Electronics of Radio

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/s
- Kirhhoff 1: $\sum_{loop} V_i = 0$ (Conservation of energy)
- Kirhhoff node: $\sum_{node} I_i = 0$ (Conservation of charge)
- V$(t) = V_p \cos(\omega t)\,, \omega = 2\pi f\,,$ 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: $P_a = \int_0^{1/f} V(t)I(t)dt = V(t)I(t) = \int_0^{2\pi/\omega} V_p I_p\, cos^2(\omega t)dt = \frac{V_p I_p}{2}$
- 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

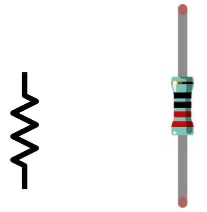
Signals

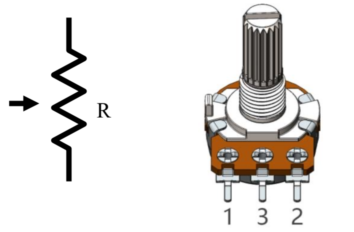
- Gain (G) expressed in decibels: $G = 10\, log_{10}(^{P_{out}}/_{P_{in}})$
- Mixer:
  - $V(t) = \cos(\omega_1 t)\cos(\omega_2 t) = \frac{1}{2}[\cos(\omega_+ t) + \cos(\omega_- t)]$, $\omega_+ = \omega_1 + \omega_2$, $\omega_- = \omega_1 - \omega_2$
- Modulation

Name	Equation
AM	$V(t) = a(t)\cos(\omega_c t)$
FM	$V(t) = V_c\cos((\omega_c + a(t))t)$
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

Resistors, capacitors, inductors

- Resistors
  - Analytic model: $IR = V$
  - Energy dissipated: $E = \int_{t_i}^{t_f} IV \,\mathrm{dt} = \int_{t_i}^{t_f} I^2 R dt$





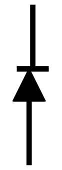
- Capacitors
  - Analytic model: $CV = q, C\frac{dV}{dt} = i$
  - Capacitor Energy stored: $E = \int_{t_i}^{t_f} CV\frac{dV}{dt}\,\mathrm{dt} = \frac{1}{2}CV^2$

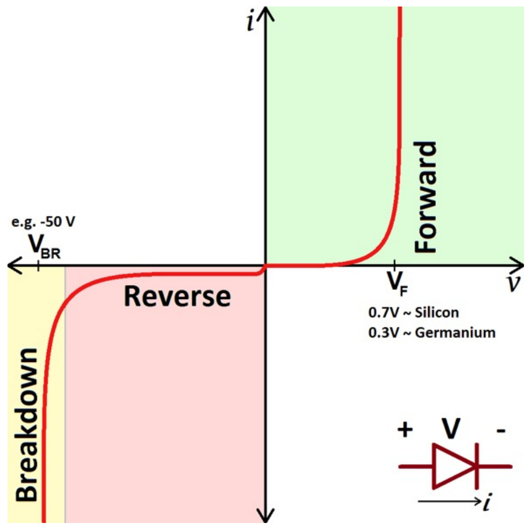
- Inductors
  - Analytic model: $V = L\frac{di}{dt}$
  - Inductor Energy stored: $E = \int_{t_i}^{t_f} IV\,\mathrm{dt} = \int_{t_i}^{t_f} LI\frac{dI}{dt} dt = \frac{1}{2}LI^2$

Credit: Make Electronics

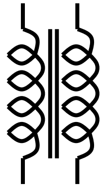
Diodes, transformers

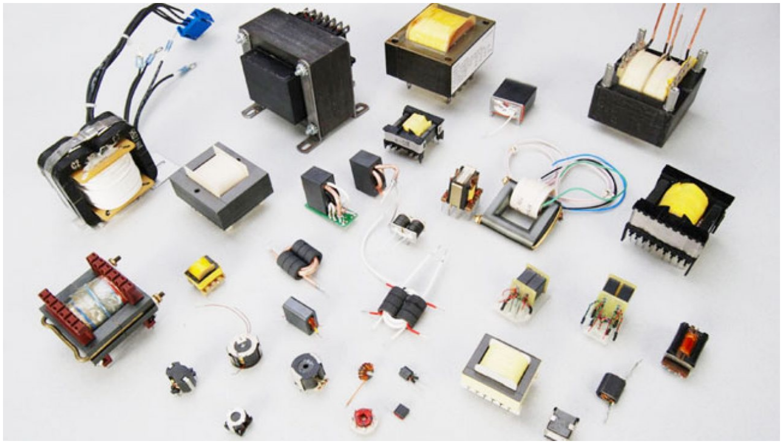
- Diodes
  - Devices that allow current to flow only in one direction
  - Silicon diodes, for example have, essentially infinite resistance if $V_{ac}<0$, that is if the cathode is at a higher potential than the anode and very low resistance if $V_{ac}> .7V$.
  - The cathode is usually labelled with a band
- Transformers
  - AC only: $\frac{N_2}{N_1} = \frac{V_2}{V_1}$



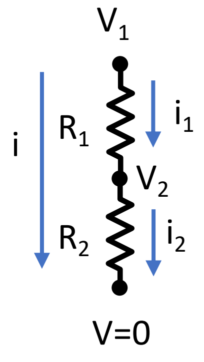


Credit: Make Electronics

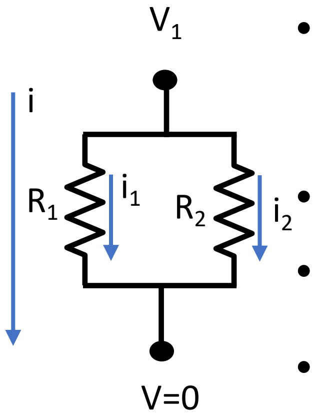




Simple circuit analysis with Kirchhoff

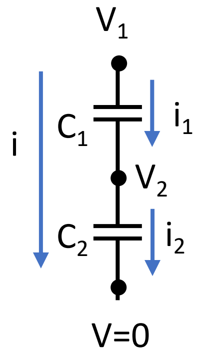


- R_{eq} is the equivalent resistance, replacing the top left circuit with a single resistance.
- By Kirchhoff's node rule, $i_1 = i_2 = i$, so
- $\frac{V_1-V_2}{R_1} = \frac{V_2}{R_2} = \frac{V_1}{R_{eq}}$ thus $\frac{R_1}{R_{eq}} V_1 = V_1 - V_2$ and $\frac{R_2}{R_{eq}} V_1 = V_2$. Adding, we get $\frac{R_1}{R_{eq}}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

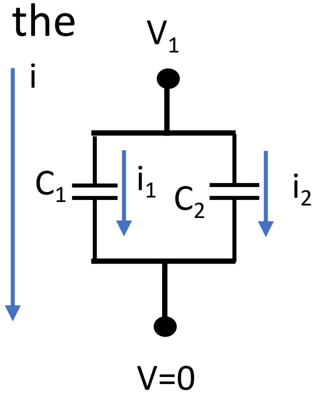


- Again let R_{eq} is the equivalent resistance, replacing the bottom left circuit with a single resistance.
- By Kirchhoff's node rule, $i_1 + i_2 = i$, so
- $\frac{V_1}{R_1} + \frac{V_1}{R_2} = \frac{V_1}{R_{eq}}$.
- Solving, we get. $\frac{R_1 R_2}{R_1+R_2} = R_{eq}$

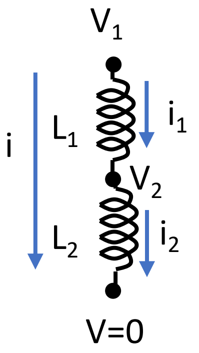
- C_{eq} is the equivalent capacitance, replacing the top right circuit with a single capacitor.
- By Kirchhoff's node rule, $i_1 = i_2 = i$, so
- $C_1 \frac{d\,(V_1-V_2)}{dt} = C_2 \frac{d\,(V_2)}{dt} = C_{eq} \frac{dV_1}{dt}$
- $\frac{C_{eq}}{C_1} \frac{d\,(V_1)}{dt} = \frac{d\,(V_1-V_2)}{dt}$ and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



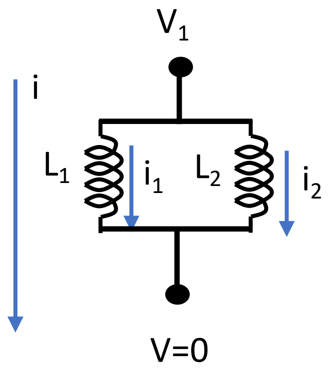
- C_{eq} is the equivalent capacitance, replacing the bottom right circuit with a single capacitor.
- By Kirchhoff's node rule, $i_1 + i_2 = i$
- $C_{eq} \frac{dV_1}{dt} = C_1 \frac{dV_1}{dt} + C_2 \frac{dV_1}{dt}$, so
- $C_{eq} = C_1 + C_2$



Simple circuit analysis with Kirchhoff

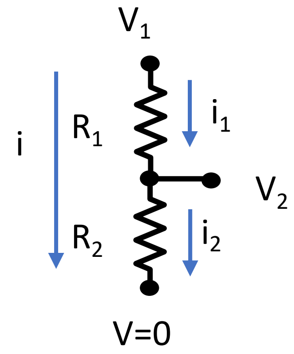


- Let L_{eq} be the equivalent inductance, replacing the top left circuit with a single inductor.
- By Kirchhoff's node rule, $i_1 = i_2 = i, \text{so}$
- $L_{eq}\frac{di}{dt} = V_1, L_1\frac{di_1}{dt} = V_1 - V_2, L_1\frac{di_2}{dt} = V_2$
- $V_1 = L_{eq}\frac{di}{dt} = L_1\frac{di}{dt} + L_2\frac{di}{dt}$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

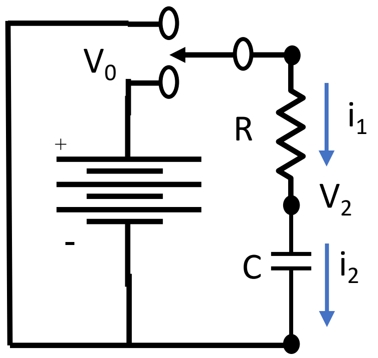


- Let L_{eq} be the equivalent inductance, replacing the bottom left circuit with a $\frac{di}{dt} = \frac{V_1}{L_{eq}}, \frac{di_1}{dt} = \frac{V_1}{L_1}, \frac{di_2}{dt} = \frac{V_1}{L_2},$
- single inductor.
- By Kirchhoff's node rule, $i_1 + i_2 = i, \text{so}$
- $\frac{V_1}{L_{eq}} = \frac{V_1}{L_1}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

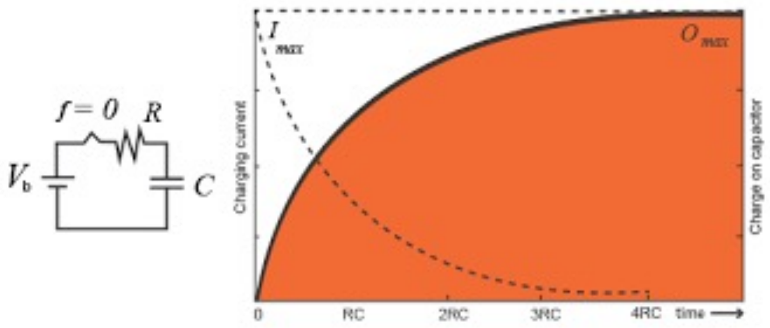
- The circuit on the right, is useful and is called a *voltage divider*.
- $i = i_1 = i_2$ so $\frac{V_1 - V_2}{R_1} = \frac{V_2}{R_2}, V_1 - V_2 = \frac{R_1}{R_2}V_2$
- Thus, $V_1 = (1 + \frac{R_1}{R_2})V_2$ and so
- $V_2 = \frac{R_2}{R_1 + R_2}V_1$

