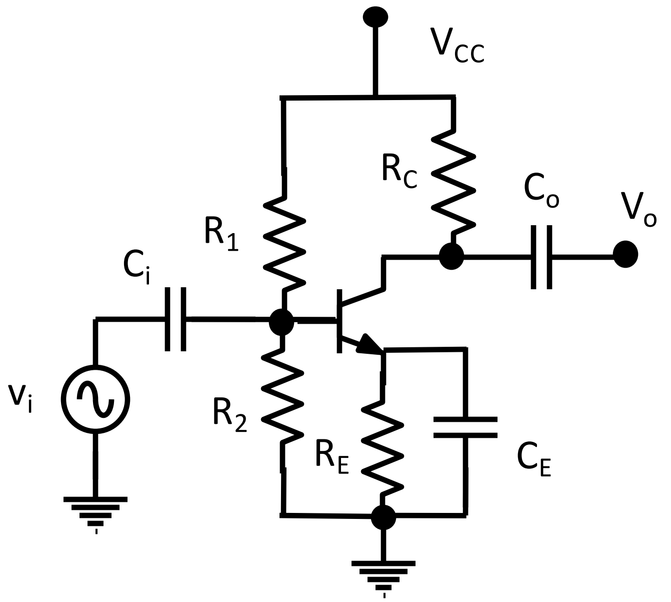
- Here's how to design a common emitter amplifier. We use a 2n3904 transistor with β=150. This circuit will work! Build it.

1. Pick the supply voltage V_{cc}=12V.
2. Choose a gain (amplification factor), A = 5.
3. Choose the "Q point" of the conducting transistor (4mA) and $V_{ce,q} = 5v$.
4. $V_{cc} = (i_c \cdot R_C) + V_{ce} + i_e R_E \sim i_e \cdot (R_C+R_E) + V_{ce}$ with i_c=4mA.
   We get $(R_C+R_E) = (V_{cc} - V_{ce})/(4mA)$ =1.75 k**Ω**.
5. Since A = 5 and $A=R_C/R_E$, R_C= 5 R_E so $R_E \sim 270$ **Ω** (this is a standard resistor value) and R_C= 1.5k**Ω**.

- $Z_{in} = \beta R_E$
- $Z_{out} \approx R_C$
- $\frac{V_o}{V_i} = \frac{\beta R_c}{r_b+(\beta+1)R_E}$

Credit: Ward, Hands on Radio.

BJT common emitter amplifier continued



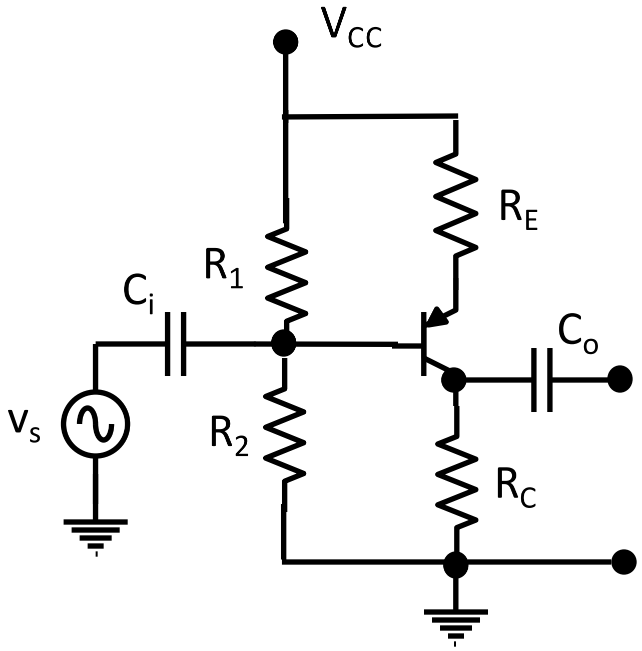
Common emitter amp

Credit: Ward, Hands on Radio.

6. i_b = 4mA/β = 27 $\boldsymbol{\mu}$A.
7. Since V_{be} must be greater than .7V throughout the input signal range, we want the voltage across R_2 to satisfy $V_{be} + i_cR_E$ = 1.8V.
8. Rule of thumb is current through R_1 and R_2 is $10i_b$. We insert a voltage divider consisting of R_1 and R_2, so that R_1= (12-1.8)/270 $\boldsymbol{\mu}$A ~ 39 k$\boldsymbol{\Omega}$. $R_2 = 6.7k\Omega$
9. C_o and C_i are picked to offer small resistance to the frequency range we're interested in and C_o = C_i = 5 $\boldsymbol{\mu}$F.

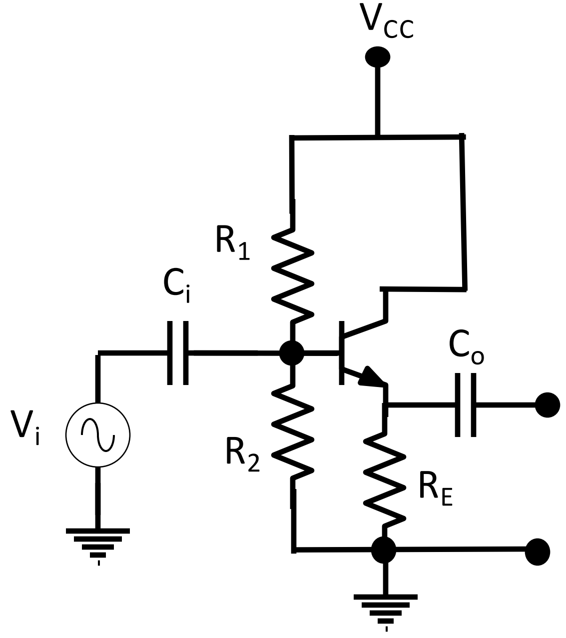
- I haven't explained why we want R_E, but it provides thermal stability for the transistor over the range we care about. The fact that $A=R_C/R_E$ can be calculated using Kirchhoff's laws.

BJT common emitter amplifier (PNP)



- 2n3906 transistor with β=150.
  1. Pick the supply voltage V_{cc}=12V.
  2. Choose a gain (amplification factor), A = 5.
  3. Choose the "Q point" of the conducting transistor (4mA) and $V_{ce,q} = 5v$.
  4. $V_{cc} = (i_c \cdot R_C) + V_{ce} + i_e R_E \sim i_e \cdot (R_C+R_E) + V_{ce}$ with i_c=4mA.
  5. We get $(R_C+R_E) = (V_{cc} - V_{ce})/(4mA)$ =1.75 k**Ω**.
  6. Since A = 5 and $A=R_C/R_E$, R_C= 5 R_E so $R_E \sim 270$ **Ω** and R_C= 1.5k**Ω**.
  7. V_{be} must be less than -.7V throughout the input signal range.
  8. i_b = 4mA/β = 27 μA. Rule of thumb is current through R_1 and R_2 is $10i_b$.
  - $Z_{in} = \beta R_E$
  - $Z_{out} \approx R_C$
  - $\frac{V_o}{V_i} = \frac{\beta R_C}{r_\pi + (\beta+1)R_E}$

BJT common collector amplifier

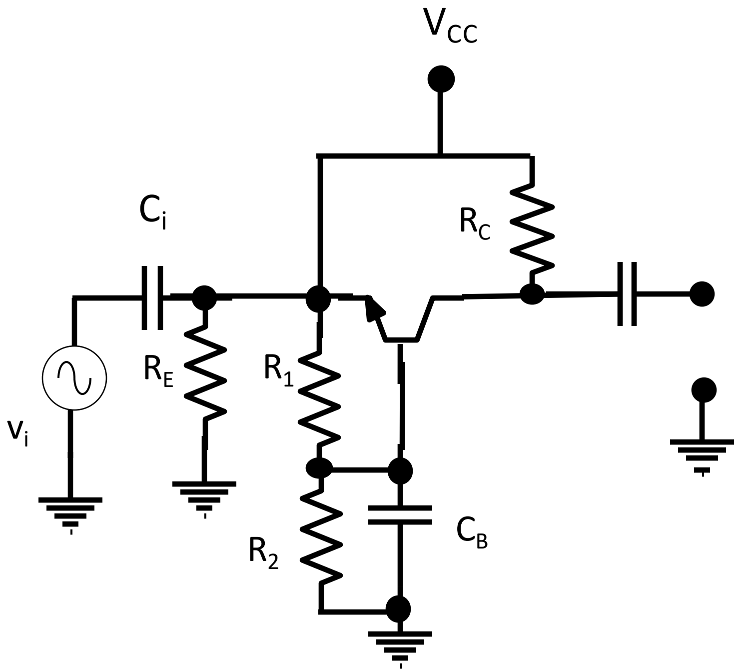


Common collector amp (Emitter Follower)

1. $\beta = 150,\ A_V = 1, V_{cc} = 12v$
2. Q-pt: $i_{ce} = 5mA$, $V_{ce,q} = 6v$ (rule of thumb), $v_{be} = .7V$.
3. $i_{R_1 \to R_2} = 10 i_b$ (ROT), $V_{ce} = \ v_{be} +\ i_{ce,q} R_E$, $R_E = 1.2k\Omega$, $i_b = \frac{V_{ce,q}}{\beta} = 33\mu A$
4. $V_{R_2} = \ V_{be} + i_C R_E = 6.7V$, $V_{R_1} = 5.3V$
5. $R_2 = \frac{6.7}{330\mu A} = 20k\Omega$, $R_1 = \frac{5.3}{330\mu A} = 16k\Omega$
6. $Z_{in} = \ R_1 \Big|\Big| R_2 \Big|\Big| (\beta + 1) R_E$, $Z_{out} = R_E || Z_E$, $Z_E = \frac{R_1||R_2}{(\beta+1)} + r_e{}'$
7. $R_{in} = 50\Omega$, $Z_{out} = 5\Omega$

Credit: Ward, Hands on Radio.

BJT common base amplifier



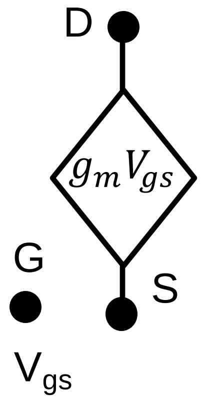
Common base amp

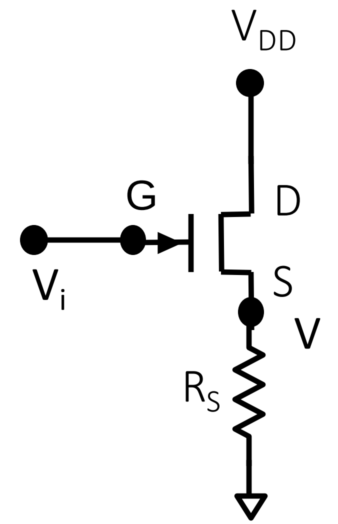
- $A_I = \frac{i_C}{i_E} = \frac{\beta}{\beta+1}$, $A_V = \frac{R_C||R_L}{r_e}$, $Z_{out} \approx R_C$

1. $V_{CC} = 12, V_{be} = .7V, R_E = 50\Omega, R_L = 1k\Omega, i_{ce,q} = 5mA, V_{ce,q} = 6V$
2. $i_b = \frac{i_{ce,q}}{\beta} = 33\mu A$, $i_{R_1 \to R_2} = 10\, i_b = 330\mu A$ (ROT)
3. $V_{R_2} = V_{be} + i_C R_E = 6.7V$, $V_{R_1} = 5.3V$
4. $R_1 = \frac{5.3}{330\mu A} = 16k\Omega$, $R_C = \frac{V_{cc} - i_{c,Q} R_E - V_{ce,Q}}{i_{c,Q}} = 1.35k\Omega$
5. $A_V = \frac{R_C||R_L}{26/i_e} = 115$

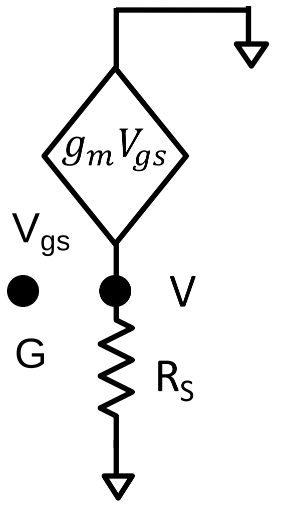
JFETs

- JFET circuit model (active region):
  - $I_d = I_{dss}(1 - \frac{V_{gs}}{V_c})^2$, provided $0 < v_{gs} < V_c$ and $V_{ds} > V_{gs} - V_c$. i_{dss} is drain to source current when gate is at 0. $v_{gs} \leq 0$
- $g_m = \frac{dI_d}{dV_{gs}} = \frac{\Delta i_{ds}}{\Delta v_{gs}} \approx -\frac{2I_{dss}}{V_c}(1 - \frac{V_{gs}}{V_c})$
- For circuit on right, $g_m \Delta v_{gs} = \Delta i_{ds}$ and so $g_m R_S \Delta v_{gs} = V$
- V_c is cutoff voltage. When $v_{gs} < V_C$ there's no channel conduction. Some people call this V_T or V_P. JFET input impedance is high ($10^{10}\Omega$).
- For J309, $V_c \approx -2.6V$, $i_{dss} \approx 23mA$, $g_m \approx 12$.
- DC: $V_b = -i_b R_S$, AC: $V = R_S g_m V_{gs}$, $v_{gs} = V_g - V$
- $V = R_S g_m v_{gs}$
- $v_{gs} = V_i - V, V = \frac{RV_i}{R + \frac{1}{g_m}}$,
- $G_v = \frac{V}{V_i} = \frac{Rg_m}{1 + Rg_m} \approx 1$
- $Z_0 = \frac{1}{g_m}$





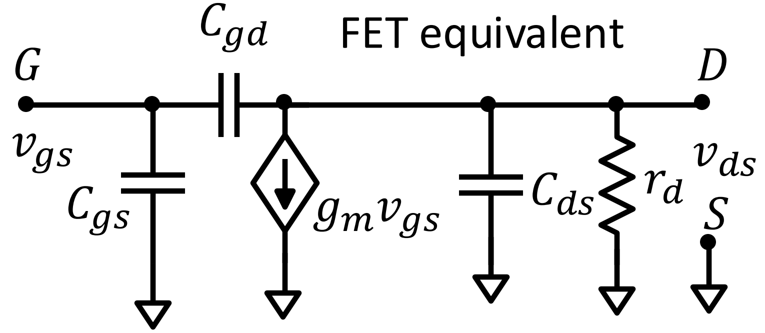
Region	Characteristic
ohmic	i_d linear in v_{ds}, v_{gs} <0
active	v_{gs} controls i_d linearly, v_{gs} <0
breakdown	v_{ds} is so high channel breaks
Pinch-off	V_{gs}<<0 and i_d=0 independent of v_{ds}



AC model

FETs

- $\Delta i_d = \frac{\partial i_d}{\partial v_{GS}} \Delta v_{GS} + \frac{\partial i_d}{\partial v_{DS}} \Delta v_{DS}$
- $\frac{\partial i_d}{\partial v_{GS}} = g_m, \frac{\partial i_d}{\partial v_{DS}} = g_d, r_d = \frac{1}{g_d}$
- $i_d = g_m v_{gs} + \frac{v_{ds}}{r_d} \approx g_m v_{gs}$, since r_d is large
- $Z_{in} = 10^9 \Omega$
- $Z_{out} = R_D || r_d || \frac{1}{j\omega(C_{GS}+C_{DS})}$

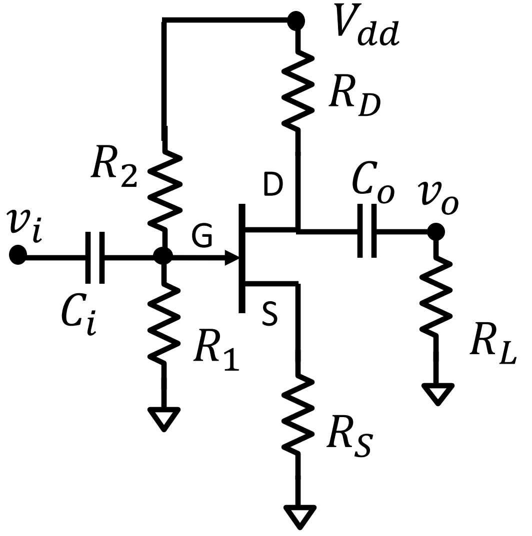


More on FETs

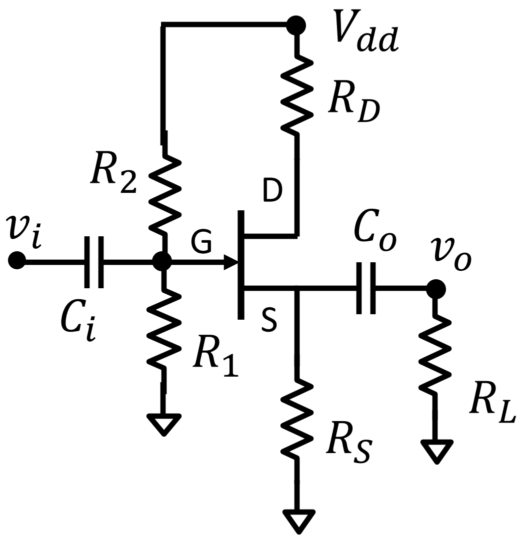
- For common source
  - $R_o = [r_d + R_S(1 + g_m r_d)]||R_D$, if $r_d \gg R_S, R_D, R_o \approx R_D$
  - $R_i = R_G = R_1||R_2$
  - $v_i = v_{gs} + i_d R_S = v_{gs}(1 + g_m R_S)$
  - $v_o = -i_d(R_D||R_L) = -g_m v_{gs}\,(R_D||R_L)$
  - $A_v = -\frac{R_D||R_L}{R_S + {}^1/_{g_m}}, A_i = -\frac{R_G}{R_S + {}^1/_{g_m}} \frac{R_D}{R_D + R_L}$
  - $R_G = 10^6\Omega, R_S = 10^4\Omega, R_D = 25k\Omega,\ g_m = 2000\mu S, g_d = 20\mu S = \frac{1}{r_d}$

- For common drain
  - $R_i = R_G = R_1||R_2$
  - $i_o = \frac{v_{gs}}{R_S} + g_m v_{gs},\ R_o = \frac{i_o}{v_{gs}} = \frac{1}{R_S} + g_m$
  - $v_0 = g_m v_{gs}(R_S||R_L), v_i = v_{gs} + g_m v_{gs}(R_S||R_L)$
  - $A_v = -\frac{g_m(R_S||R_L)}{[1 + g_m(R_S||R_L)]}, A_i = -\frac{R_G}{R_S + R_L} \frac{R_S}{[(R_S||R_L) + {}^1/_{g_m}]}$
  - $R_o = \frac{1}{\frac{1}{r_d} + \frac{1}{R_S} + g_m} \approx 196\Omega,$
  - $R_G = 10^6\Omega,\ , R_D = 100k, R_S = 10^4\Omega, R_L \approx 10^6\Omega$

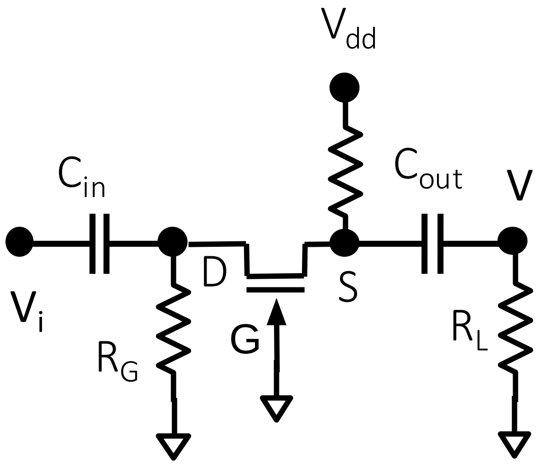
FET common source



FET common drain



JFET common gate amplifier



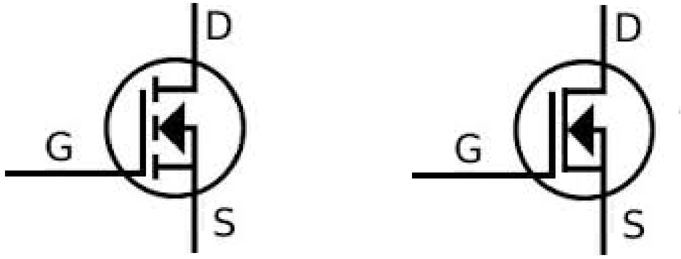
- $A_V = g_m(R_D||R_L), Z_{out} \approx r_0(g_m R_S + 1)||R_D, Z_{in} = R_S||\frac{1}{g_m}$
- $V_{DD} = 12V, i_{dss} = 60mA, V_P = -6, A_V = 10, R_L = 1k\Omega, R_S = 50\Omega$
- $i_{d,q} = \frac{V_P}{2{R_S}^2 i_{dss}}\left(V_P + \sqrt{{V_P}^2 - 4R_S i_{dss} V_P}\right) - \frac{V_P}{R_S}$

1. Solve for R_D: $10 = g_m \times R_D||R_L$, $R_D = 2k\Omega$
2. Find $i_{d,q} = \frac{V_P}{2{R_S}^2 i_{dss}}\left(V_P + \sqrt{{V_P}^2 - 4R_S i_{dss} V_P}\right) - \frac{V_P}{R_S} = 10\text{mA}$

Mosfet

- $r_{out} = \frac{1}{\lambda i_d} = \frac{\partial i_{ds}}{\partial V_{ds}}, g_m = \frac{\partial i_d}{\partial V_{gs}}$
- Weak Inversion ($V_{gs} < V_{th}$)
  - $i_d = i_0 \exp(\frac{V_g - V_{th}}{nV_t}), n = 1 + \frac{C_{th}}{C_{ox}}$
- Linear ($V_{gs} > V_{th}, V_{ds} < V_{gs} - V_{th}$)
  - $i_d = \mu_n C_{ox} \frac{W}{L} [(V_{gs} - V_{th}) V_{ds} - \frac{{V_{ds}}^2}{1}](1 + \lambda V_{ds})$
- Saturation ($V_{gs} > V_{th}, V_{ds} \geq V_{gs} - V_{th}$)
  - $i_d = \mu_n C_{ox} \frac{W}{2L} (V_{gs} - V_{th})^2 [1 + \lambda(V_{ds} - V_{dsat})]$

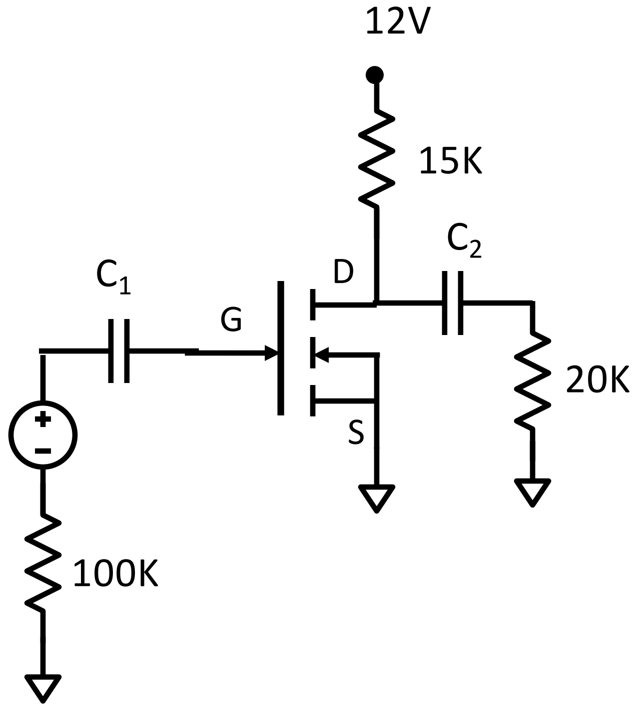
N channel MOSFET

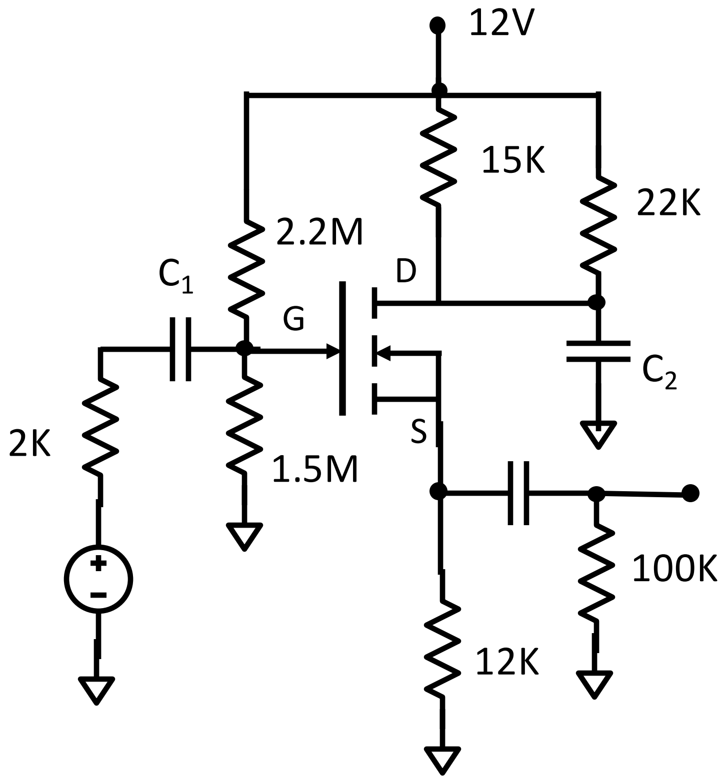


Enhancement type

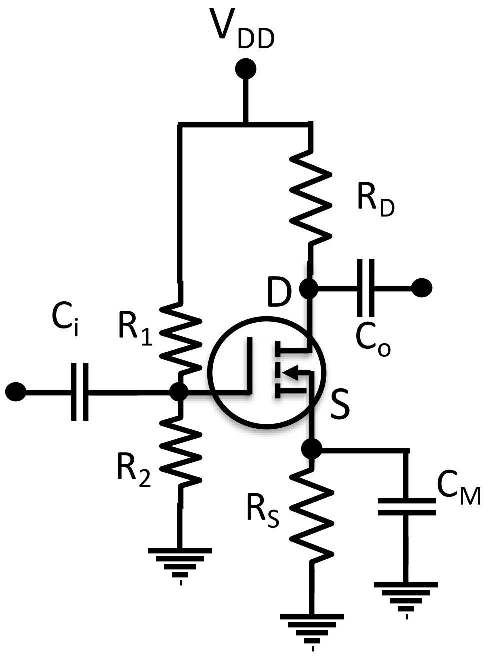
Depletion type

Mosfet amps


12V
15K
C1
C2
D
G
S
20K
100K


12V
15K
22K
2.2M
C1
D
G
C2
S
2K
1.5M
100K
12K

CMOS common emitter amplifier



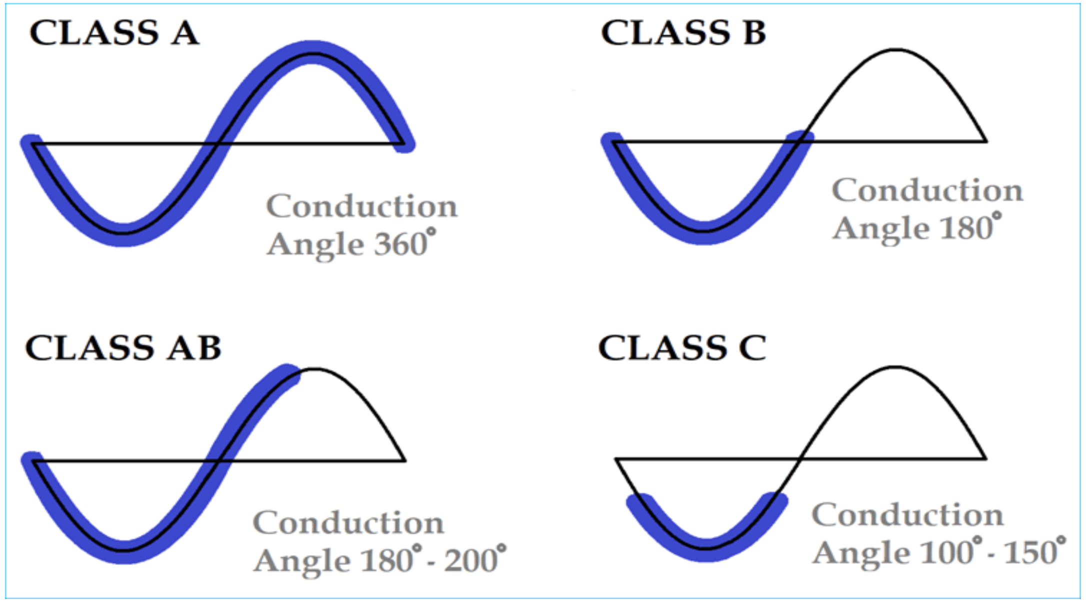
- Pick power
- $V_{DD} = i_D R_D + V_{DS} + i_D R_S$
- $V_{GS} = V_G - i_S R_S$
- $V_G = V_{DD} \frac{R_1}{R_1 + R_2}$
- $i_D = k(V_G - V_{TH})^2$
- Bias around $\frac{V_{DD}}{3}$
- Pick gain, $A = \frac{R_D}{R_S + \frac{1}{g_m}}$
- For 2N7000, $g_m \approx 200$