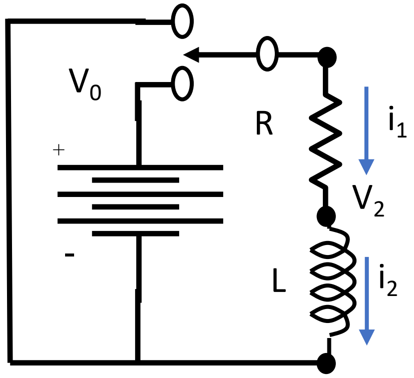


RC/RL circuit analysis with Kirchhoff

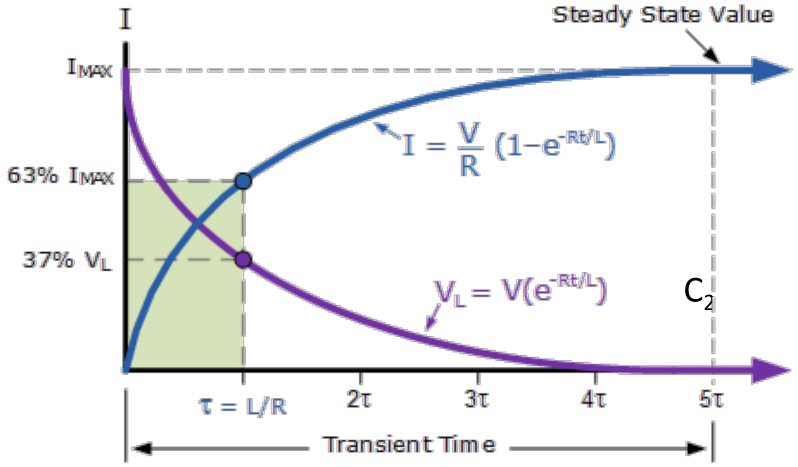


- RC behavior: charging
  - $V_0 - V_2 = i_1 R = V_R, i_1 = \frac{V_R}{R}$
  - $i_2 = C\frac{dV_2}{dt}, V_C = V_2$
  - $i_1 = i_2, V_C = V_0 - V_R$
  - $\frac{V_R}{R} = C\frac{dV_C}{dt}$, RC$\frac{dV_C}{dt} = V_0 - V_C$, or RC$\frac{dV_C}{dt} + V_C = V_0$
  - Solution is $V_C = V_0 - V_0 e^{-\frac{t}{RC}}$





- 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 $V_L = V_0 e^{-\frac{Rt}{L}}$



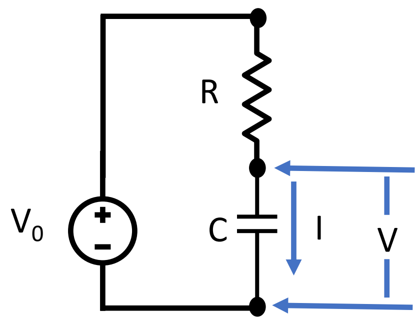
Phasors

- $V(t) = RI(t)$
- $V(t) = L\dot{I}(t)$
- $I(t) = C\dot{V}(t)$
- 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.
- $V(t) = Re(e^{j(\omega t+\theta)})$, and $\mathrm{I}(t) = Re(e^{j(\omega t+\phi)})$
- Now define $V = Ae^{j\theta}$ and $I = Be^{j\phi}$, so $|V| = A$, $|I| = B$, $\angle V = \theta$, and $\angle I = \phi$. V and I are called phasors and do not include time. Note that $V(t) = Re(Ve^{j\omega t})$ and $\mathrm{I}(t) = Re\left(Ie^{j\omega t}\right)$.
- Note that $I = CVj\omega$, for a capacitor and $V = LIj\omega$, for an inductor

Circuit analysis and impedance

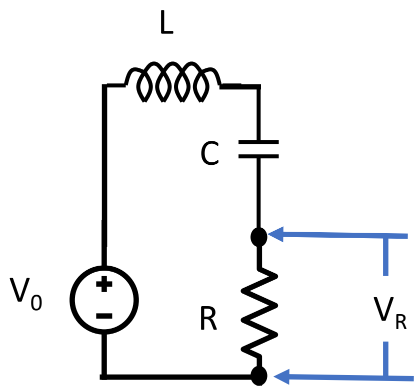
- Impedance unifies the “simple” ohms law with capacitance and inductance.
- $Z = R$, for resistors, $Z = j\omega L$, for inductors and $Z = \frac{1}{j\omega C}$, for capacitors.
- In general, $Z = R + jX$ and all the ohm like laws hold for resistors, capacitors and inductors .
  - $Z_{eq} = Z_1 + Z_2$ for two components with impedance Z_1, Z_2 connected in series
  - $Z_{eq} = \frac{Z_1 2}{Z_1 + Z_2}$ 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



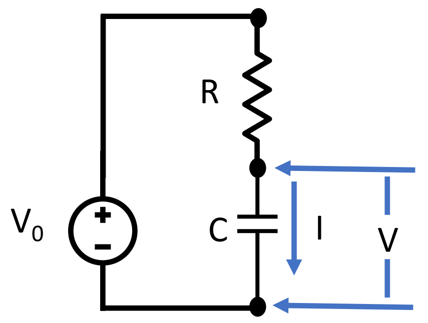
- For the circuit on the right, $Z = R + \frac{1}{j\omega C}$ is the impedance for the resistor and capacitor in series.
- The phasor $I = \frac{V_0}{Z}$ and the phasor $V = \frac{I}{j\omega C} = \frac{V_0}{1+j\omega RC}$
- Further, $|I| = \frac{V_0}{|Z|}$, $\angle I = \angle \frac{V_0}{|Z|}$ and $|V| = \frac{|I|}{|j\omega C|} = |\frac{V_0}{1+j\omega RC}|$
- 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.

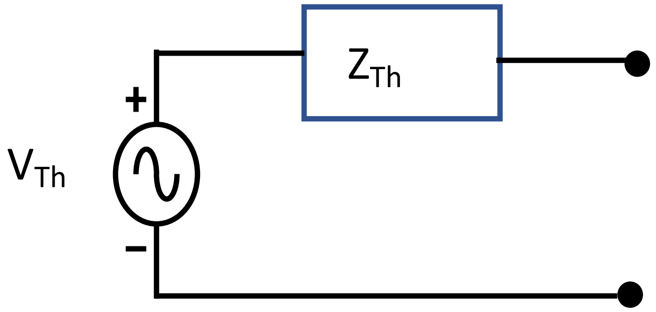
Phasors, impedance and power



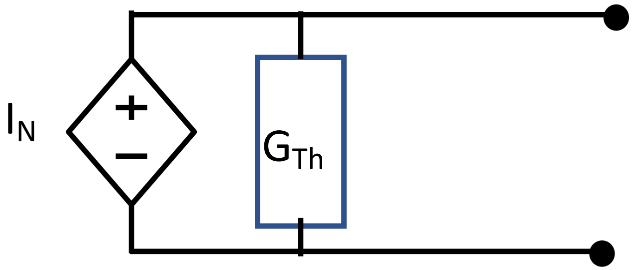
- For the circuit on the right, $Z = R + \frac{1}{j\omega C}$ is the impedance for the resistor and capacitor in series.
- The phasor $I = \frac{V_0}{Z}$ and the phasor $V = \frac{I}{j\omega C} = \frac{V_0}{1+j\omega RC}$
- Further, $|I| = \frac{V_0}{|Z|}$, $\angle I = \angle \frac{V_0}{|Z|}$ and $|V|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

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



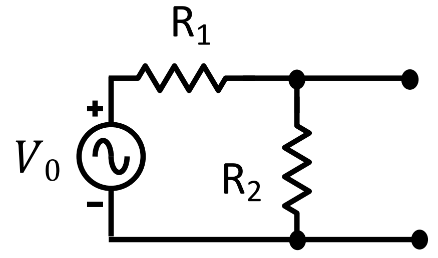
- Norton: Any combination of linear sources and passive elements terminating in two terminals is equivalent to a pure current source in parallel with a conductance



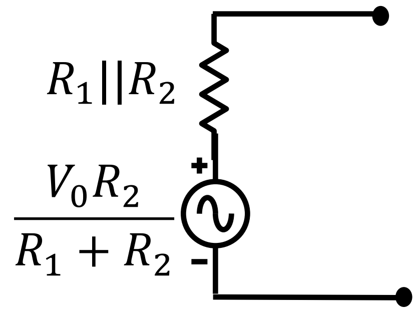
- Similar theorems for two terminal input and output devices (with transfer function)

Thevenin and Norton

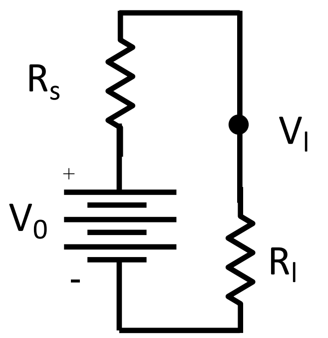
- We can use lookback resistance to calculate the Thevenin equivalent resistance and ideal source.
- To find the lookback resistance, short the source and apply the usual laws.
  - Here $R_s = R_1||R_2$
- To find the new ideal source, notice R_1 and R_2 form a voltage divider.
  - The new source voltage is $\frac{V_0 R_2}{R_1+R_2}$



Is equivalent to

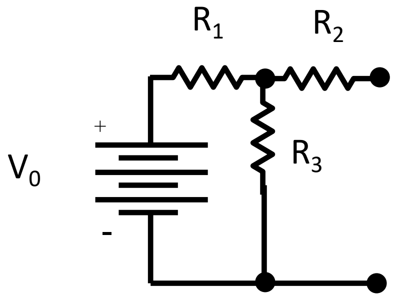


Exercise 1: Resistors



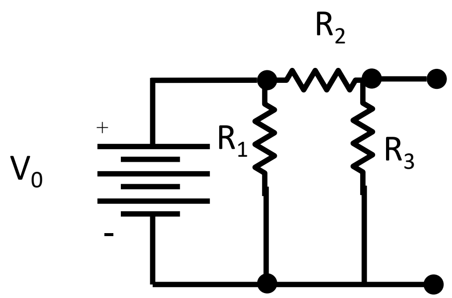
A

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 $R_l = R_s$



B

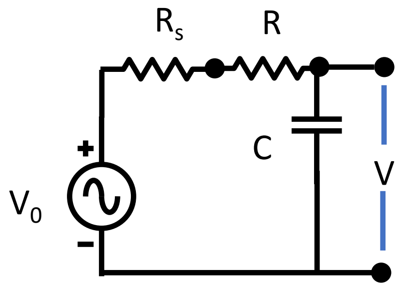
2. Find the Thevenin and Norton parameters for (B).
   - $V_{Th} = \frac{R_3}{R_1+R_3}V_0$
   - $R_{Th} = R_2 + R_1||R_3$
3. Find the Thevenin and Norton parameters for (C).
   - $V_{Th} = \frac{R_3}{R_2+R_3}V_0$
   - $R_{Th} = R_2||R_3$



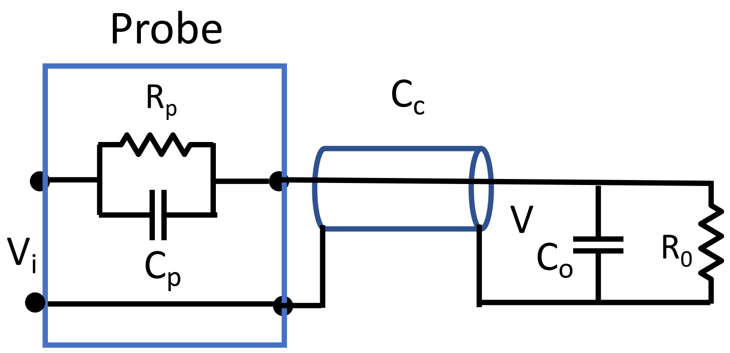
C

Exercise 3: Capacitors

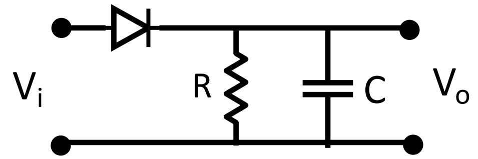
1. In the circuit on the right, V_0 is a 2 volt pp ideal square wave source of frequency 20Hz, $R_S = 50\Omega$, $R = 300k\Omega$ and $C = 10\ nF$. Period is $50\ millisec$



2. What is the voltage, V, at the output? The scope has an input resistance of $1M\Omega$.
   - About a volt at peak
3. Let t_2, the time to discharge to 0V. Calculate τ and t_2.
   - $\tau = 3{\times}10^5{\times}10^{-8}\ sec = 3\ millisec$
   - $t_{12} \approx 1.5ms$
4. Capacitance on the scope prevents the delay from being 0. Measure the new t_2 with these changes.
5. Given C_0 and C_p and $R_{p.}$
   - $C_0 = 100pf/m, C_o = 50pF, C_p = 10pF$
6. Now calculate the new t_{12}.
   - $\tau = 6\mu$-sec