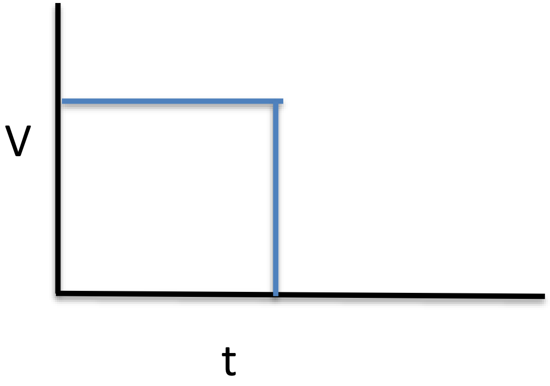
Amplifier classes

Class	Efficiency	Characteristics
A	35%	Full bias
B	60%	Low bias
C	75%	Saturating
D	75%	Switch in pass-band
E	90%	Voltage switch
F	80%	Harmonic resonators

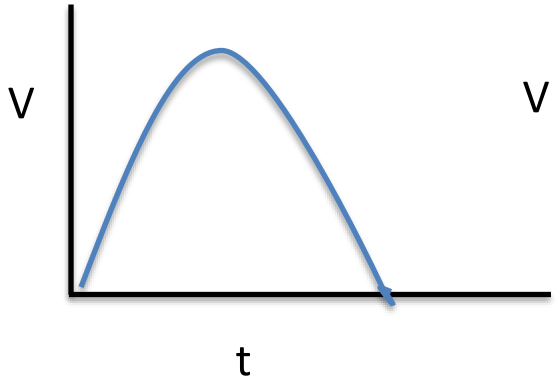
$$\eta = \frac{P}{P_0}, P_d = P_0 - P_i$$

$$P_d = P_a + P_{on}$$

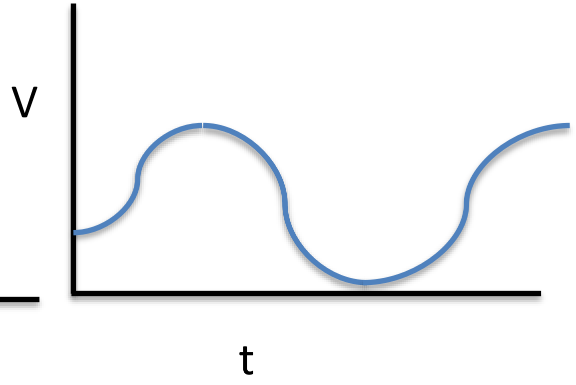




Class D



Class E



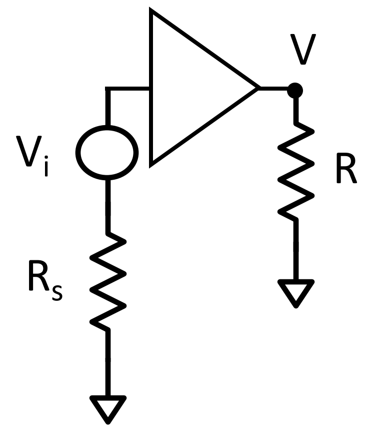
Class F

Efficiency of class A amplifiers

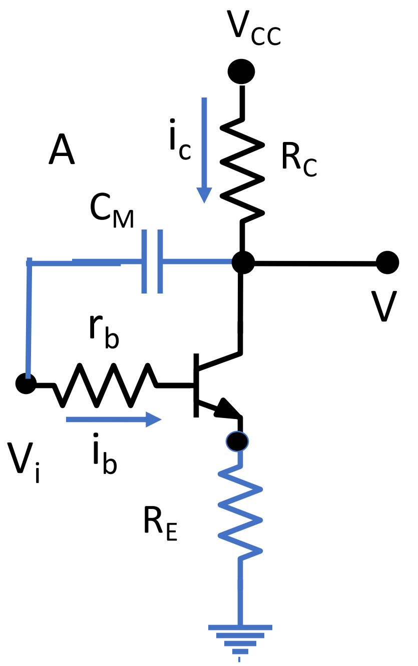
- Here, P_o is the power from the supply, P_d is the dissipated power, P is the output power. R is the collector resistor.
- $\eta = \frac{P}{P_0}$, P_0 is DC power
- $P_0 = V_{cc} I_0$, where $I_0 = < i_c >$, so $I_0 = \frac{V_{cc}}{2R}$ (R is the collector resistance). Thus, $P_o = \frac{{V_{cc}}^2}{2R}$.
- AC load power is $P = \frac{V_{pp} I_{pp}}{8} = \frac{{V_{cc}}^2}{8R}$. So maximum efficiency $\eta = \frac{P}{P_o} = 25\%$.
- DC load power is $\frac{{V_{cc}}^2}{4R}$ and so is transistor power.
- Half the power in a class A is lost to load resistance. If we replace resistance with transformer, $P_0 = \frac{{V_{cc}}^2}{R'}$, where R' is the effective load resistance and $P = \frac{V_{pp} I_{pp}}{8} = \frac{{V_{cc}}^2}{2R'}$, giving 50% efficiency. Transformer turns ratio controls peak-to-peak current. Maximum current is $I_m = \frac{2V_{cc}}{R'} = \frac{2V_{cc}}{n^2 R'}$, where n is the turns ratio.

Amplifier gain

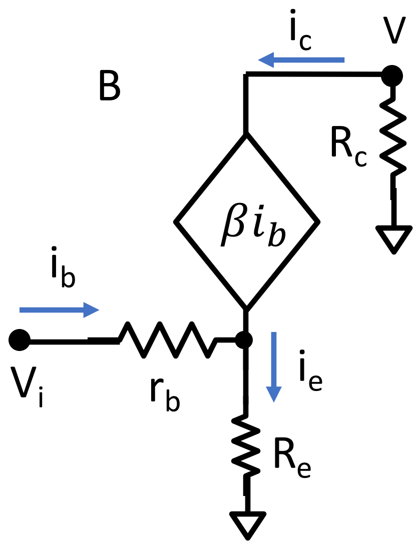
- Let P_+ be the maximum input power (when load is matched) and V_+ is the voltage at maximum power. $V_+ = \frac{V_0}{2}$
- $G = 10\log(\frac{P}{P_+})$
- $P = \frac{{V_{pp}}^2}{8R}$
- $P_+ = \frac{{V_{+,pp}}^2}{8R_S}$
- $G_S = 10\log(\frac{V}{V_+})$



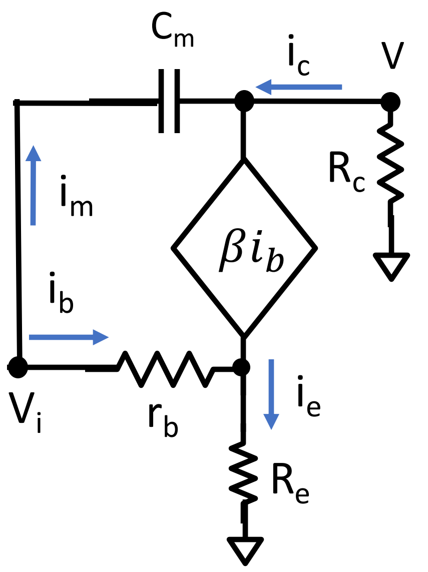
Emitter degeneration



- To the usual transistor circuit (A), on the right, we add R_E. (B) is an equivalent circuit.
- $V_{bb} \approx V_f + i_c R_E$. Let V be the output AC and V_i be the input AC.
- The gain is $G_v = \frac{V}{V_i}$.
- $V_i = i_b r_b + i_E R_E \approx i_C R_E, Z_i = \frac{V_i}{i_b}$,
- $V = -i_c R_C$.
- So $G_v = -\frac{R_C}{R_E}$ (Doesn't depend on β).
- $V_i \approx \beta i_b R_E$
- $Z_i = \frac{V_i}{i_b}$, so $Z_i = \beta R_E$.

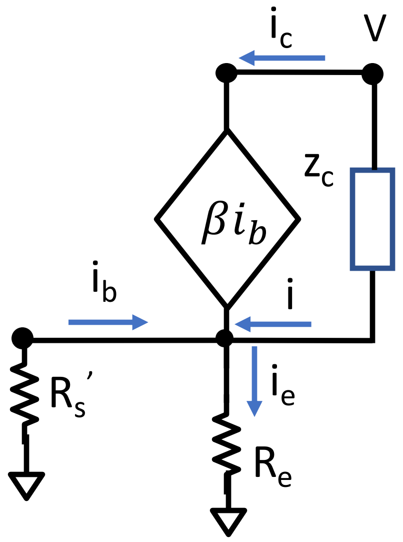


Emitter degeneration



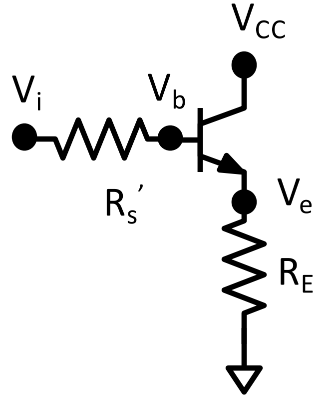
- C_M is called a Miller capacitor and it arises from a capacitance, C_C, between the collector and emitter. $i_m = j\omega C_C(V_i - V) = j\omega C_M(1 + |G_v|)V_i$. i_m is the maximum current between base and emitter in the equivalent circuit on the right.

- With the Miller capacitor, $Z_i = \beta R_E||(1 + |G_v|)C_M$
- $r_c \approx \frac{V_{early}}{i_c}$, r_c is the collector resistanc.e
- $R_s{}' = R_S + r_b$, r_b is the base resistance. $R_s{}'$ is the combined source resistance.
- z_C is called the collector impedance and $z_c = r_c||C_C$, C_c is specified in data sheet (8pF).
- $Z_o = \frac{V}{i_C}$, $i = i_c - \beta i_b$,
- $i_b = -\frac{i_C R_s}{R_s{}' + R_E}$,
- $i = i_c(1 + \frac{\beta R_E}{R_s{}' + R_E})$
- $V = iz_c + i_C\,(R_s{}'||R_E)$
- $Z_o = \frac{V}{i_C} = z_C\left(1 + \frac{\beta R_E}{R_s{}' + R_E}\right) + R_s{}'||R_E$.
- $|z_c| \gg R_E$, so $Z_o = z_C\left(1 + \frac{\beta R_E}{R_s{}' + R_E}\right)$



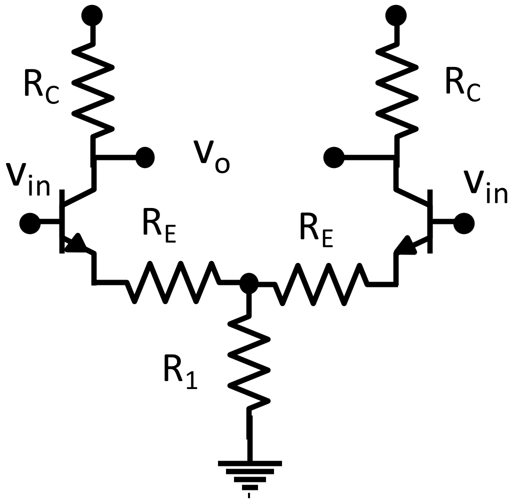
More on emitter follower

- $Z_0 = \frac{v_e}{i_e}$
- $v_b = -R_s{}' i_b, R_s{}' = R_s + r_b$
- $i_e \approx \beta i_b$
- $Z_0 \approx \frac{R_s{}'}{\beta}$



Differential Amplifier

- Two port model
- $\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \begin{pmatrix} V_1 \\ V_2 \end{pmatrix}$

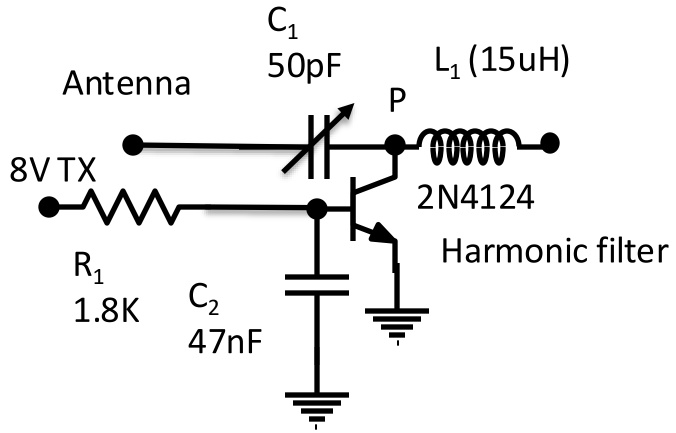


Differential amplifier

- Pick power 12
- Choose collector current ($2mA$) by picking R_1
- Pick gain, $A = \frac{R_C}{2R_E}$
- $G_d = -\frac{R_c}{R_e}$
- $Z_d = 2R_c$

Exercise 19: Norcal receiver switch

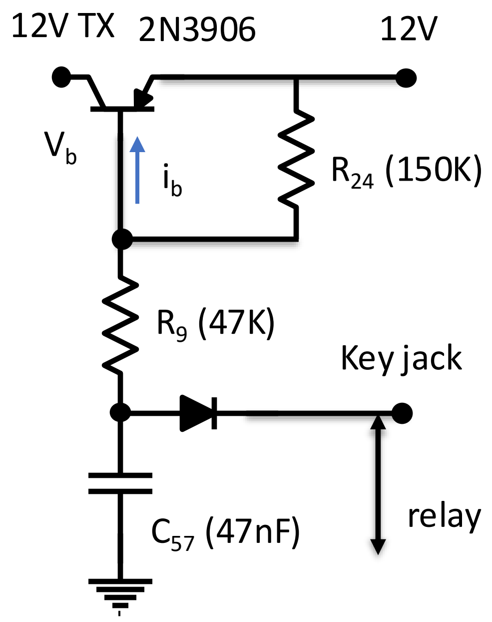
1. Consider the rising part of the base voltage waveform. Calculate slope.
2. Do the same for the falling part for voltage below .6V. Calculate t_2.
3. Measure the switch attenuation
   - When the transistor is saturated, the drop across ce is 1.4V. At full power, $P = \frac{{V_m}^2}{8R}$ and $V_m = 33.9V$. $\frac{P_{new}}{P_{original}} = \frac{1}{33.9^2}$, so $loss = 10\log\left(\frac{1}{33.9^2}\right) = -31dB$.
4. Measure the voltage with the switch on. Measure output voltage and calculate on-off rejection ratio R=20 log(V_{off}/V_{on})



Exercise 20: NorCal transmitter switch

- For transmitter switch, saturation resistance is $\approx 2\Omega$, $G_s = \frac{i_b}{15mV}$.
- i is current into the load. In Norcal, $i = 7mA$. For 2n3906, to ensure saturation, $i_b = \frac{2i}{100} = 140\mu A$.