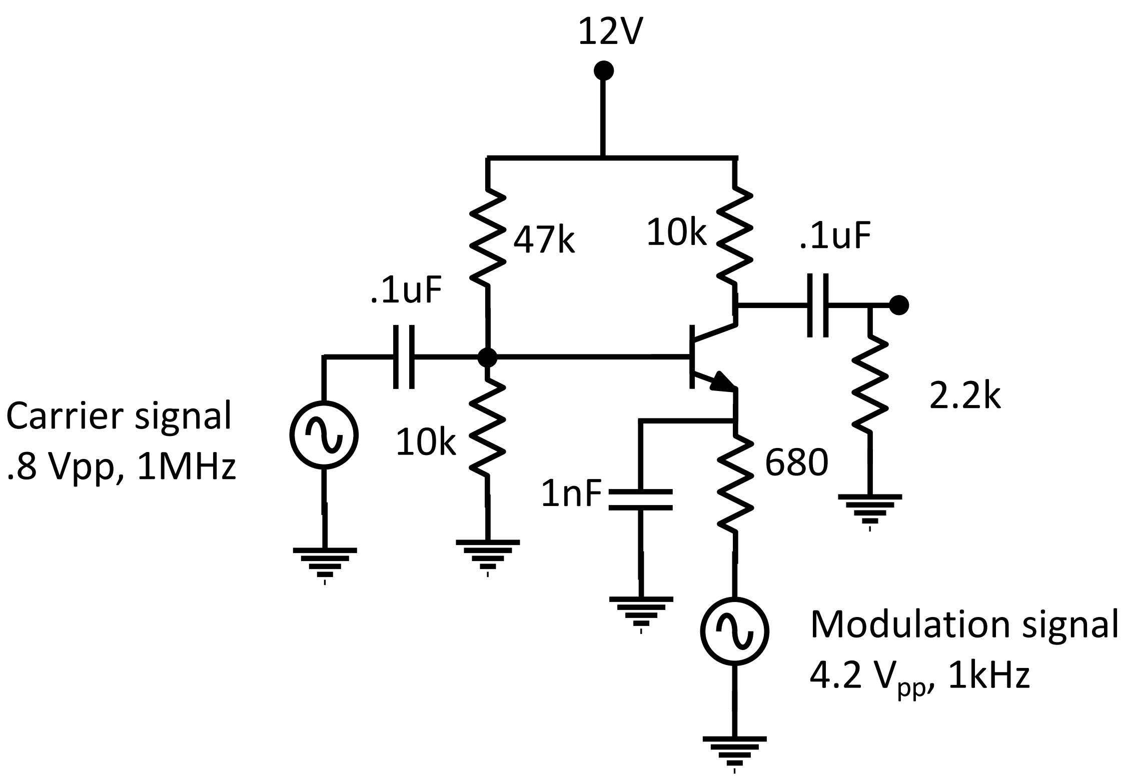
Exercise 4: Diode detectors



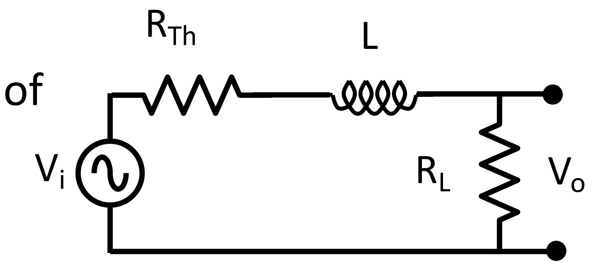
- For AM, $V(t) = V_c \cos(\omega_c t) + a(t) \cos(\omega_c t)$, Define the modulation depth $m = \frac{a_p}{V_c}$
- In circuit on the right, $R = 10k\Omega$, $C = 10\ nF$
- Set function generator for $f_c = 1MHz$, $V_{c,pp} = 5V$, $f_m = 1kHz$, $m = .7$
  1. Calculate τ for the RC circuit. $\tau = 10^4 \times 10^{-8}$ 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

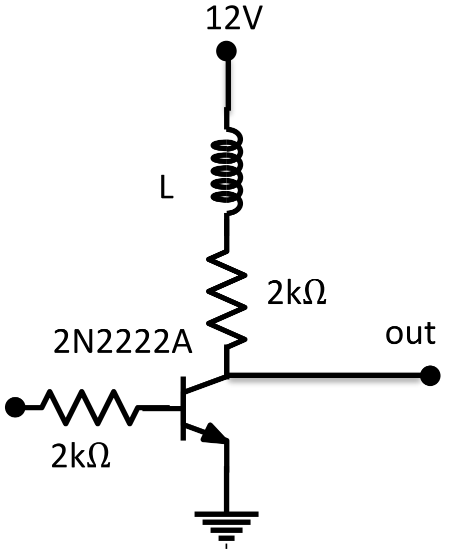
- I didn't have a signal generator that produced an AM signal, so I used the modulator on the right with the indicated inputs to produce the AM needed for the detector in the previous exercise.



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
     - $Z_{eq} = 150 + jL\omega$, $\omega = 2\pi\times10^3$, $V_i = Re(V_{i,p}e^{j\omega t})$
     - 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.
     - $1\mu\,\text{sec}\, rise\ time$. Ringing at 10MHz. $\frac{1}{\sqrt{LC}} = 62.8\times10^6$. $C = \frac{10^3}{(62.8\times10^6)^2} \approx\ .25pF$



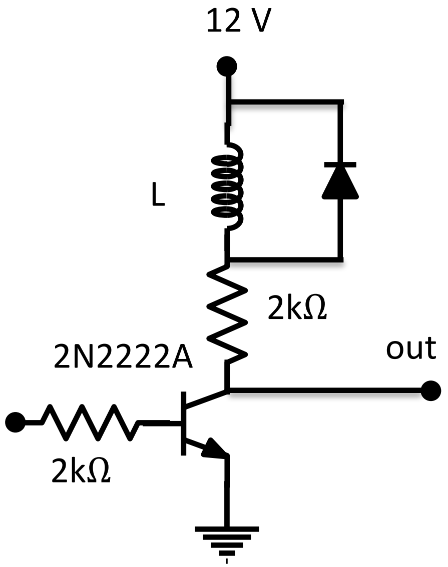
- Note: I made the pull-up 100K.

Exercise 6: Diodes and snubbers

- Add indicated snubber diode.

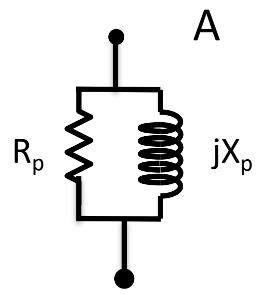
1. Swing up is nearly immediate with snubber
2. Ringing at 10MHz. $\frac{1}{\sqrt{LC}} = 62.8\times10^{6}$. $C = \frac{10^3}{(62.8\times10^6)^2} \approx .25pF$
3. What is its effect on ringing?
   - Ringing is uniform at 5 MHz
4. Diode should be on when transistor is off.

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

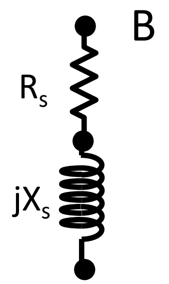


Exercise 7: Parallel to Series conversion

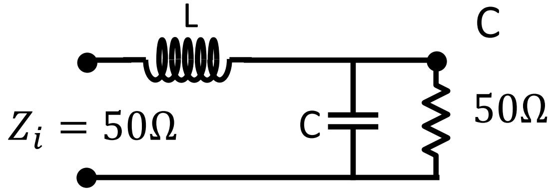
- For series: $Z_s = R_s + j\omega L$, $Q_s = \frac{X_s}{R_s}$
- For parallel: $\frac{1}{Z_p} = \frac{1}{R_p} + \frac{1}{j\omega L}$, so $Z_p = \frac{j\omega L R_p}{R_p + j\omega L}$ and $Q_p = \frac{R_p}{X_p}$
- If $Q_p = Q_S$, $X_p X_s = R_p R_S$



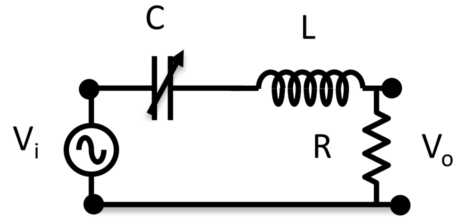
1. If circuits (A) and (B) have the same impedance, what is the relationship between R_p, X_p and R_s, X_s?
   - $\frac{1}{Z_p} = \frac{1}{R_p} + \frac{1}{jX_s}$, $Z_p = \frac{jR_p X_p}{R_p + jX_p} = \frac{{X_p}^2 R_p + j{R_p}^2 X_p}{{R_p}^2 + {X_p}^2}$, $Z_s = R_s + jX_s$
   - $R_s = \frac{{X_p}^2 R_p}{{R_p}^2 + {X_p}^2}$, $X_s = \frac{{R_p}^2 X_p}{{R_p}^2 + {X_p}^2}$, $R_s = X_p \frac{X_p R_p}{{R_p}^2 + {X_p}^2}$, $X_s = R_p \frac{X_p R_p}{{R_p}^2 + {X_p}^2}$, 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

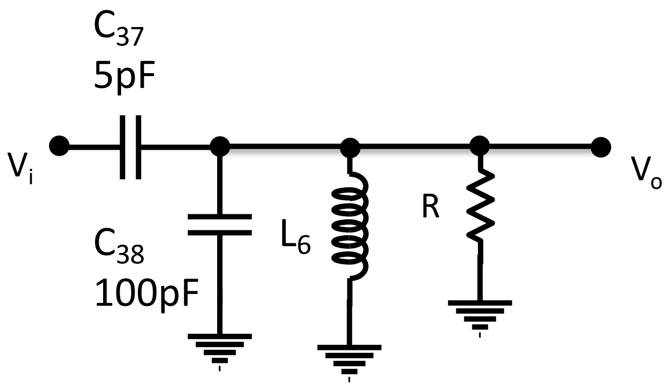


- For the circuit on the right, C$= 8 - 50pf$, $L = 15\mu H$ forming a bandpass filter. $R = 100\Omega$
- If $C = 34pf$, the resonant frequency is $\omega = \frac{1}{\sqrt{35\times10^{-12}\times15\times10^{-6}}} = \frac{10^9}{\sqrt{525}} \approx 44.2$, so the resonant frequency is $\frac{44.2}{2\pi}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

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, $L_6 = 3.1\mu H$

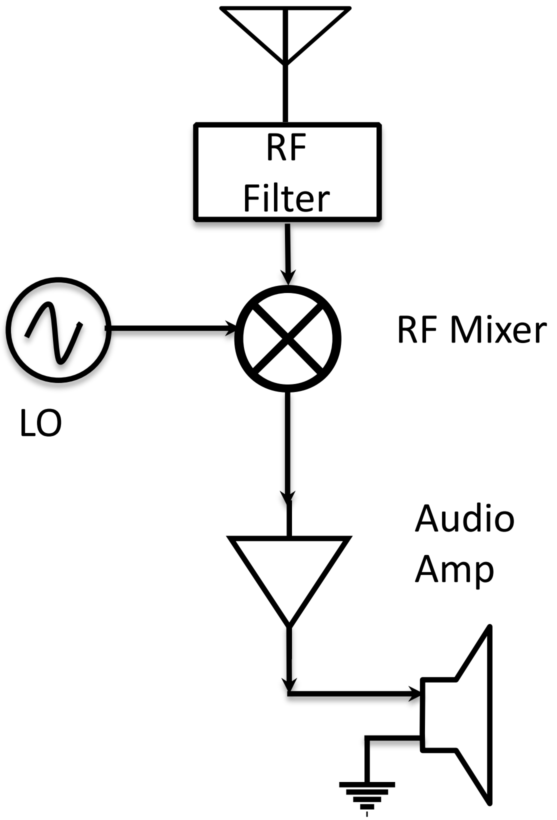


1. Again, find the resonant frequency, the frequencies corresponding to a 3db falloff, the bandwidth and the Q of this circuit. This circuit is in the transmit oscillator
   - For $C_{38}, L_6, R = 100\Omega$, network: $Q = 100\sqrt{\frac{10^{-10}}{3.1\times10^{-6}}} = 5.6.$
   - $BW = \frac{f_r}{Q} = \frac{7MHZ}{5.6} = 1.25MHz.\ f_u = f_r + \frac{BW}{2} = 7.625MHz,\ f_l = f_r - \frac{BW}{2} = 6.375MHz.$

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

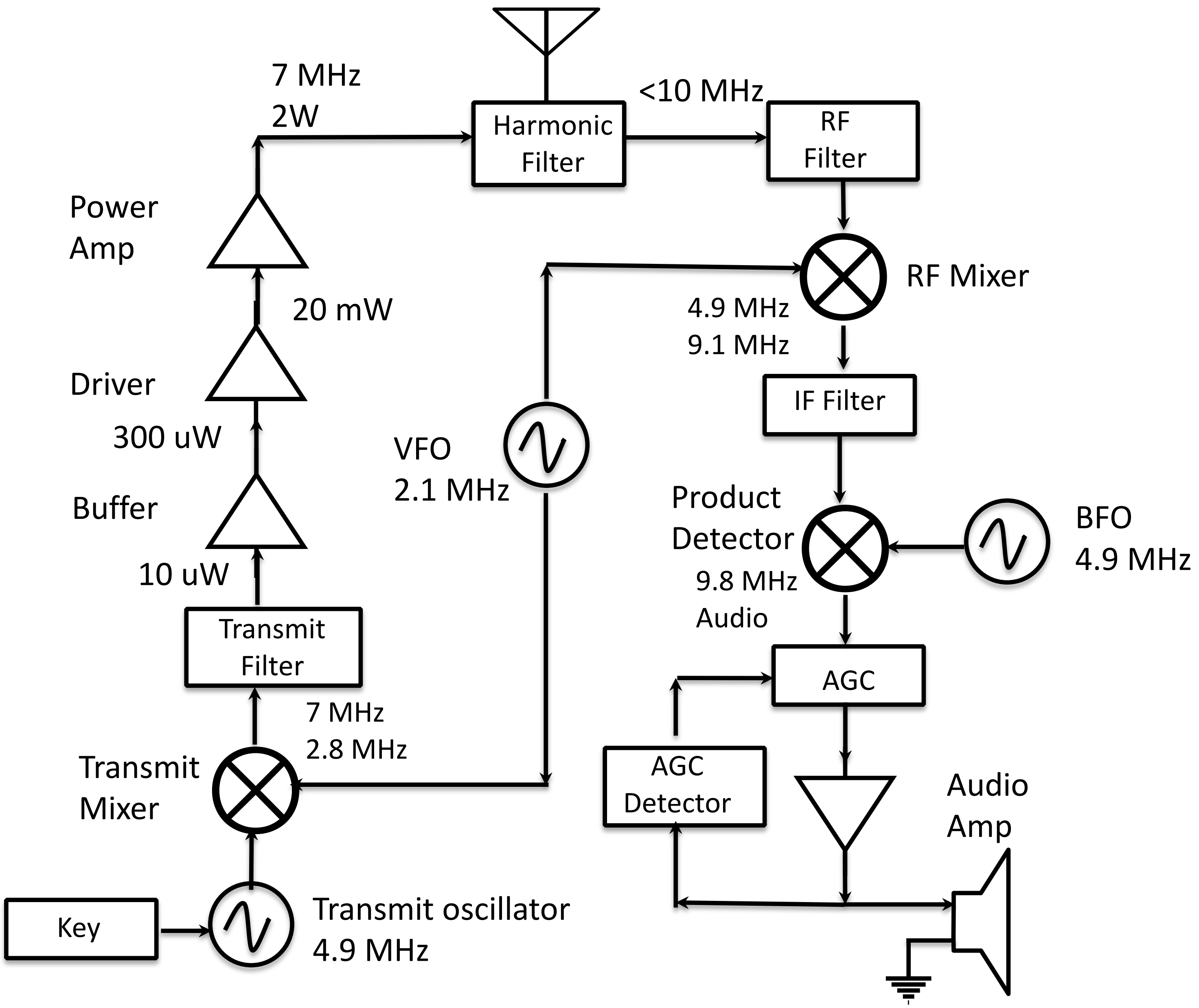
Direct conversion and superhet receivers

- Image frequency
  - $\omega_{rf} = \omega_{LO} - \omega_a$
  - $\omega_i = \omega_{LO} + \omega_a$
- Superheterodyne designs
  - $\omega_{rf} = \omega_{IF} + \omega_{VFO}$
  - $\omega_{vi} = \omega_{IF} - \omega_{VFO}$
  - $\omega_{IF} = \omega_{BFO} - \omega_a$
  - $\omega_{bi} = \omega_{BFO} + \omega_a$
  - $\omega_{usb} = \omega_{VFO} + \omega_{BFO} + \omega_a$
  - $\omega_{lsb} = \omega_{VFO} + \omega_{BFO} - \omega_a$



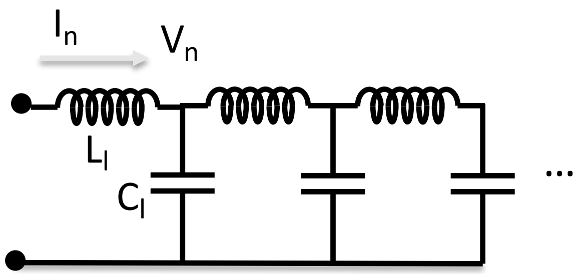
Direct conversion

Norcal 40A

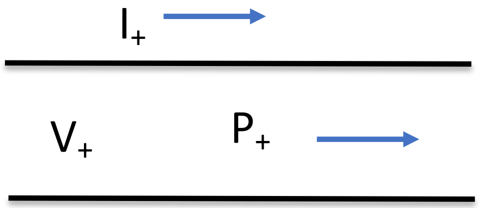


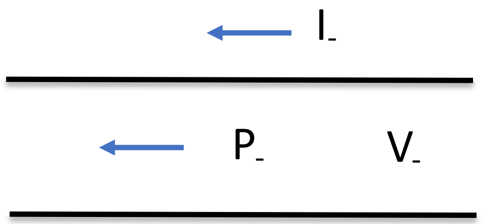
7 MHz
2W
Harmonic Filter
<10 MHz
RF Filter
Power Amp
RF Mixer
20 mW
4.9 MHz
9.1 MHz
Driver
IF Filter
300 uW
VFO
2.1 MHz
Buffer
Product Detector
BFO
4.9 MHz
10 uW
9.8 MHz
Audio
Transmit Filter
AGC
7 MHz
2.8 MHz
AGC Detector
Transmit Mixer
Audio Amp
Key
Transmit oscillator
4.9 MHz

Transmission Lines



- $V_{n+1} - V_n = -L_l \frac{\partial I_{n+1}}{\partial t}, \; L = \frac{L_l}{l}$
- $I_{n+1} - I_n = -C_l \frac{\partial V_n}{\partial t}, \; C = \frac{C_l}{l}$
- $\frac{\partial^2 V}{\partial z^2} = LC \frac{\partial^2 V}{\partial t^2}$ and $\frac{\partial^2 I}{\partial z^2} = LC \frac{\partial^2 I}{\partial t^2}$
- Solution is $V(z - vt), v = \frac{1}{\sqrt{LC}}$, which is the velocity 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
- $V' = vLI', \; \frac{V}{I} = \sqrt{\frac{L}{C}}, \; Z_0 = \sqrt{\frac{L}{C}}, Z_0$ is the forward or “characteristic” impedance
- Another solution is $V(z + vt), v = \frac{1}{\sqrt{LC}}$ which is the velocity for the reverse wave
- $Z_0 = \frac{V_+}{I_+}, -Z_0 = \frac{V_-}{I_-}, V = V_+ + V_-, -Z_0$ is the backwards looking impedance
- $P_+(t) = \frac{{V_+}^2}{Z_0}, \; P_-(t) = -\frac{{V_-}^2}{Z_0}$





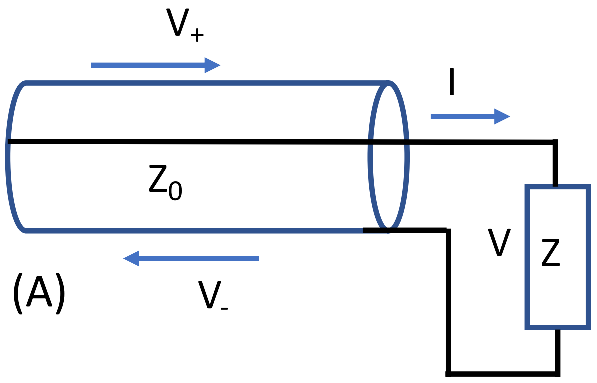
Transmission Lines - continued

- $V(z - vt) = Acos(\omega t - \beta z), v = \frac{\omega}{\beta}$; in phasor notation, $V = Ae^{-j\beta z}$.
- To compute power, let $V_+ = Ae^{-j\beta z}, V_- = Ae^{j\beta z}$
- 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 $Ae^{j(\omega t - jk)}$, and its phasor is $V = Ae^{-jkz}$. Let $Z = \frac{V}{I}$,
- $\frac{dV}{dz} = -ZI, \frac{dI}{dt} = -YV$. …(1)
- 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 (1), we get $jkV = ZI, -jkI = YV$. Solutions are $jk = \sqrt{ZY},\ Z_0 = \frac{V}{I} = \sqrt{\frac{Z}{Y}}$, all complex
- Example: 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)

Transmission Lines - dispersion

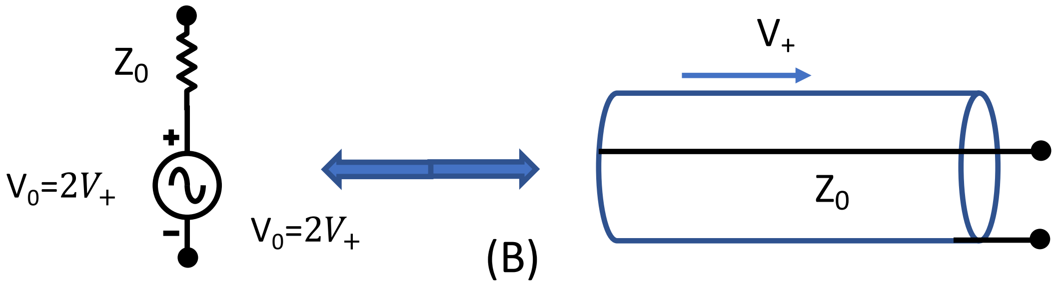
- α and v can vary with frequency; this is dispersion.
- Heaviside: Adjust parameters so $\frac{R}{L} = \frac{G}{C}$, then α and v are constants and we get:
  - $jk = j\omega\sqrt{LC}(1 + \frac{R}{j\omega L})$ and $v = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}}$, $\alpha = \sqrt{RG}$
  - We also get $Z_0 = \sqrt{\frac{L}{C}}$ 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× $10^{-3}\ {}^{nepers}/_{m}$, $\alpha l = 140dB$
  - $\alpha l \approx 140dB$ and highly dispersive

Transmission Lines-reflections



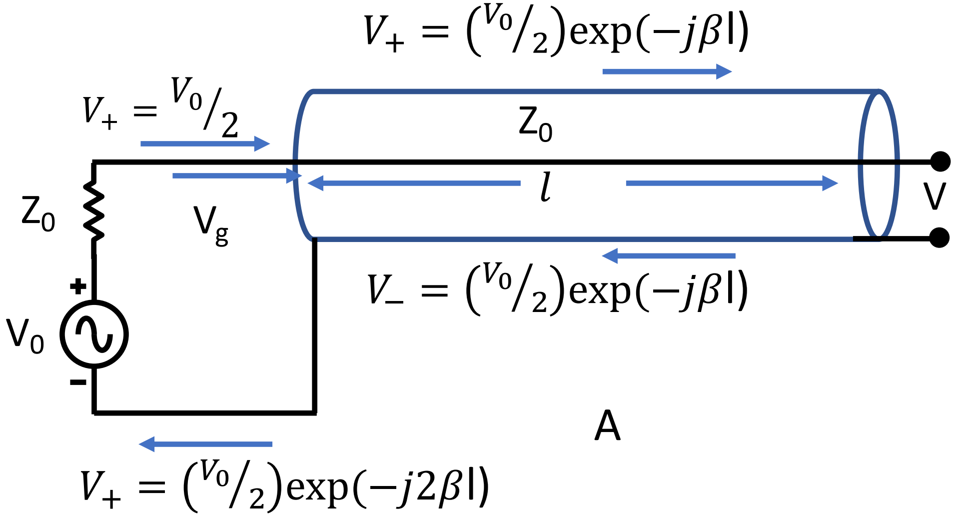
- $\rho = \frac{V_-}{V_+}$, $\rho V_+ = V_-, V = V_+ + V_- = (1+\rho)V_+$
- $\rho_i = \frac{i_-}{i_+} = -\rho$
- $\tau = \frac{V}{V_+} = 1 + \rho = \frac{2Z}{Z+Z_0}, V = 2V_+$
- Consider the circuit in the upper right (A): $V = V_+ + V_-, I = I_+ + I_-,\ Z = \frac{V}{I}$
- $Z = \frac{V}{I} = \frac{V_+ + V_-}{I_+ + I_-}$
- $\frac{Z}{Z_0} = \frac{1+\rho}{1-\rho}, \rho = \frac{Z-Z_0}{Z+Z_0},\ \rho_{open-circuit} = 1.$