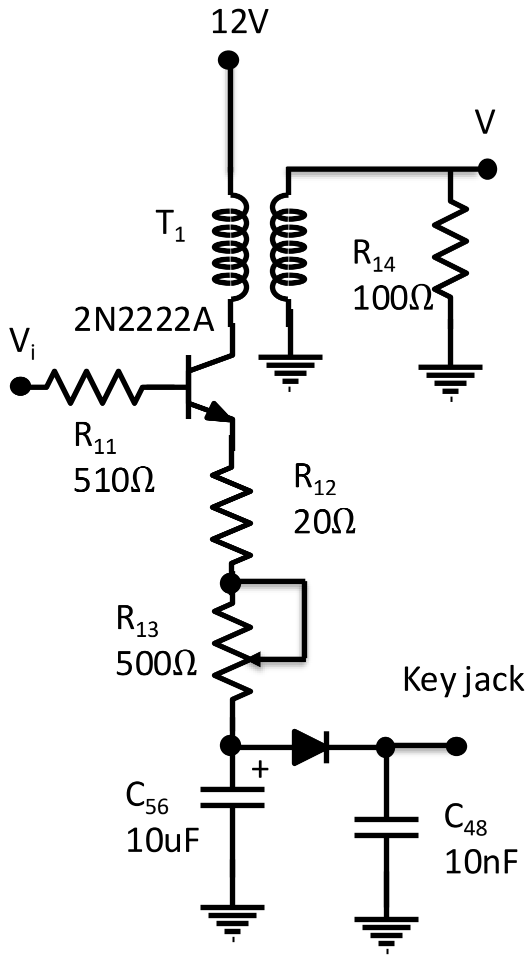
1. Calculate voltage on C_{57}. Measure time for capacitor to charge half-way. Calculate what the time should be.
   - $\tau = 197 \times 10^4 \times 47 \times 10^{-9} = 9.2 \times 10^{-2} sec = 92 msec.$
2. Calculate the approximate current i_c when Q_4 is on. Assume base voltage on Q_1 is 700 mV. Neglect saturation voltage on Q_4. Calculate base current i_b required to produce this collector current assuming $\beta = 100$.
3. Calculate i_b at key down assuming a 700 mV drop-in base-emitter of Q_4 and at 600mV at D_{11}
   - $V_b = \frac{R_9}{R_9 + R_{24}}(12) \approx 3V$, $i_b = i_{bs}\exp(\frac{V_b}{V_t})$,
4. Sketch collector voltage at Q_4 showing where transistor is saturated. What is the delay in going active?
5. Use the delay to measure β.



Exercise 21: Norcal Driver

1. Measure the output voltage and calculate the power, P.
2. Calculate the power from the power supply.
3. Measure the voltage gain $G_v = \frac{V}{V_i}$ with R13 at minimum and maximum gain.

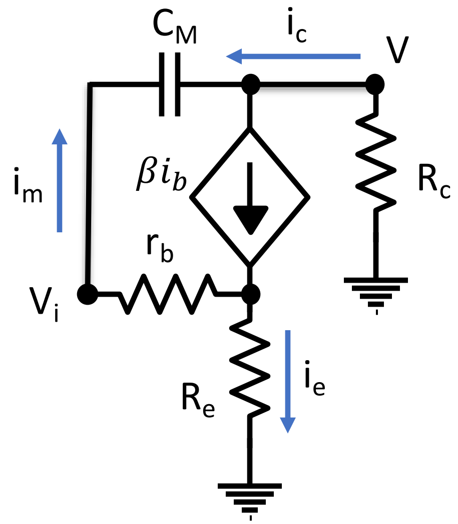
- $\omega = 4.4 \times 10^7, L_{p,T1} = 68.6\mu H$. This is a class A amplifier.
- $R' = n^2 R, n = \frac{14}{4}, \ R' = 1225\Omega$
- $Z_{eq}(R) = (20 + R) + j\omega L_{p,T1}, 0 \leq R \leq 500. R = R_{13}$
- $Z_{eq}(0) = 20 + 2992j,$
- $i_c = \frac{V_{cc} - V_{ce}}{20 + R'}, i_c = \beta i_b, V_e = \ 20 i_c$
- From text, $P = \frac{(V_{cc} - V_e)^2}{2R'}$
- $P_o(R) = \frac{{V_{cc}}^2}{(20 + R + R')}$
- Gain is between 2.5 and 60



Exercise 22: Emitter degeneration

- In Driver amplifier, add probe to R_{11}, this allows us to measure the AC voltage, V_i

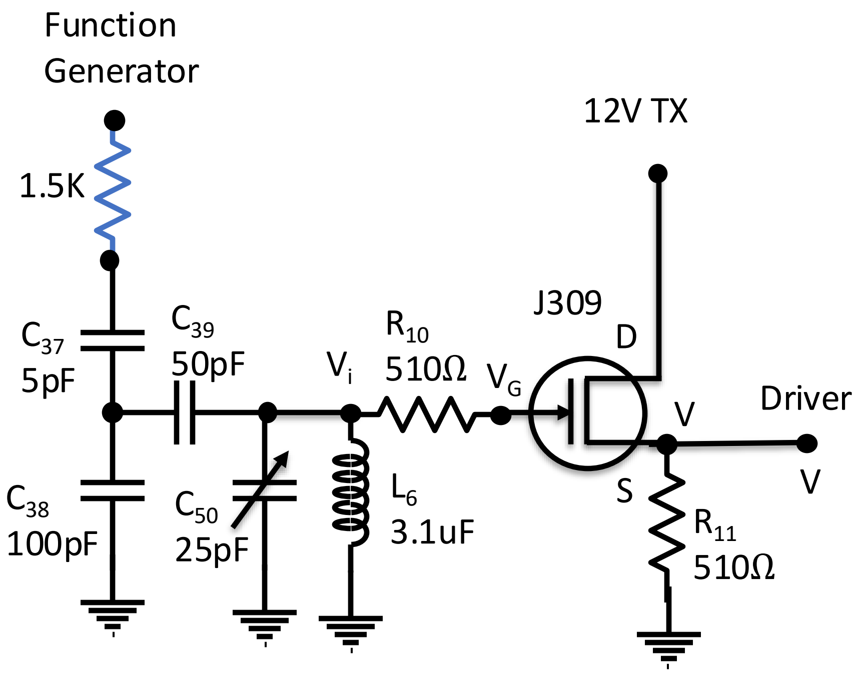
1. Measure $G_v = \frac{V}{V_i}$ with R_{13} turned fully counterclockwise and then fully clockwise.
   - $R' = 1225\Omega$
   - When R_{13} is fully counter-clockwise $R_{E,effective} = 520\Omega$, $G_v = \frac{1225}{520} = 2.36$
   - When R_{13} is fully clockwise $R_{E,effective} = 20\Omega$, $G_v = \frac{1225}{20} = 61$
   - Calculate the expected voltage gain for each setting of R_{13}
2. Measure V_i at the maximal gain setting
3. The open circuit voltage is $V_0 = 2V$, calculate V_i in terms of C_M
   - $Z_i = \beta R_E || (G_v + 1) C_M$, $V_i = Z_i i_b$, $V = -1225\, i_c$,
   - So, $\frac{V}{V_i} = -\frac{1225}{Z_i} \cdot \frac{i_c}{i_b} = -\beta \frac{1225}{Z_i}$



Exercise 23: Norcal Buffer amplifier

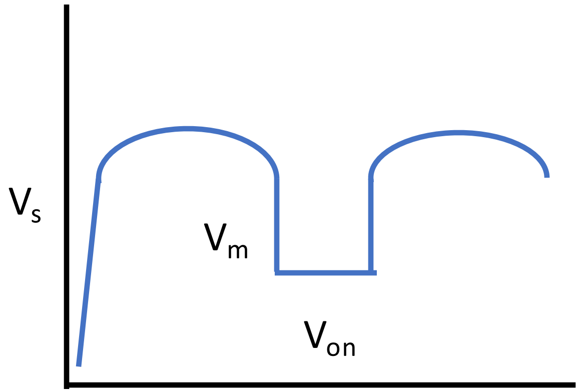
1. R_{11} determines the bias. Measure the DC voltage at source of the JFET (V).
2. Calculate the drain bias current. Calculate the source and drain voltages you should expect ($R = R_{11}$)
   - $V_{gs} = V_i - V$, $i_d = i_{dss}(1 - \frac{V_{gs}}{V_C})^2$, $V = g_m V_{gs} R$
   - $g_m \approx 12$, $i_{dss} = 23mA$, $V_C = -2.6V$
   - $V_{gs} = \frac{V_i}{1+g_m R}$, substitute into $i_d = i_{dss}(1 - \frac{V_{gs}}{V_C})^2$ to get i_d.

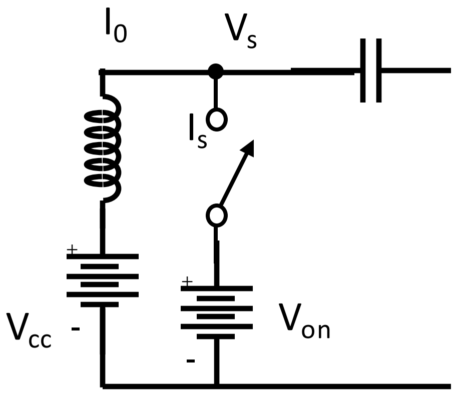
     $i_S = \frac{V}{R}$
3. Calculate and measure the voltage gain of the buffer.
   - $G_V = \frac{V}{V_i} = \frac{1}{1+\frac{1}{g_m R}}$, or about 1 since $g_m \approx 12$
4. Find the transconductance using the measured voltage gain.
5. Calculate the available power P_+ from the function generator through a $1.5k\Omega$load. Calculate gain in dB.



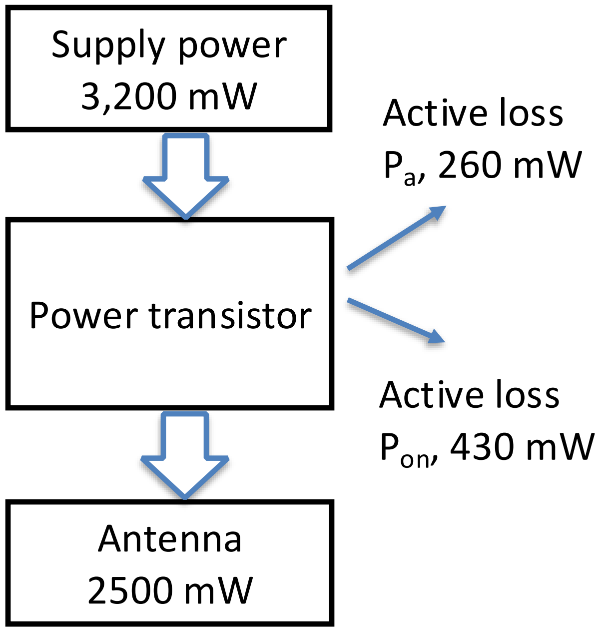
Class C amplifiers and Norcal 40 Power amp

- Here P_o is the power from the supply, P_d is the dissipated power, P is the output power. The Norcal Power amplifier is a class C amplifier. For switch model, switch represents the transistor, when the transistor is on, the switch is open.
- $V_s = V_{on} + V_m \cos(\omega t)$, (switch off), V_{on} (switch on)
- $V_{cc} = V_{on} + \frac{V_m}{\pi}, V_m = \pi(V_{cc} - V_{on})$
- $P_0 = V_{cc} I_0, P_d = V_{on} I_0$
- $P = P_o - P_d = \frac{(V_{cc} - V_{on})}{\pi}$
- $\eta = \frac{P}{P_0} = \frac{(V_{cc} - V_{on})}{V_{cc}}$
- $P = \frac{{V_m}^2}{8R}$, R is input filter impedance
- $P_d = P_0 - P = 3.2W - 2.5W = 700mW$
- Cap energy: $E = \frac{CV^2}{8R} = 37nJ$
- $P_a = Ef = 260mW$
- $i_c = i_0 - i_c = 215mA$
- $P_{on} = V_{on} i_{on} = 430mW$
- $P_d = P_a + P_{on} = 690mW$