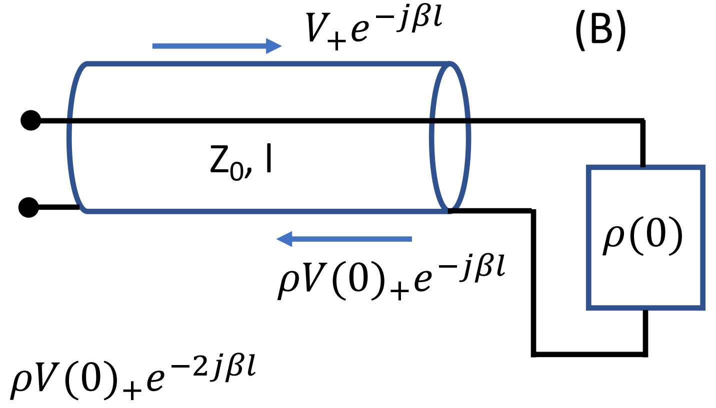
- For (B):
- Lookback resistance is $R_s = Z_0$, short circuit for (B) is $i_s = \frac{V_0}{Z_0}$
- Thevenin equivalent for open circuit is (B)
- $P_+ = \frac{{V_+}^2}{2Z_0} = \frac{{V_0}^2}{8R_s}$, This is the total available power



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: $Z(\lambda/4) = \frac{1}{z(0)}$
  - $Z_0 = \sqrt{Z(\lambda/4)Z(0)}, Z_0 = \sqrt{R_S R_L}$

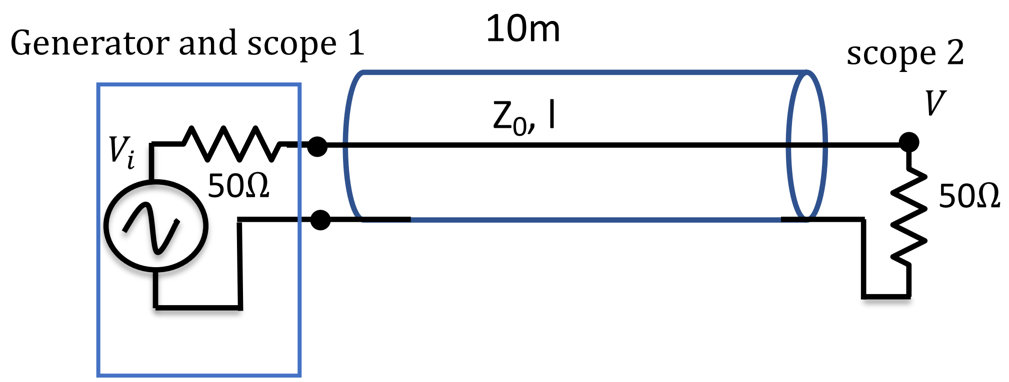


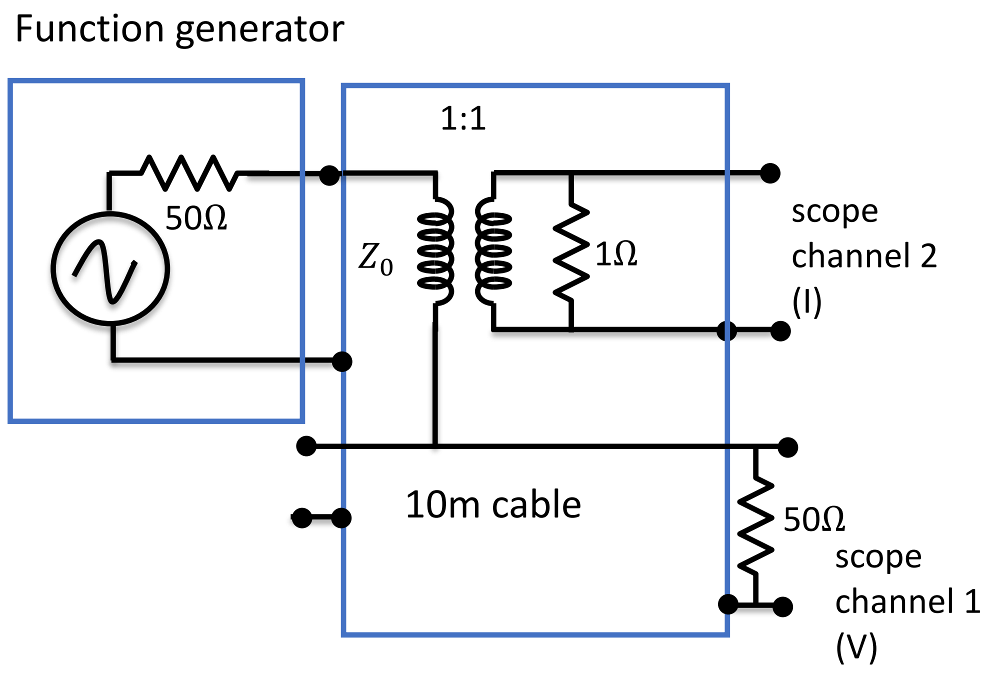


Exercise 10: Coax

- 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.
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 $v = 2\times10^8\ m/s$. $Z_0 = 50\Omega$. For high impedance, $Z_0 = \sqrt{\frac{L}{C}}$ and $v = \frac{1}{\sqrt{LC}}$. So, ${Z_0}^2 C = L$ and $v^2 = \frac{1}{LC}$, so ${Z_0}^2C^2v^2$=1. $C = \frac{1}{Z_0 v} = 10^{-10}F$. $2500\times10^{-10}F = \mathrm{L} =$ $250\mathrm{nH}$, which is what we use in the next problem.

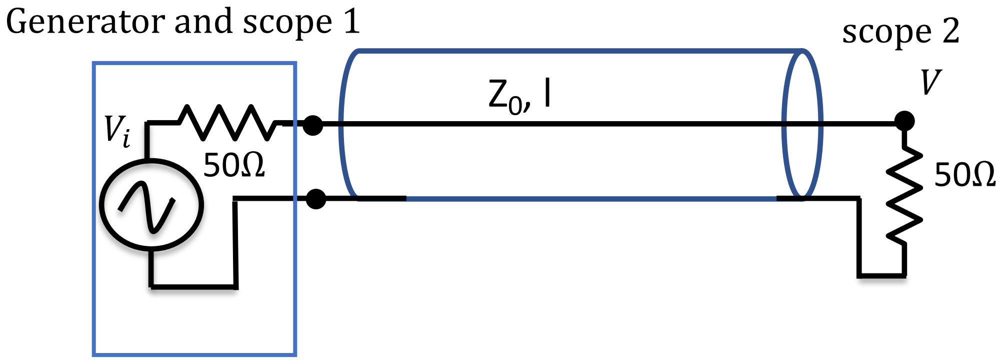




Exercise 11: Waves

- 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.
   - $Z(f) = j\omega L + R = 50\times10^{-3} + j\cdot 2\pi f\cdot 250\times10^{-9}$
   - $Y(f) = j\omega C + \frac{1}{R} = \frac{1}{50\times10^{-3}} + j\ 2\pi f\cdot 10^{-10}$
   - $Z_0(f) = \sqrt{\frac{Z(f)}{Y(f)}}$
   - $Z_0(500) = 400\Omega, Z_0(1000) = 282\Omega, Z_0(500) = 200\Omega,$
   - High resistance approximation: $\alpha(f) = \sqrt{\frac{\omega RC}{2}},$
   - $\alpha(500) = 8.8\times10^{-5}, \alpha(1000) = 12.6\times10^{-5}$
   - $\alpha(2000) = 17.6\times10^{-5}\times10^{5}$
   - For $100km$, loss is $\alpha\times10^{5}$



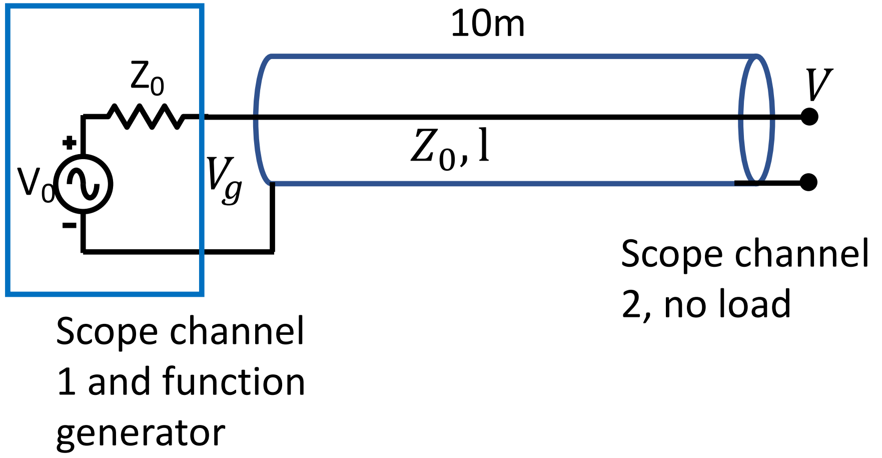
Exercise 11: Waves

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: $\alpha(f) = \frac{R}{2Z_0}$, $Z_0 = \sqrt{\frac{L}{C}} = 1000\Omega$
   - $\alpha(f) = \frac{R}{2Z_0(f)}$, $\alpha(500) = \alpha(1000) = \alpha(2000) = \frac{5\times 10^{-2}}{2000} = 5.5\times 10^{-5}\ nepers/m$
   - For $100km$, loss is $\alpha\times 10^5$=5.5 or $5.5\ \times 8.868 \approx 49dB$

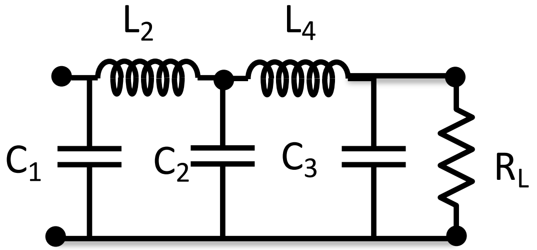
Exercise 12: Resonance

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

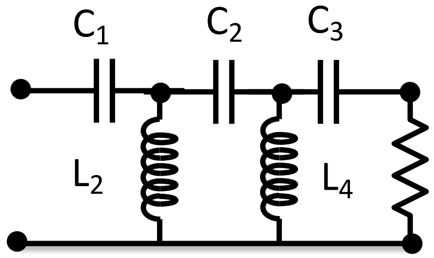
1. Derive an expression for $|\frac{V_g}{V}|$ and assuming α is small by finding the first resonance where V_g is minimum.
   - $V_g = V_0 e^{-j\beta l}\cos(\beta l), V = V_0 \exp(-j\beta l)$. $\left|\frac{V_g}{V}\right| = \cos(\beta l)$. So, at $l = \frac{\lambda}{2}, \left|V_g\right| = |V|$
2. Find 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 $2\times10^8\ m/s$]
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 $f_l = f_r - \frac{BW}{.2}$



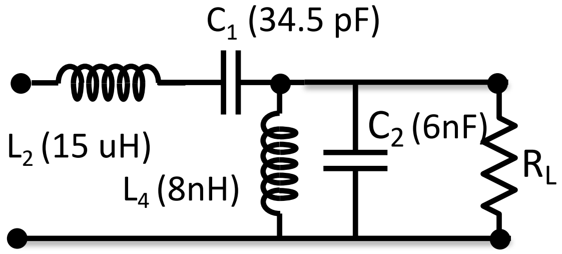
Filters



Low pass

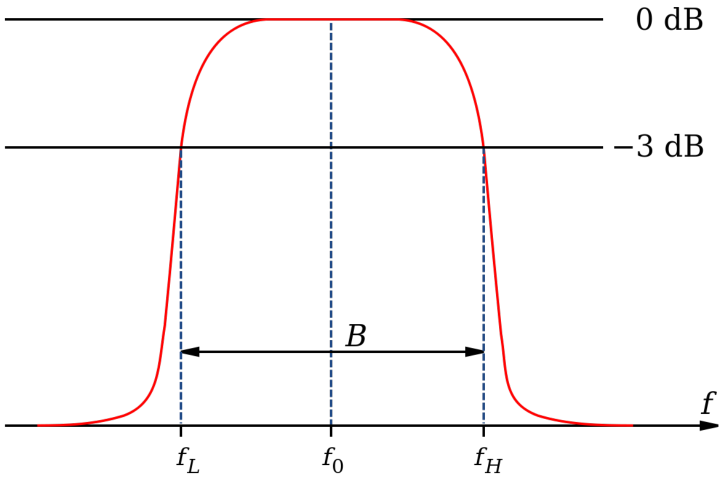


High pass

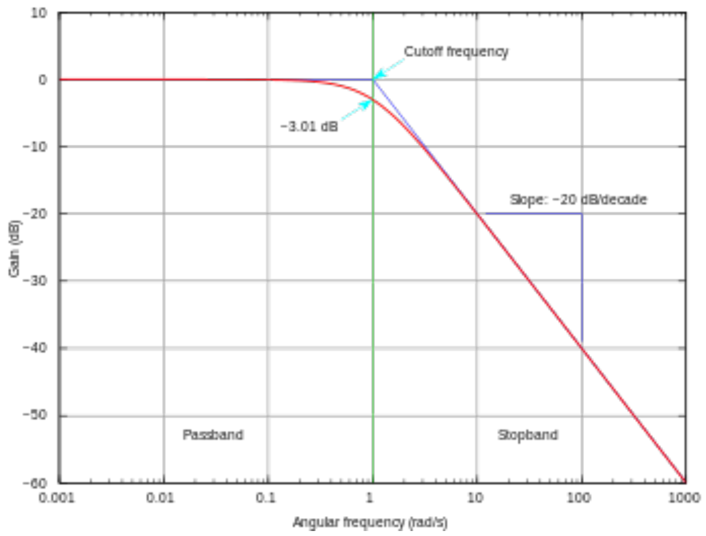


7 MHz bandpass

- 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!

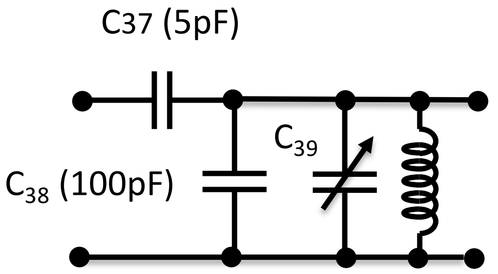


Bandpass - WIkipedia



Lowpass - WIkipedia

Norcal transmit bandpass filter



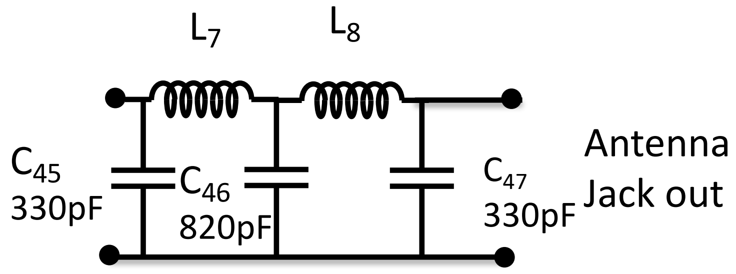
- $C_{39} = 50pF,$
- L_6 is 36 turns #28 on T37-2 which has $A_l = 4\frac{nH}{turn^2}, L_6 = A_l \cdot 36^2 = 3.1\mu H$
- $Z_2 = -\frac{j}{(C_{38}+C_{39})\omega_o},\ Z_3 = jL_6\omega_o, Z_1 = \frac{j}{C_{37}\omega_o}$
- $Z_{2,3-eq} = \frac{jL_6\omega_0}{L_6(C_{38}+C_{39})\omega_0{}^2-1}$ L6
- Resonance 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 $Z_{2,3-eq}$ $\omega_o = \frac{10^9}{\sqrt{465}} \approx 50.8\times10^6, f_0 = \frac{\omega_o}{2\pi} \approx 7.1\ MHz$
- 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.

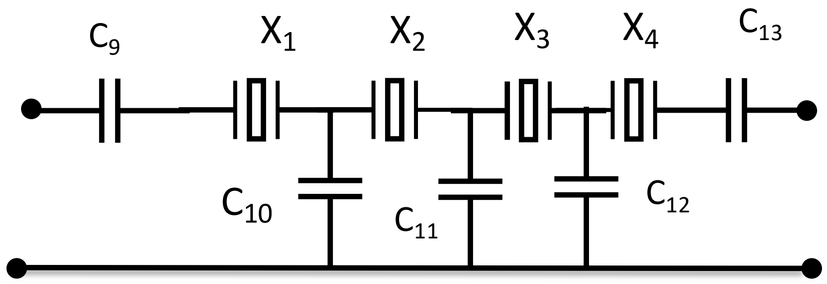
- Need Puff (a simulator) to get losses. However, there is a 6dB drop-off at every frequency doubling

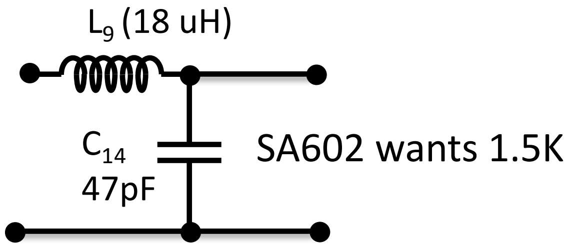


Exercise 14: Norcal IF Cohn Filter

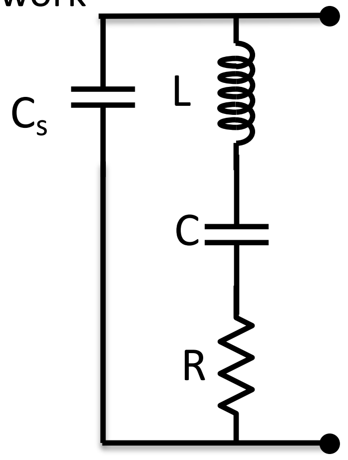
- X_1 through X_4 are 4.91 MHz
- C_{10}, C_{11}, C_{12} are 270 pF
- Set function generator to $50mV_{pp}$ from function generator
- Calculate R and X for filter

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

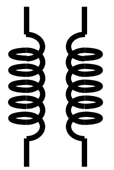
- For solenoid, $\oint B \cdot ds = \mu_0 n I$ inside
- $LI = \Phi_B$. Since there are n turns in the solenoid, over the solenoid, $LI = \mu_0 n^2 I$, so $L = \mu_0 n^2$.
- This is the source of $L = A_l n^2$
- $V_s = \frac{N_s}{N_p} V_p$
- $Z_p = (\frac{N_p}{N_s})^2 Z_s$

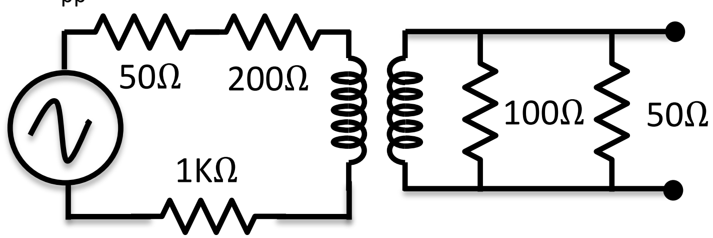
Exercise 15: Norcal Driver Transformers

- T_1 uses FT 37–43. $L(\mu H) = \frac{A_L t^2}{1000}$, $A_L = 350$. $f_r = 7\times10^6 MHz$, $n_p = 14$, $n_s = 4$, $\omega_r = 2\pi\times7\times10^6 MHz = 4.4\times10^7$

1. Measure the output V_{out}.
2. Calculate V_{out}
   - $L_p = 68.6\mu H,\ L_s = 5.6\mu H$
   - $Z_{eq,in}(\omega) = 1250 + j(\omega L_p)$, $Z_{eq,in}(\omega_r) = 1250 + 3016j$, $|Z_{eq,in}(\omega_r)| = 3264$
   - $Z_{eq,out}(\omega) = 33 + j\omega L_s$, $Z_{eq,out}(\omega_r) = 33 + j246$, $|Z_{eq,out}(\omega_r)| = 248$
   - $V_{t,in} = \frac{3016}{3264} V_{in}$
   - $V_{out} = V_{t,out} = \frac{n_s}{n_p} V_{t,in} = .29 V_{t,in} = .29\times\frac{3016}{3264}\times5 = 1.3V$
   - $i_p(\omega) = \frac{V_{in}}{|Z_{eq,in}|} e^{j\theta_p(\omega)}$, $\theta_p(\omega) = \arctan\left(\frac{\omega L_p}{1250}\right)$; $i_s(\omega) = \frac{V_{out}}{|Z_{eq,out}|} e^{j\theta_s(\omega)}$,$\theta_s(\omega) = \arctan\left(\frac{\omega L_s}{33}\right)$.
   - $P_{in,a} = Re\left(\frac{V_{in}\overline{I_{in}}}{2}\right) = Re\left(\frac{{V_{in}}^2}{2|Z_{eq,in}(\omega)|} e^{j\theta_p(\omega)}\right)$
   - $P_{out,a} = Re\left(\frac{V_{out}\overline{I_{out}}}{2}\right) = Re\left(\frac{{V_{out}}^2}{2|Z_{eq,out}(\omega)|} e^{j\theta_s(\omega)}\right)$

T_1, 14:4





Exercise 15: Norcal Driver Transformers

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

3. Measure the 3dB cutoff, f_c.

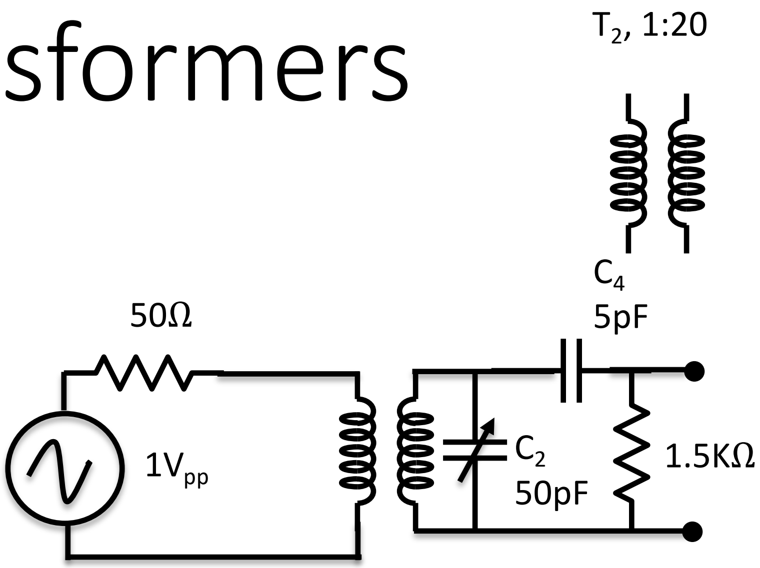
- $\frac{P_{out,a}(\omega)}{P_{in,a}(\omega)} = (\frac{V_{out}}{V_{in}})^2 \frac{|Z_{eq,in}(\omega)|}{|Z_{eq,out}(\omega)|} \frac{\cos(\theta_s(\omega))}{\cos(\theta_p(\omega))} = .15$

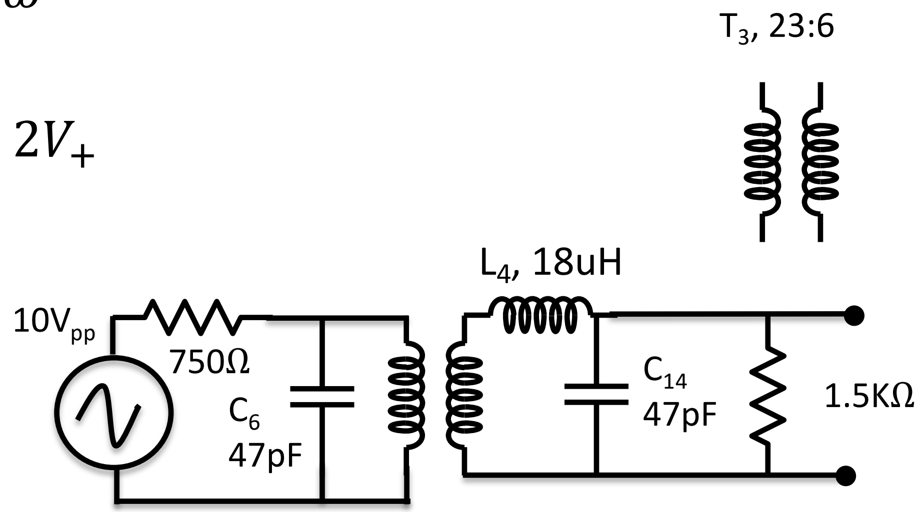
Exercise 16: Norcal Tuned Transformers

- 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. $P_+ = \frac{{V_+}^2}{2Z_0}$. $L_{in} = 55nH, L_{out} = 22\mu H, \omega = 4.4\times10^7$
  - $Z = 50 + 2.4j, Z_0 = 203 - 4030j, \rho = 1$, so $V_+ = V_-, V = 2V_+$
  - $\frac{P}{P_+} = \frac{1}{4}$
- Similar calculation for T_3.