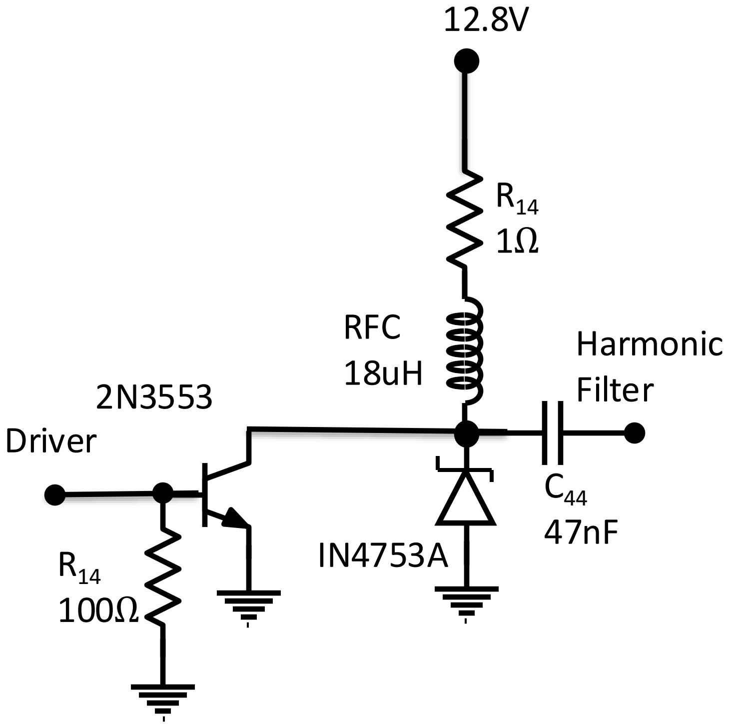


Switch model for class C amp



Exercise 24: Norcal Power Amp

1. Measure the peak-to-peak voltage across 50ohm load required for output of 2W. Calculate it and compare. Calculate the gain in dB
   - $V_{cc} = 12.8V$, $\mathrm{R} \approx 50\Omega$, $I_0 = 250mA$
   - $P_{on} = V_{on}I_{on} = 430\mathrm{mW}$, $P_a = Ef = 260mA$, $P_d = P_a + P_{on} = 690mW$
   - $P_o = V_{cc}I_0 = 3.2W$.
   - $P = \frac{(\pi(V_{cc}-V_{on}))^2}{8R} = 2.6W$
2. Find pp output voltages or 5, 10, 15, 20, 25 and 30V. Calculate power supply current subtracting 2mA for regulator
3. Calculate the output power, efficiency and and dissipation power.
   - $P_d = P_0 - P = 3.2W - 2.5W$
   - $\eta = \frac{P}{P_0} = \frac{2.5}{3.2} = .78$



Thermal modelling

- T is heat sink temperature, T_0 is ambient temperature, P_d is power dissipated.
- $R_t = \frac{T - T_0}{P_d}$, R_t is the thermal resistance
- $C_t \dot{T} = P_d$, C_t is the thermal capacitance
- $R_j = \frac{T_j - T}{P_d}$, T_j is the junction temperature
- $f(t) + \tau \dot{f(t)} = f_\infty, f(t) = f_0 e^{-\frac{t}{\tau}}$
- $P_d = \frac{T(t) - T_0}{R_t} + C_t \dot{T(t)}, \tau = C_t R_t, T_\infty = P_d R_t + T_0$
- $T(t) + \tau \dot{T(t)} = T_\infty, \tau = C_t R_t.$
- $T_\infty = P_d R_t + T_0$
- $T(t) = T_\infty - P_d R_t e^{-\frac{t}{\tau}}$
- $T_j(t) = T(t) + R_j P_d$

Exercise 25: Thermal modelling

- For Motorola 2N3553, $T_j = 25\ {}^{C}/_{W}$

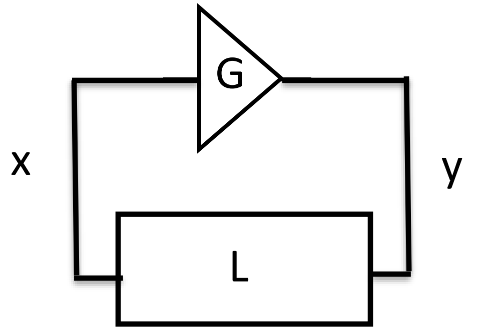
1. Measure ambient temperature
2. Turn function generator until output is $30V_{pp}$
3. After 20 minutes, measure T_∞. Use this to calculate R_t and T_j
4. Plot heat sink temperature vs time. Measure t_2 and calculate C_t

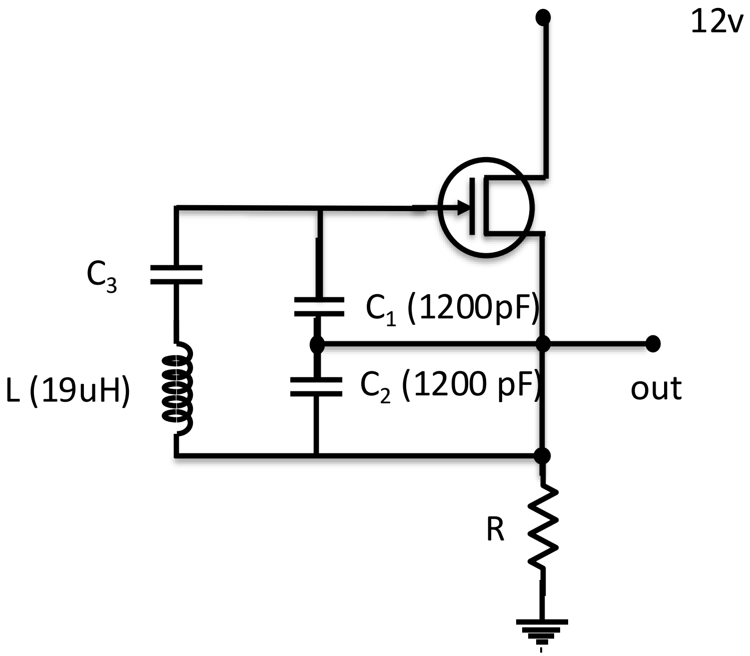
- Need measurements

Clapp oscillator

- Oscillation condition
  - $Gx = y$, Ly=x
  - $|G| = |L|$ and $\angle G = \angle L$

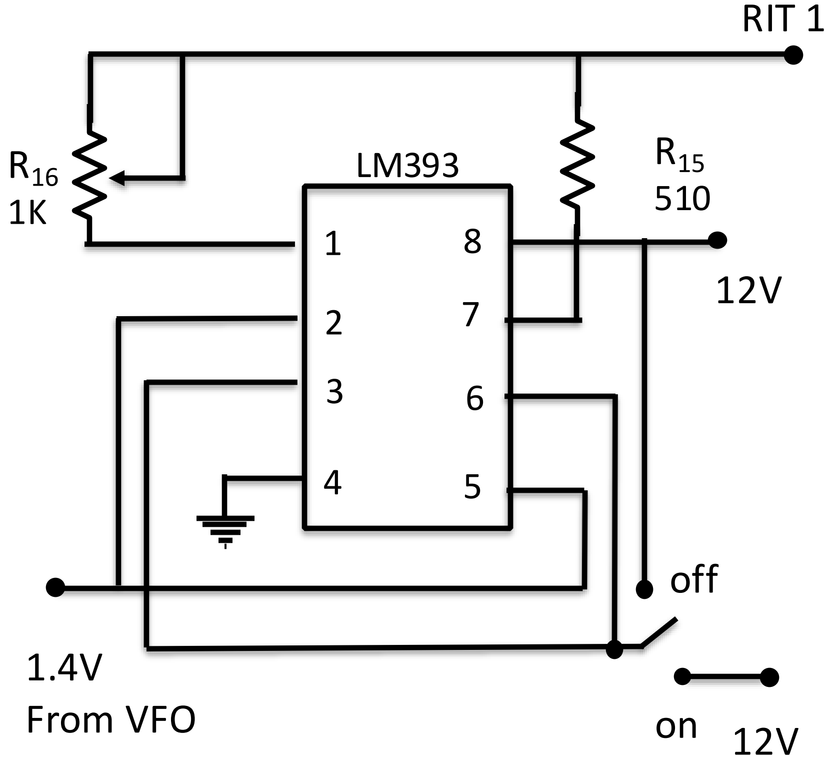
- Clapp (circuit on right)
  - $i_d = g_m v_{gs}$
  - Resonance: $-\frac{1}{j\omega_0 C_2} = j\omega_0 L + \frac{1}{j\omega_0 C_3} + \frac{1}{j\omega_0 C_1}$
  - $\omega_0 = \frac{1}{\sqrt{LC}}$, $C = C_1||C_2||C_3$
  - At resonance, $v_{gs} = Ri_d \frac{C_1}{C_2}$, $L = \frac{C_1}{RC_2}$
  - Oscillation continues if $g_m > \frac{C_1}{RC_2}$
  - $v_{gs} = 2v_s$





Norcal Receiver Incremental Tuning (RIT)

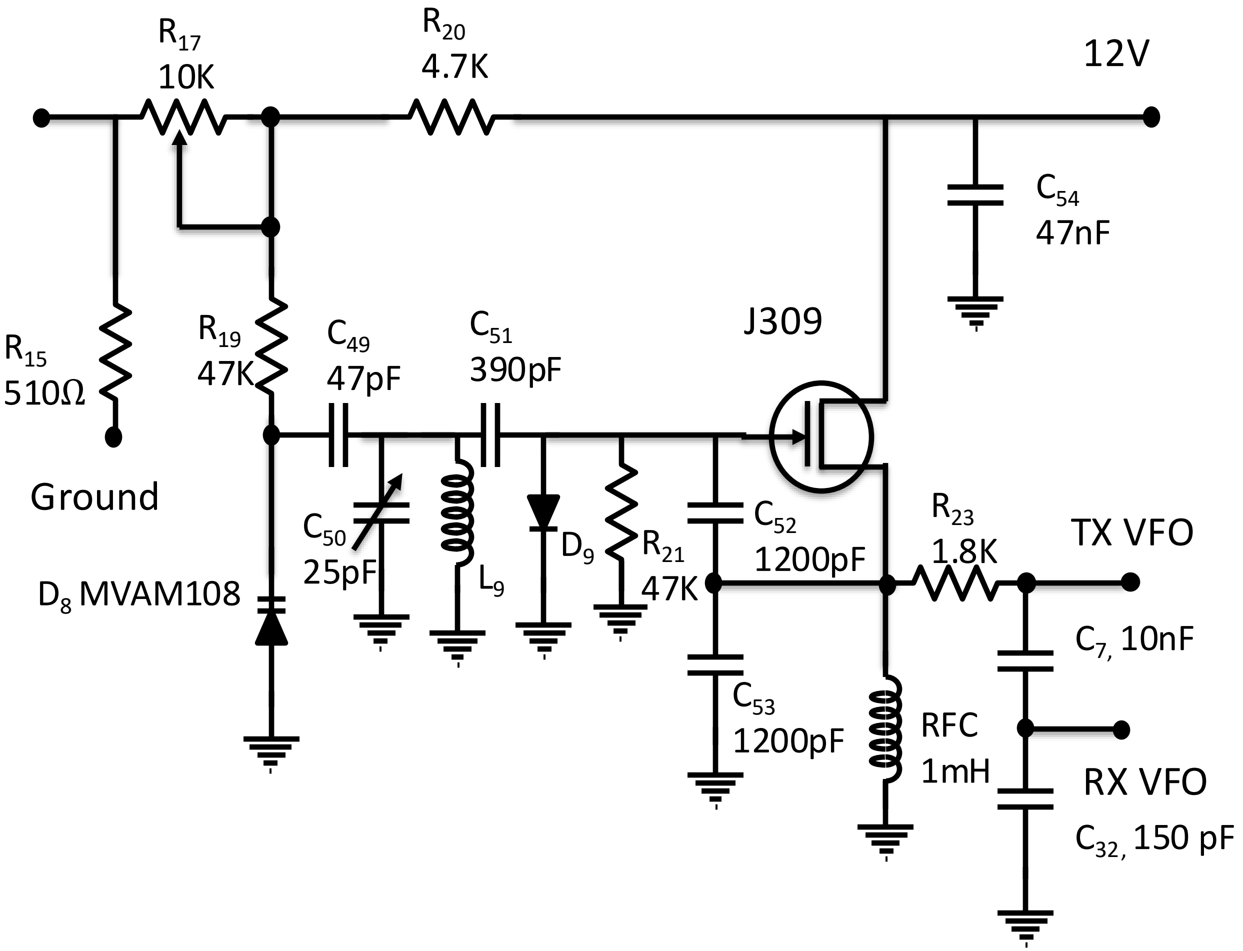
- LM393 is a comparator
- RIT allows transmit and receive frequency to be offset.
- If transmitter is on, TX will be 8V and the left comparator will be off, the right one on and R_{15} will be grounded.
- For receiving, TX is <1.4V, disconnecting R_{15} and shorting R_{16} to ground.



Exercise 26: Norcal VFO

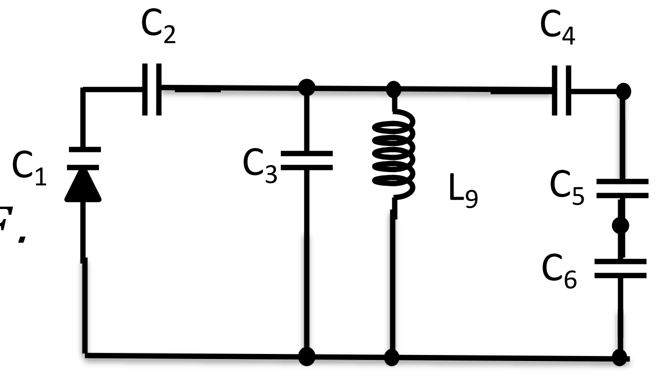
- Check MVAM108 capacitor when R_{17} is high and low
- Start resistor (R_{21}) pulls gate to ground at start
- When gain limiting diode (D_9) conducts, it pulls gate negative
- Oscillator keeps growing as long as g_m>1/R

1. Measure p-p voltage, V. What should you expect?
2. Measure DC voltage across wiper in R_{17}
3. Calculate expected V for large signal oscillation
4. How does the frequency change as R_{17} changes?
5. Calculate the oscillation frequency and the loss ratio $|V/V_l|$
6. How would this change if you took when L_9 is turned off?