Acoustics

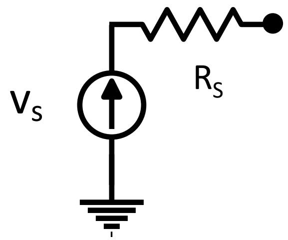
- $P = P_0 + P_e, \rho = \rho_0 + \rho_e$

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: $P = f(\rho)$, $P_0 + P_e = f(\rho_0 + \rho_e) = f(\rho_0) + \rho_e f'(\rho_0), f'(\rho_0) = \kappa = (\frac{dP}{d\rho})_0$, 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 $PV = Nkt$, $\kappa = (\frac{dP}{d\rho})_0 = \frac{\gamma k T}{n}$
- $L_p = 20 \log(\frac{P}{P_0}), P_0 = 20\ \mu Pa$

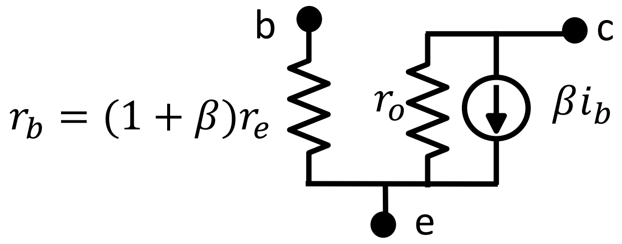
Sound	L_p	Power density
rustling leaves	10dB	1pW/m^2
broadcast studio	20dB	1pW/m^2
classroom	50dB	10nW/m^2
heavy truck	90dB	1nW/m^2
Shout at 1m	100dB	10mW/m^2
jackhammer	110db	100mW/m^2
jet takeoff at 50m	120dB	1W/m^2

Bipolar Transistors - I



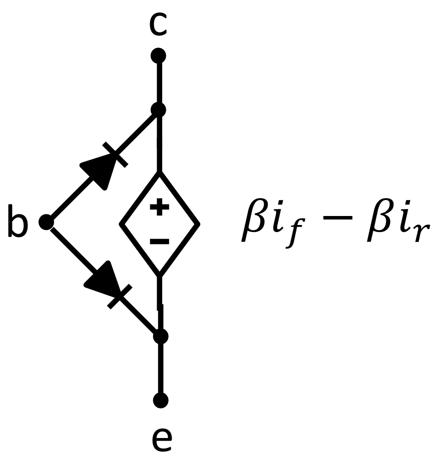
Bipolar source model

- NPN Model
  - $V_f \approx .7V, V_s \approx .2V$
  - Conducts when $V_{be} > V_f$
  - $i_c = \beta i_b$
  - $i_c = \alpha i_e$
  - $\beta = \frac{\alpha}{1-\alpha}$ [$= h_{fe}$, 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
  - $G_s = \frac{i_b}{15mV}$
  - $R_S = 2\Omega$

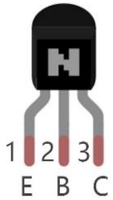


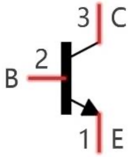
Bipolar equivalent circuit

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



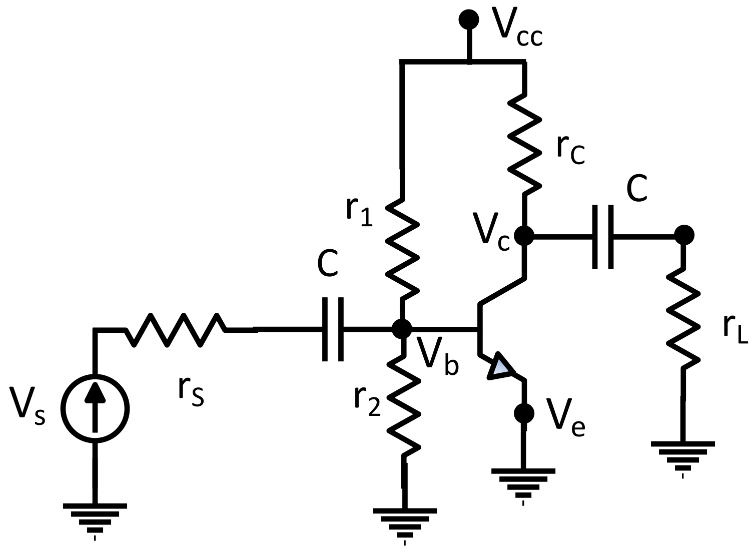
Bipolar model





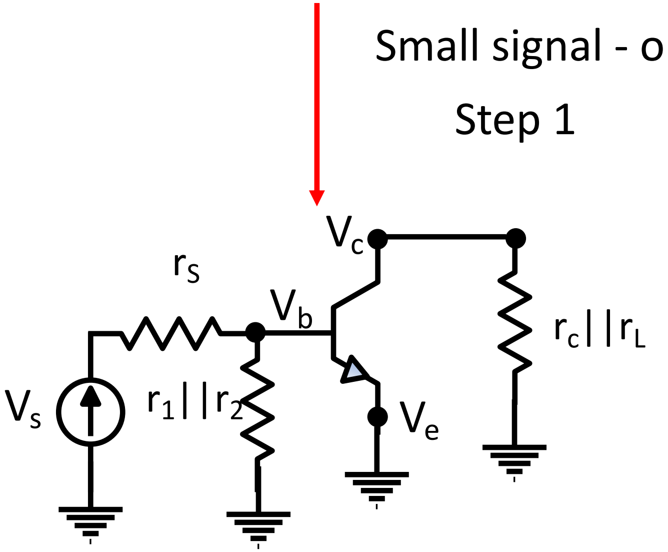
Bipolar Transistors - II

- 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. Thevienize
  3. Replace transistor with model

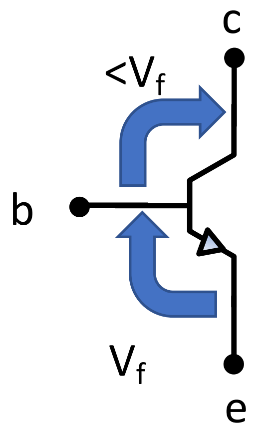


Small signal - original

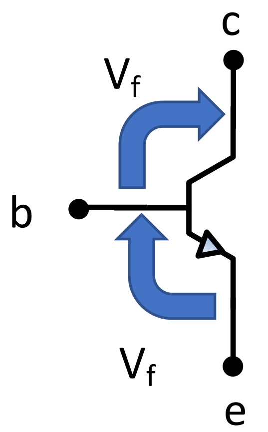
Step 1



Small signal



Active

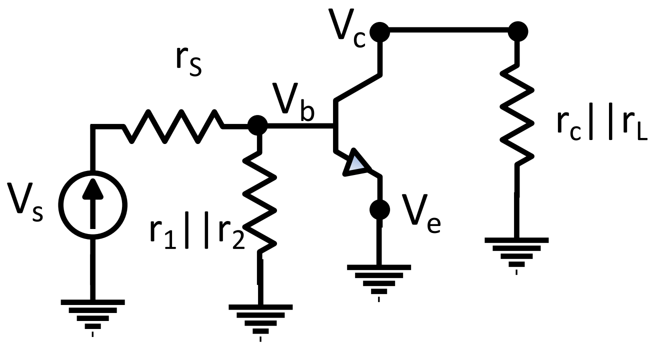


Saturated

Bipolar Transistors - III

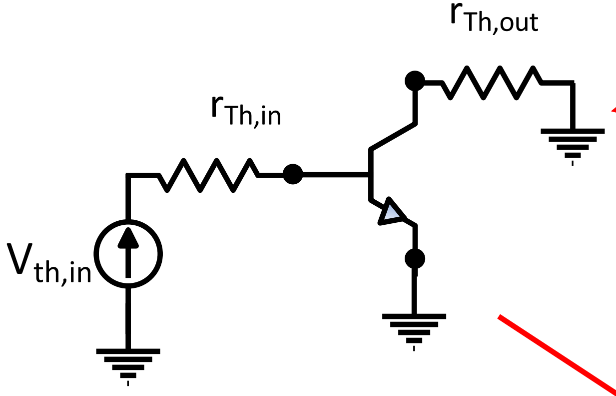
- Small signal
  1. Convert to AC only and simplify
  2. Thevienize
     - $V_{th,in} = V_s \frac{r_1||r_2}{r_1||r_2+r_s}$
     - $r_{th,in} = r_s||r_1||r_2$
     - $r_{th,out} = r_C||r_L$

  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
     - $A_{gail} = \frac{V_{out}}{V_s}$



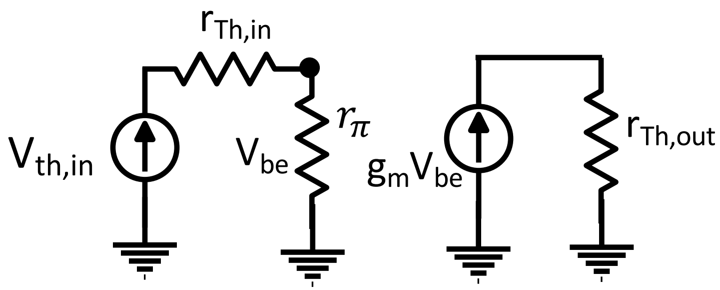
Small signal

Step 2



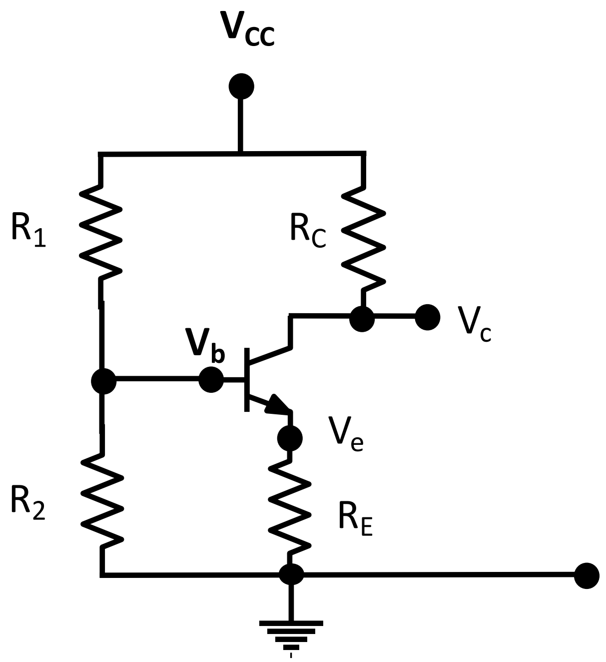
Small signal

Step 3



Small signal

Transistor experiment



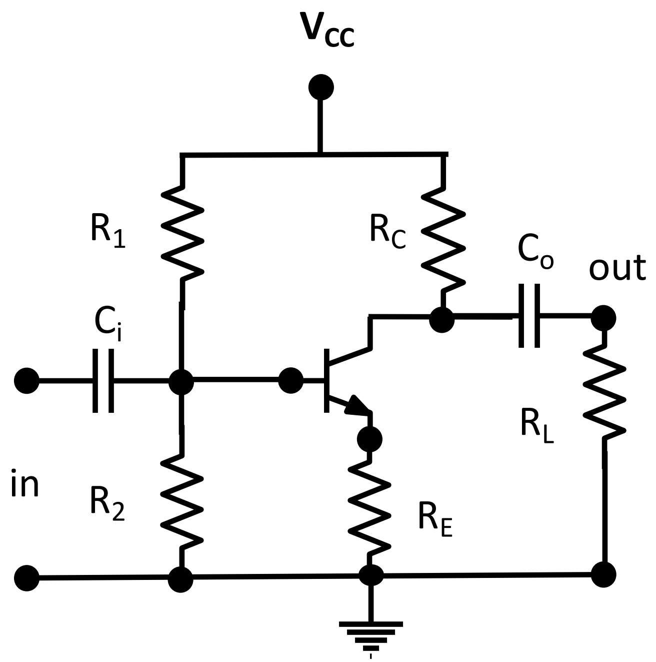
Experiment A

- $V_{cc} = 9V$, $R_1 = 22.8k\Omega$, $R_2 = 7.2k\Omega$, $R_c = 1k\Omega$, $R_E = 220k\Omega$. 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, $V_{cc} = 9V$, $R_1 = 20k\Omega$, $R_2 = 10k\Omega$, $R_c = 1k\Omega$, $R_E = 220k\Omega$. 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