VFO Problem

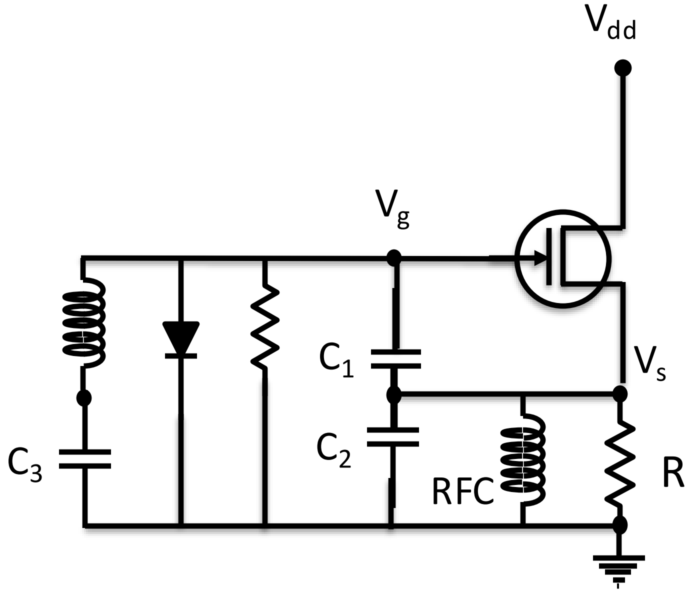
- The figure on the right is an equivalent circuit for the oscillator for the purpose of calculating resonant frequency. The varactor, C_1 varies from 30 to 600 pF depending on the voltage.
- $C_2 = 47pF, C_3 = 7pF, C_4 = 390pF, C_5 = C_6 = 1200pF, L_9 = 19.2\mu H$
- The equivalent capacitance for $C_4 - C_5 - C_6$ is $C_{R,eq} = 236pF$.
- When $C_1 = 187pF$, the equivalent capacitance for $C_1 - C_2 - C_3$ is $C_{L,eq} = 46.2pF$ and $C_{osc} = 282\text{pF}$. At the resonant frequency, $\omega L_9 = \frac{1}{\omega C_{osc}}$. $\omega^2 = \frac{1}{L_9 C_{osc}}$. $f_r = 2.16MHz$
- When $C_1 = 54pF$, the equivalent capacitance for $C_1 - C_2 - C_3$ is $C_{L,eq} = 32pF$. $C_{osc} = 268pF$ At the resonant frequency, $\omega L_9 = \frac{1}{\omega C_{osc}}$. $\omega^2 = \frac{1}{L_9 C_{osc}}$. $f_r = 2.22MHz$.



- These values are what we want for the tunable VFO.

Gain Limiting in Norcal 40

- The gain here is limited by the diode. $C_1 = C_2$. $V_g = 2V_s$
- $V_g = V_f - V$, V_f is the forward voltage of the diode.
- $V_m = V_g + \frac{V}{2}$, or $V_m = V_g - \frac{V}{2}$
- $G_m = \frac{I}{V} = \frac{1}{R}$
- $I_d = I_{dss}(1 - \frac{V_{gs}}{V_c})^2$
- $I_o = \frac{I_m}{4}$, $I \approx I_m$
- $G_m = \frac{I_m}{V}$
- Oscillation condition is $G_m = \frac{1}{R}$



Exercise 27: Gain limiting

1. Measure the voltage, V, on R_{23}
2. In deriving the oscillation condition, we neglected the inductor resistance and drain source resistance, r_d. How does this affect the conditions? L_9 has a Q of 250 and $r_d = 5k\Omega$, now what is the predicted V?
3. Calculate the loss ratio $|\frac{V}{V_i}|$.
4. Measure the temperature dependence of the VFO.
5. How much does the temperature have to change to cause a 100Hz shift?
6. What is the oscillation change if we remove one turn of the inductor?
7. What is the RIT tuning range?

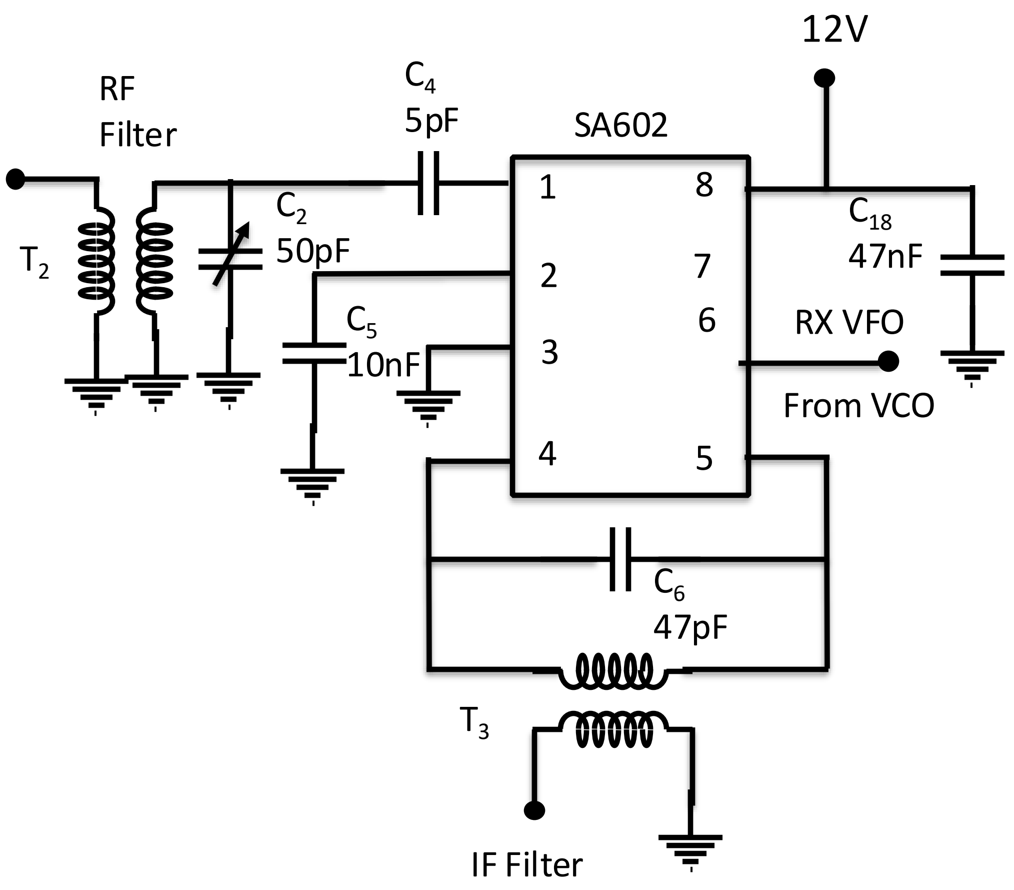
Gain limiting

- Need measurements

Exercise 28: Norcal RF Mixer

1. Measure conversion gain of the Mixer.
2. How much attenuation is provided by pot?
3. By how many dB is the image response suppressed. Look at the spur $f_{\downarrow 5}$

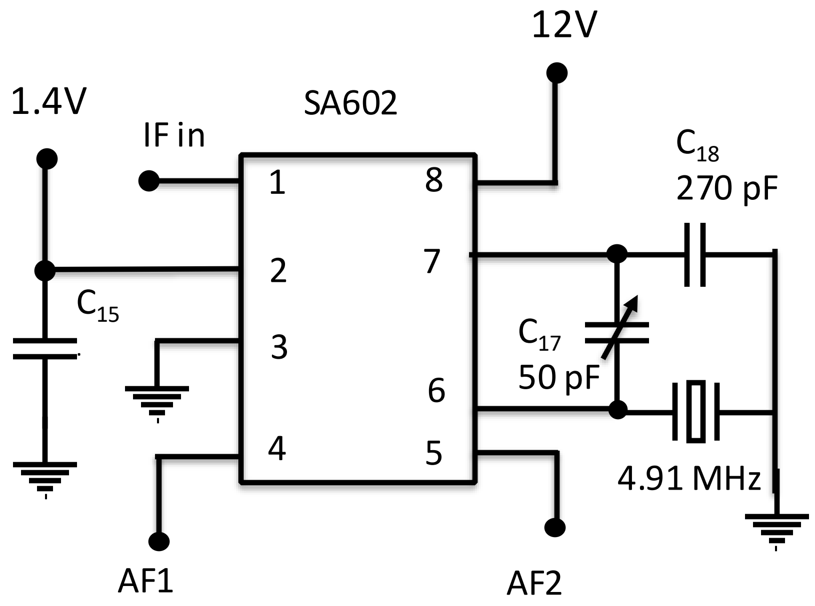
- Need measurements



Exercise 29: Norcal Product Detector (1)

1. Adjust C17 for minimum oscillation frequency and record it.
2. Calculate the minimum oscillation frequency you'd expect.
3. Measure the temperature coefficient for the BFO.
4. Measure the gain through the receiver from the antenna through the product detector.
5. Find the f5 spur calculate the expected f3.
6. By how much is the if spur suppressed?

- Need measurements

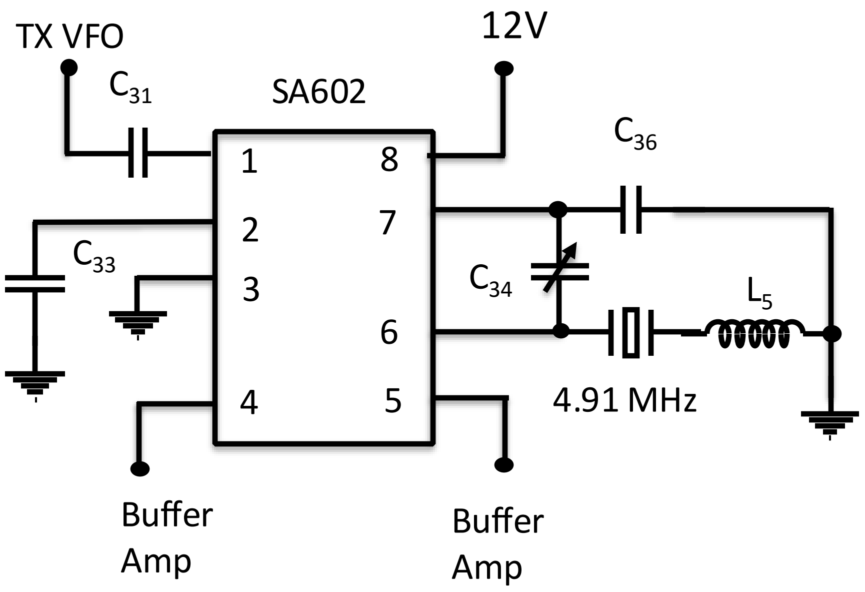


- 620 Hz output through AF1 and AF2

Exercise 30: Norcal transmit mixer and oscillator

1. How much would you expect the inductor to lower the oscillation frequency
2. Use the TX VFO and the voltage attenuation to calculate the input power from the transmit mixer. Calculate the gain through the entire chain
3. Measure the rise and fall time of keying response
4. There is a spurious $f_{mn}=mf_{vfo}+nf_{to}$.

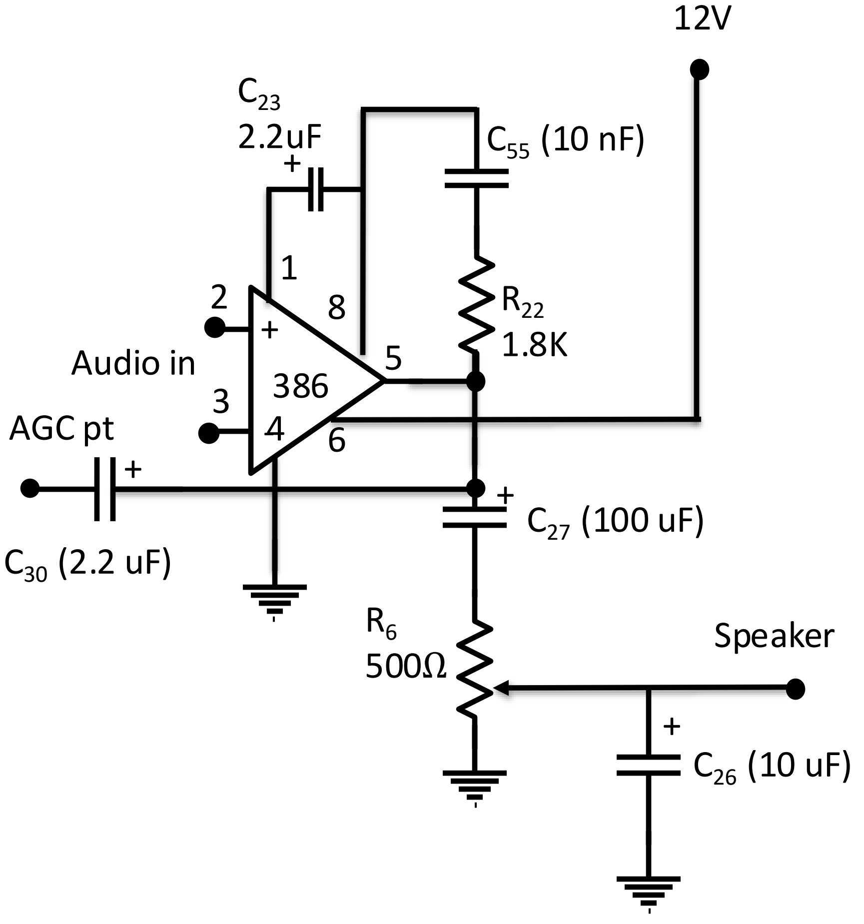
- Need measurements



Exercise 31: Norcal Audio Amp

1. Calculate input V_i assuming very high input impedance
2. Measure the voltage gain G_v at high frequency and 3dB roll-off

- Input impedance is high.
- The 386 acts like a non-inverting op amp. The internal feedback resistor is $R_f = 15k\Omega$. $G = 2\frac{R_f}{R_e}$. With pins 1, 8 open, $R_e = 1.5k\Omega$, so $G = 2\frac{15}{1.5} = 20$. pins 1 and 8 go across $1.35k\Omega$ of R_e. So, shorting them (using the non-inverting gain circuit) results in a gain of $G = 2\frac{15}{.15} = 200$.

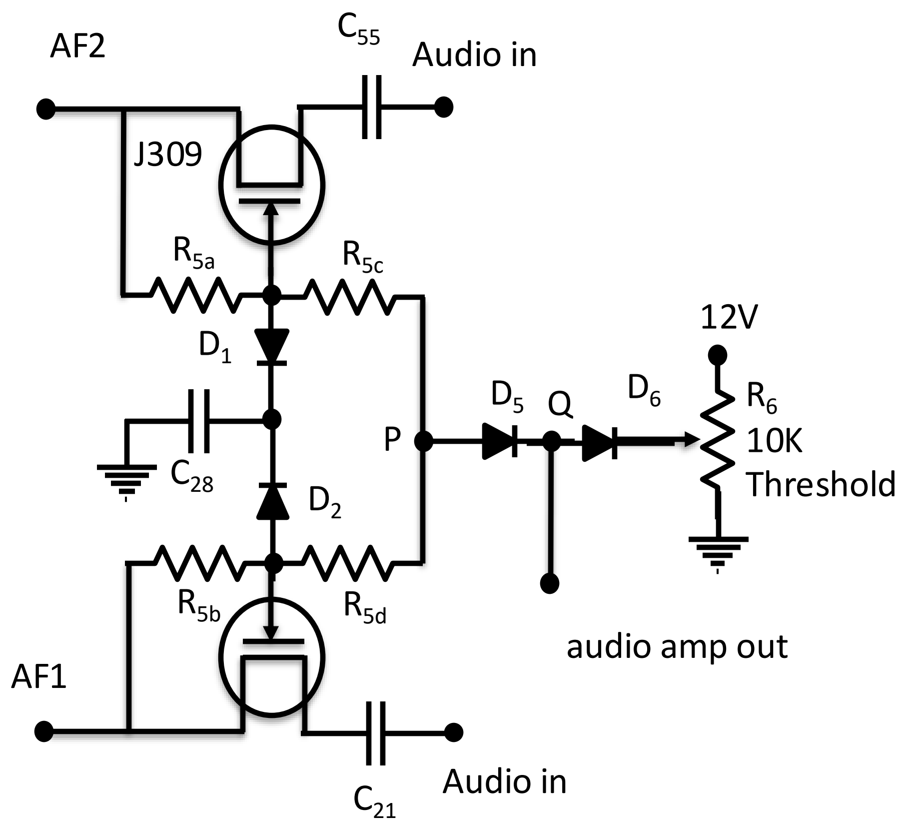


Exercise 32: Norcal AGC

- Connect function generator through 300K resistor to AF2 (620Hz sine, R_6 fully counterclockwise) and oscilloscope to audio output. Adjust input so output is 1V rms. Connect multimeter to P.

1. Plot audio output v dc control
2. What is the maximum control voltage we can measure? Infer cutoff voltage V_c
3. What is the minimum control voltage?

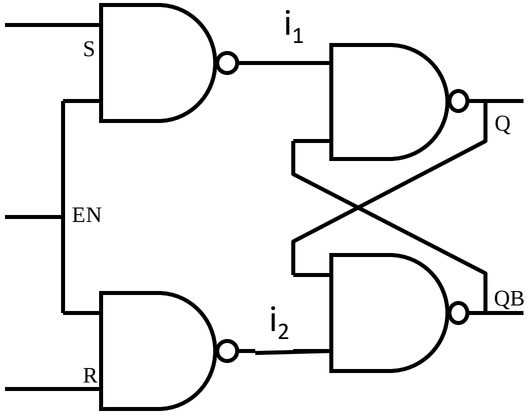
- Need measurements



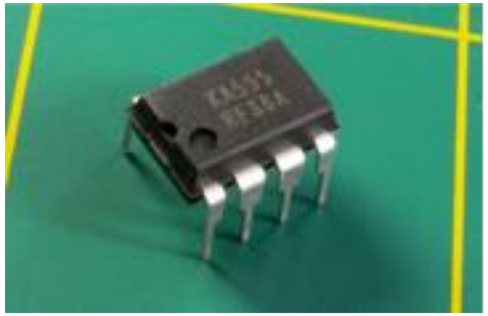
Logic symbols, flip flops, latches and shift registers

and or not and not or

- The circuit on the right is an S-R flip flop.
- QB = ¬Q, always. If EN=0, i_1=i_2=1, the other NAND input will be Q. ¬Q and Q will keep whatever values they had. If EN=1, S=1 and R=0, Q=1 and QB=0. If EN=1, S=0 and R=1, Q=0 and QB=1. The output is undefined if EN=1, and S and R have the same value.



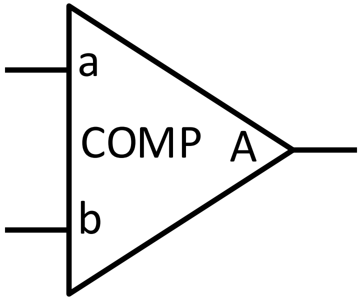
IC's: 555 Timer



Credit: Make Electronics

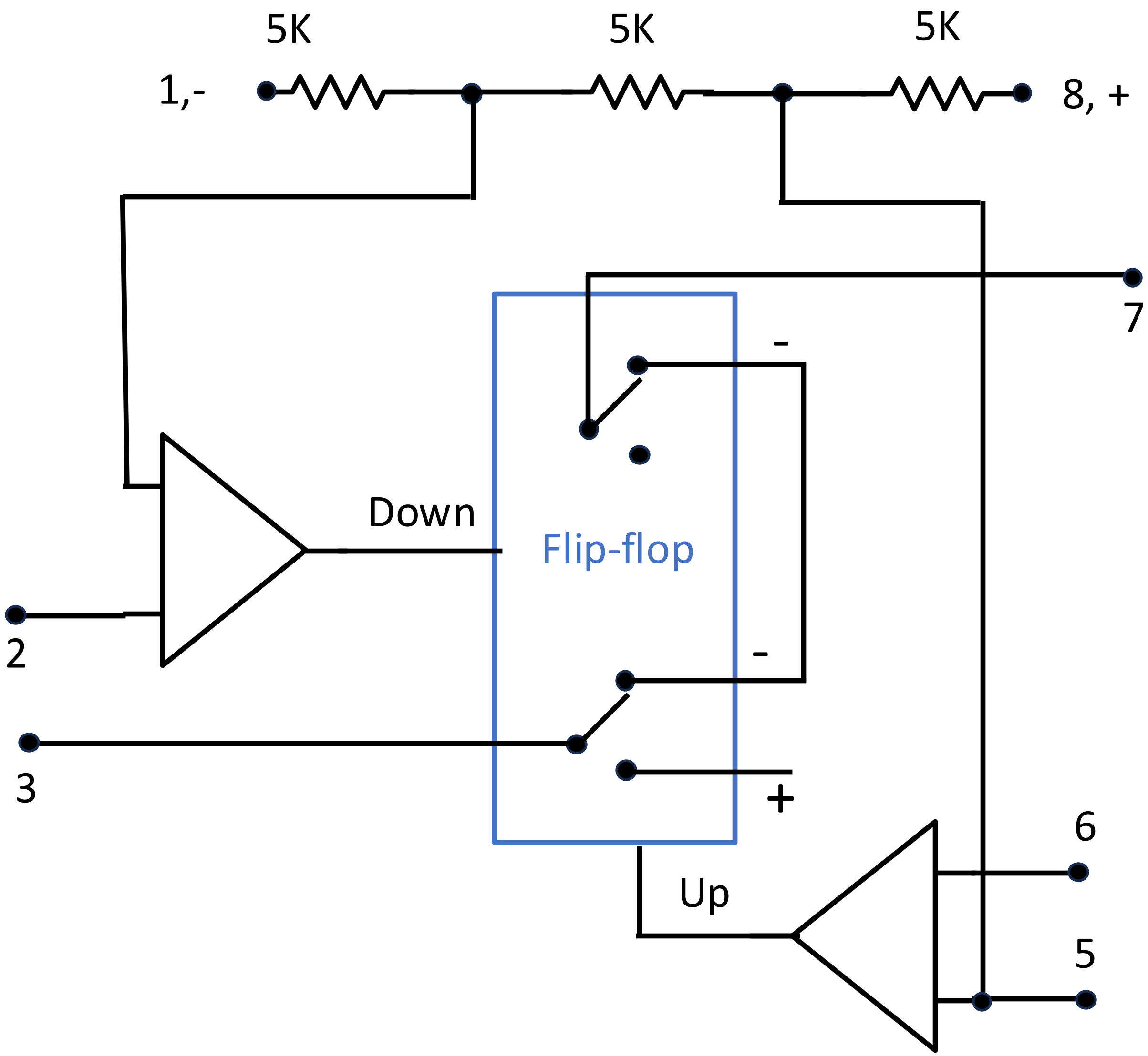
- Package**:** 8 pin dual inline package ("DIP")
  - Pin 1: -Vs
  - Pin 8: +Vs
  - Pin 2 (Trigger): Out HIGH if V < V_{CC}/3. Pin 2 has control over pin 6. If pin 2 is LOW, and pin 6 LOW, output goes and stays HIGH. If pin 6 HIGH, and pin 2 goes LOW, output goes LOW while pin 2 LOW. This pin has a very high impedance (about 10M) and will trigger with about 1uA.
  - Pin 3 (Output): (Pins 3 and 7 are "in phase.") Goes HIGH (about 2v less than rail) and LOW (about 0.5v less than 0v) and will deliver up to 200mA.
  - Pin 4 (Reset): Internally connected HIGH via 100k. Must be taken below 0.8v to reset the chip.
  - Pin 5 (Control): A voltage applied to this pin will vary the timing of the RC network (quite considerably).
  - Pin 6 (Threshold): HIGH if > 2 V_{CC} /3, make output LOW only if pin 2 is HIGH. Pin 6 has very high impedance (~10M) and will trigger with about 0.2uA.
  - Pin 7 (Discharge): Pin 7 is equal to pin 3 but pin 7 does not go high - it goes OPEN. When LOW it sinks about 200mA.

555 Equivalent

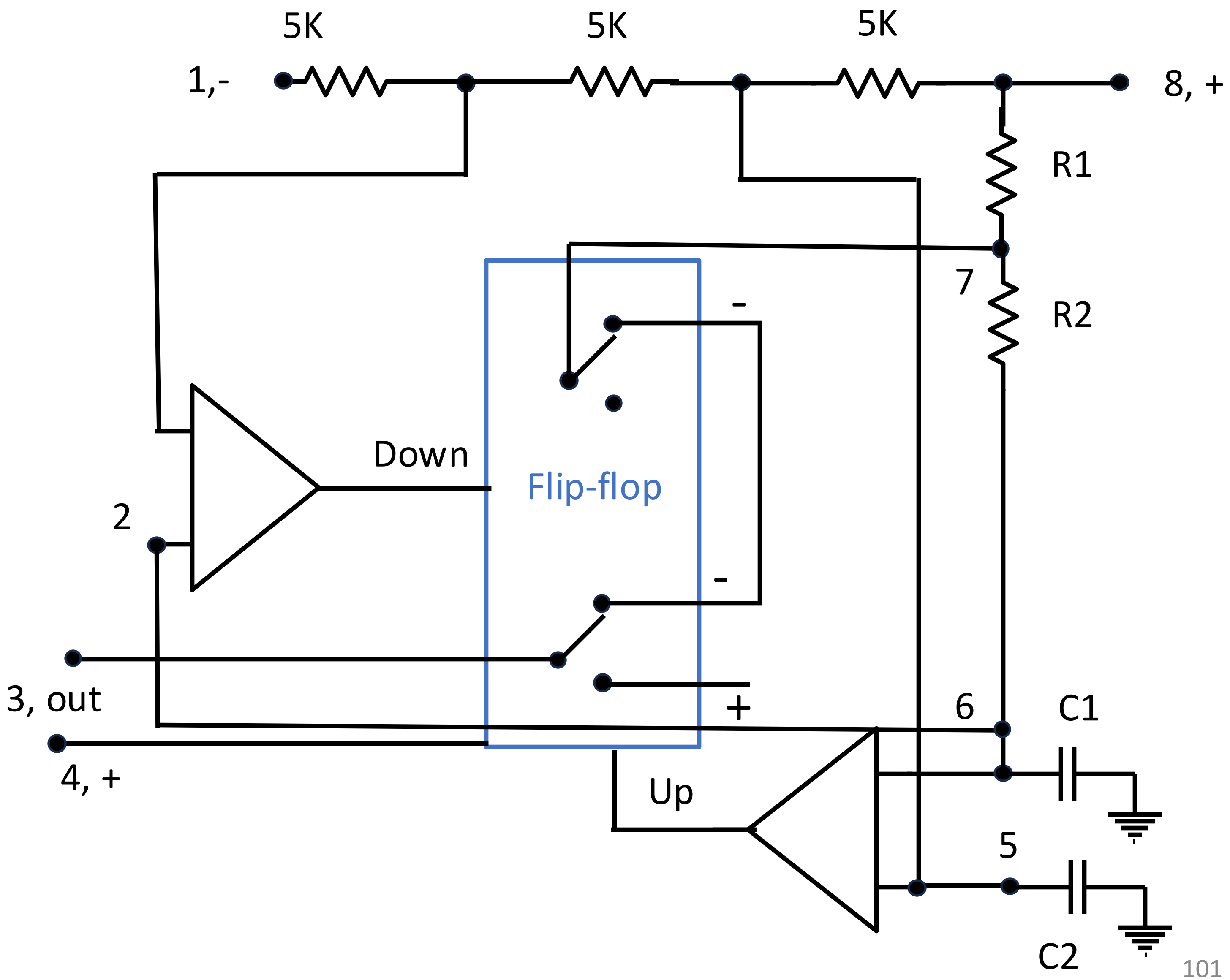


For comparator,

- A=1, if $a \geq b$
- A=0, if $a < b$

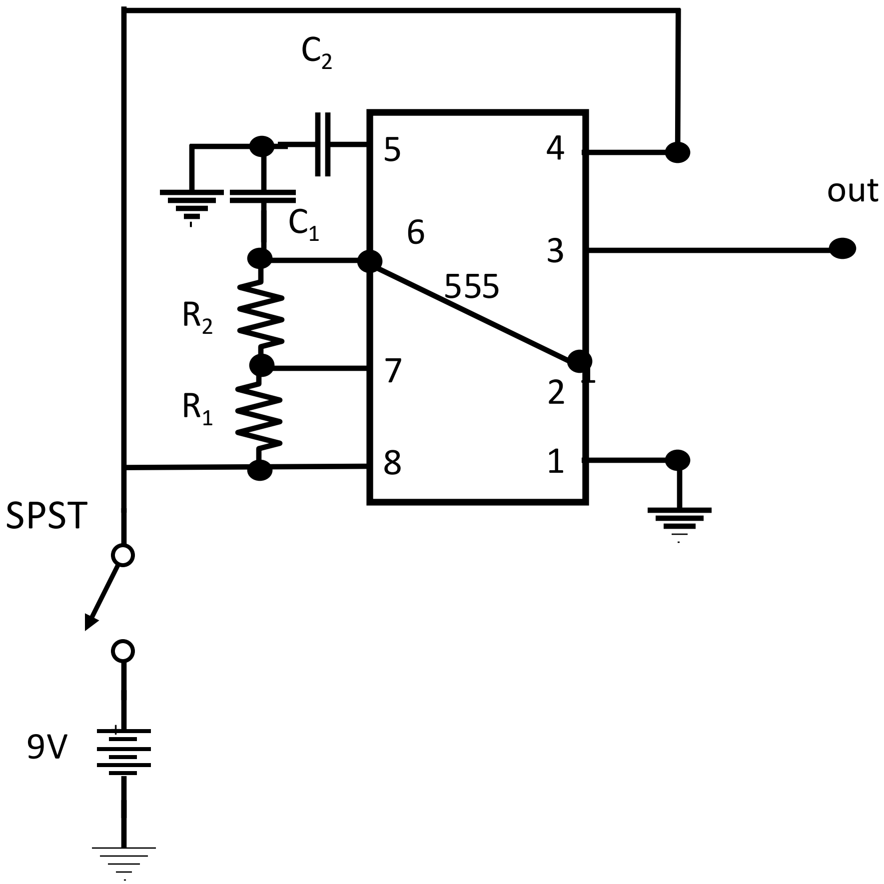


555 Astable



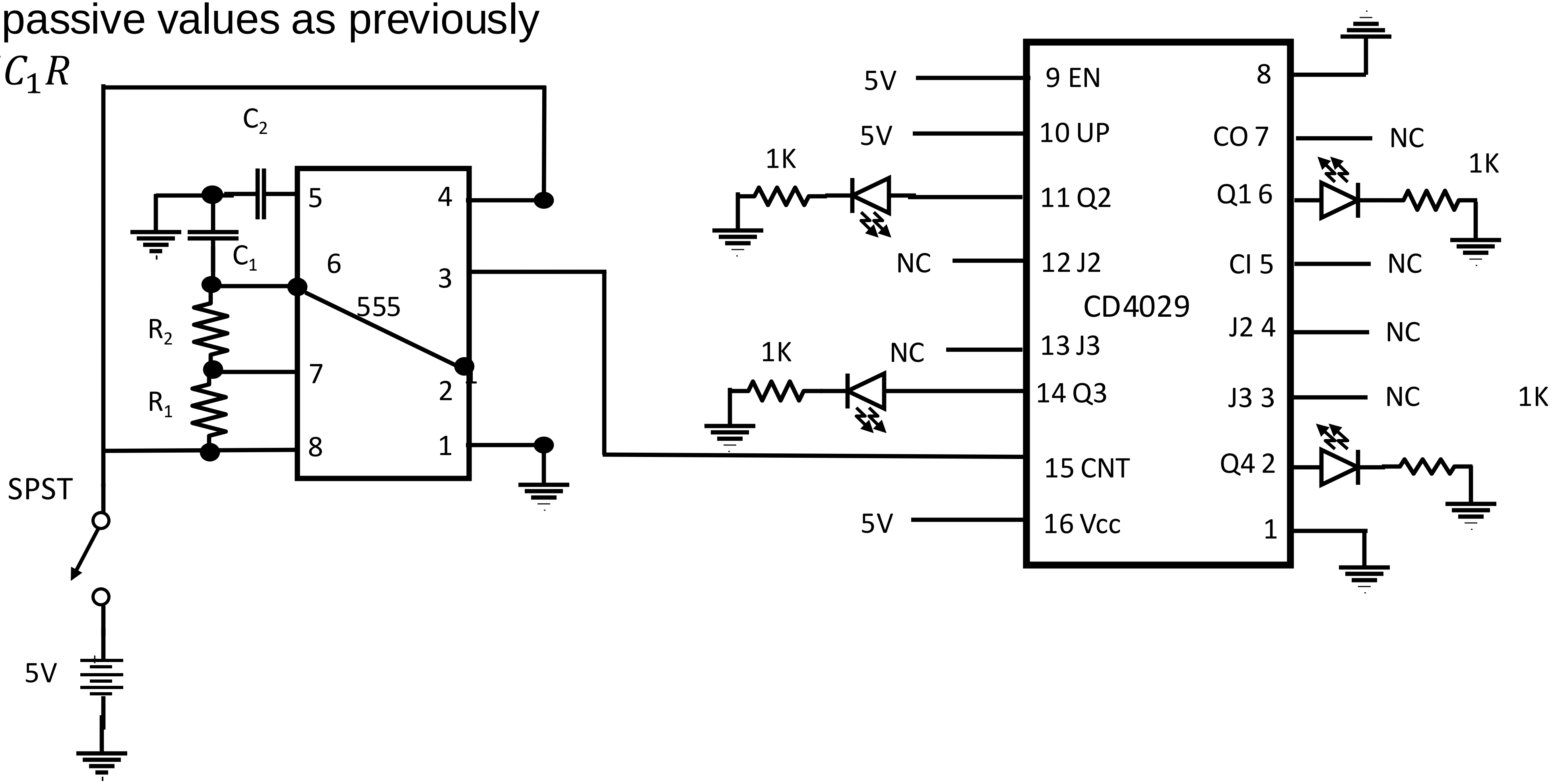
Using the 555 to generate a clock

- We use the 555 in *astable* mode
- $t_0 = .7C(R_1 + 2R_2)$
- $duty - \text{cycle} = \frac{R_1+R_2}{R_1+2R_2}$
- Try
  - $R_1 = R_2 = 10K$
  - $C_1 = 10\mu F$
  - $C_2 = .1\mu F$
  - Freq: ~50/sec
- Try
  - $C_1 = 1\mu F,$ what happens to the frequency



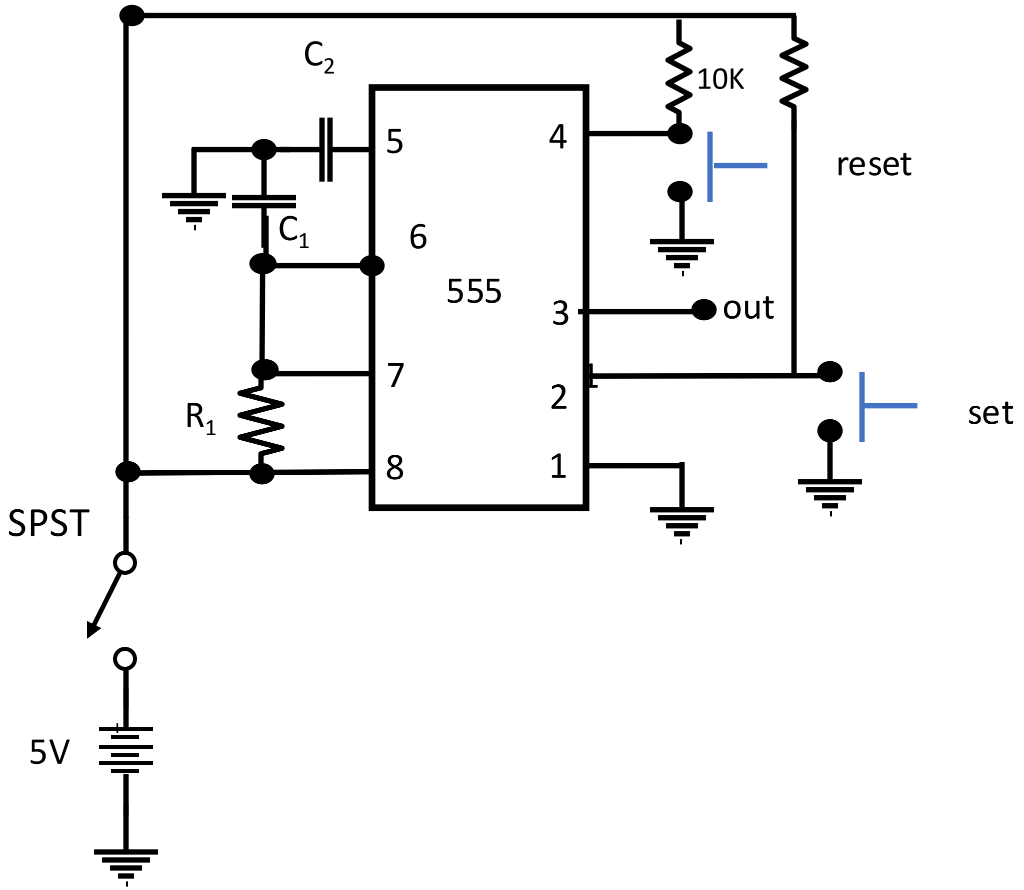
Using the 555 to clock a counter

- 555 in *astable* mode used as counter
- Same passive values as previously
- $t_0 = .7C_1R$



Using the 555 to generate a pulse

- Try
  - $R_1 = R_2 = 10K$
  - $C_1 = 100\mu F$
  - $C_2 = .1\mu F$
- If you build this and connect an oscilloscope, you'll notice the wave isn't perfectly square. As the pulse rises, you get some overshoot and on discharge there is undershoot. What happens if connect a RC circuit at the output with C around .1 μF. You can also use a Schmidt trigger to clean up messy waveforms.
- Pulse length time is $R_1 C_1$ secs



More circuits

- Differential amp

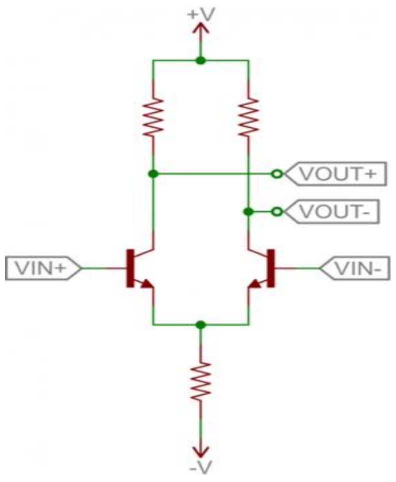
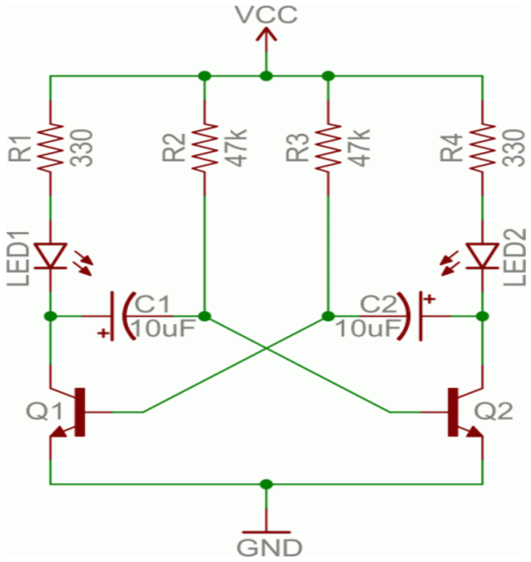


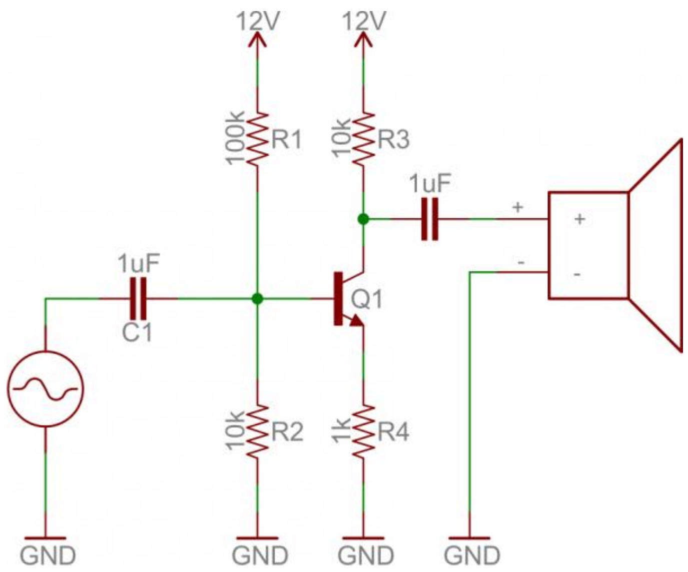
Diagram credit: Sparkfun

Multi-vibrator



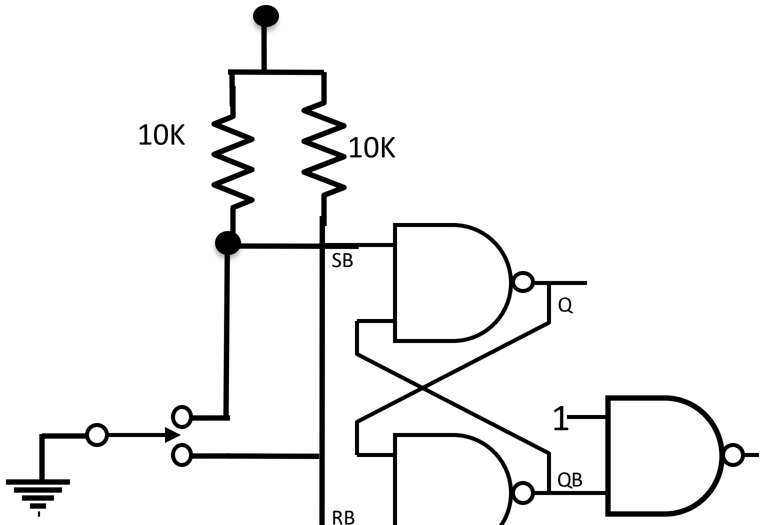
Build this and try 100 μF capacitors.
How often do the LED's blink?

Common emitter as audio amp

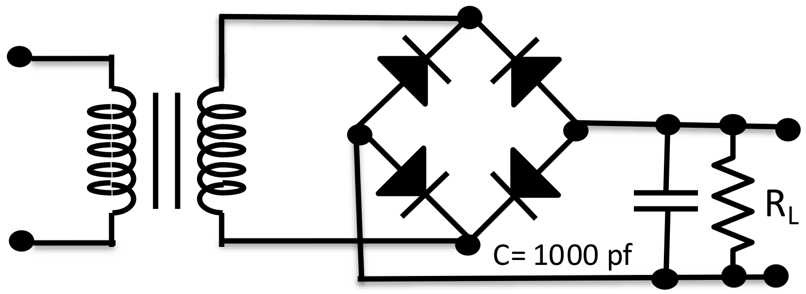


More Circuits

Debounce



Power supply



Voltage regulator