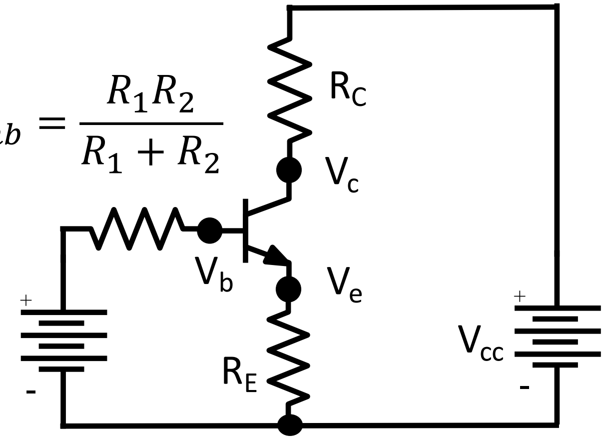


- Add C_i and C_o. Component values are:
  - $C_i = C_o = 1\mu F$
  - $R_1 = 20k\Omega, R_2 = 10k\Omega$
  - $R_C = 1k\Omega, R_E = 220\Omega$
  - $V_{cc} = 9V$

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

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



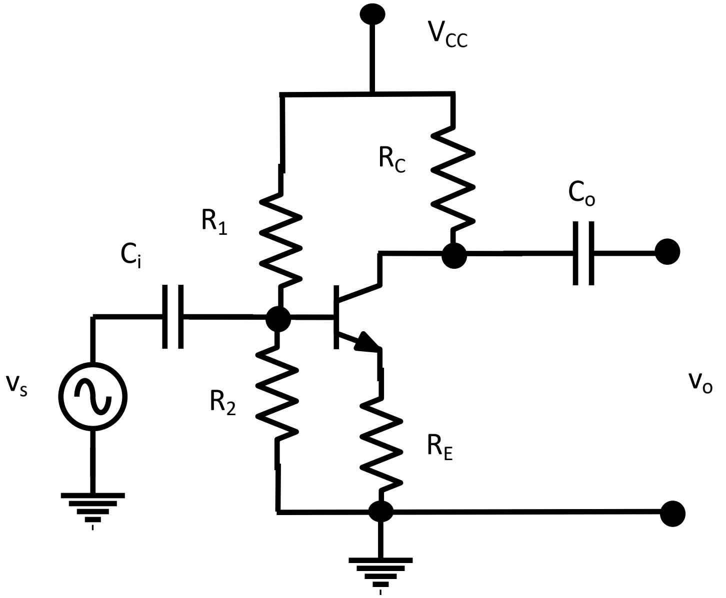
$$V_{thb} = \frac{R_1}{R_1 + R_2} V_{cc}$$

- 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



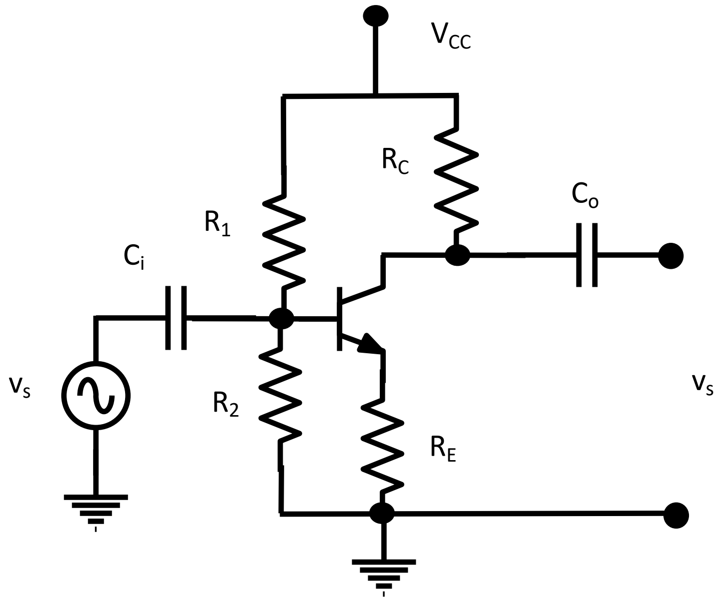
Common emitter amp

Credit: Ward, Hands on Radio.

- 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$\boldsymbol{\Omega}$.
5. Since A = 5 and $A=R_C/R_E$, $R_C= 5\ R_E$ so $R_E \sim 270\ \boldsymbol{\Omega}$ (this is a standard resistor value) and R_C= 1.5k$\boldsymbol{\Omega}$.

BJT common emitter amplifier continued



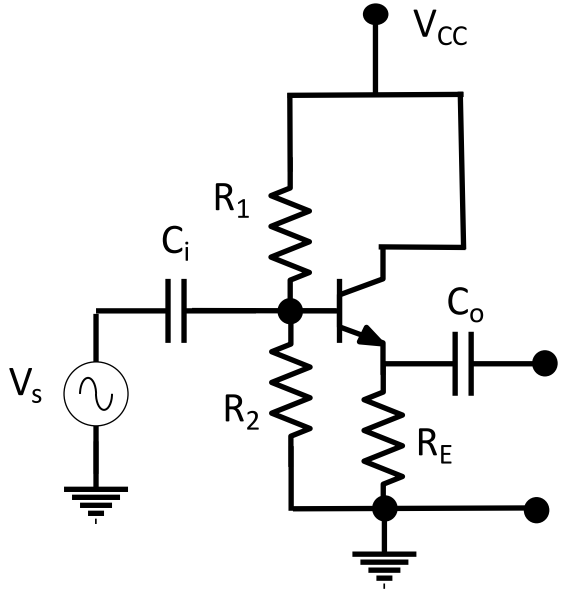
Common emitter amp

Credit: Ward, Hands on Radio.

6. i_b = 4mA/β = 27 μA.
7. Since V_{be} must be greater than .7V throughout the input signal range, we want the voltage across R_2 to satisfy $V_{be} + i_c R_E$ = 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 μA ~ 39 kΩ. $R_2 = 6.7k\Omega$
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 μF.

- 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 collector amplifier

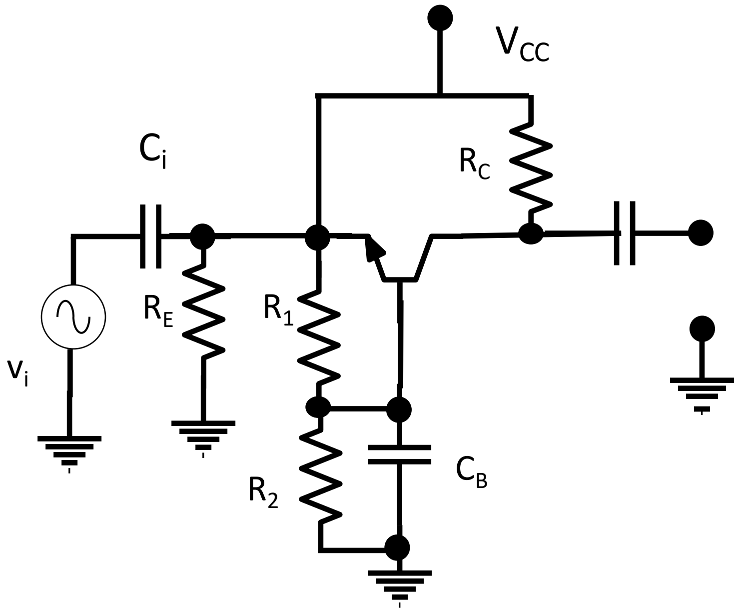


Common collector amp (Emitter Follower)

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. $i_{R_1 \rightarrow R_2} = 10 i_b$ (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. $Z_{in} = R_1 || R_2 || (\beta + 1) R_E$, $R_{in} = 50\Omega$, $Z_{out} = 5\Omega$

Credit: Ward, Hands on Radio.

BJT common base amplifier

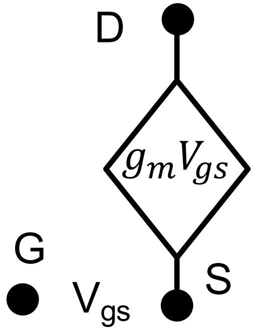


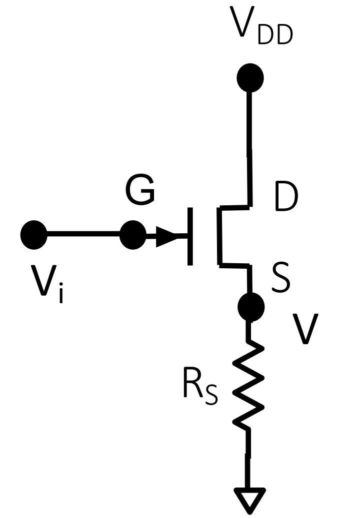
Common base amp

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

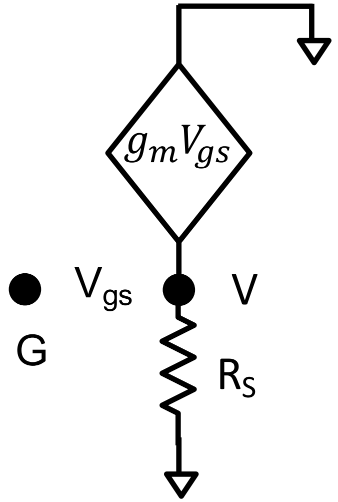
1. $V_{CC} = 12, V_{be} = .7V, R_E = 50\Omega, R_L = 1k\Omega, i_{ce,q} = 5mA, V_{ce,q} = 6V$
2. $i_b = \frac{i_{ce,q}}{\beta} = 33\mu A$, $i_{R_1 \rightarrow R_2} = 10\; i_b = 330\mu A$ (ROT)
3. $V_{R_2} = \; V_{be} + i_C R_E = 6.7V$, $V_{R_1} = 5.3V$
4. $R_1 = \frac{5.3}{330\mu A} = 16k\Omega$, $R_C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

JFETs





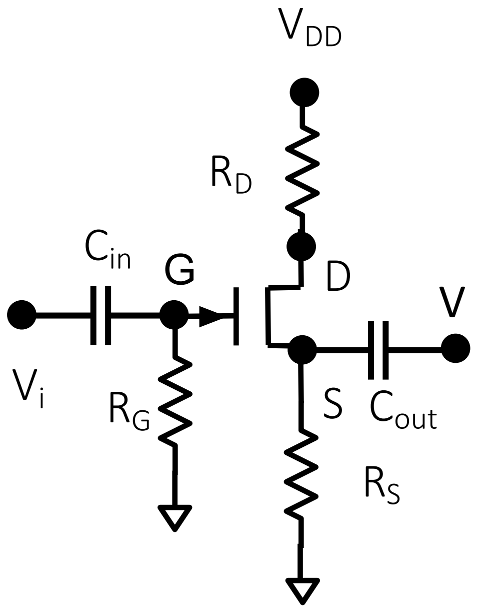
- JFET circuit model (active region):
  - $I_d = I_{dss}\,(1 - \frac{V_{gs}}{V_c})^2$, provided $0 < v_{gs} < V_c$ and $V_{ds} > V_{gs} - V_c$. i_{dss} is drain to source current when gate is at 0.
- $g_m = \frac{dI_d}{dV_{gs}} = \frac{\Delta i_{ds}}{\Delta v_{gs}} \approx -\frac{2I_{dss}}{V_c}(1 - \frac{V_{gs}}{V_c})$
- 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: $V = R_S g_m V_{gs}$, $v_{gs} = V_g - V$



AC model

JFET common drain (follower) amplifier

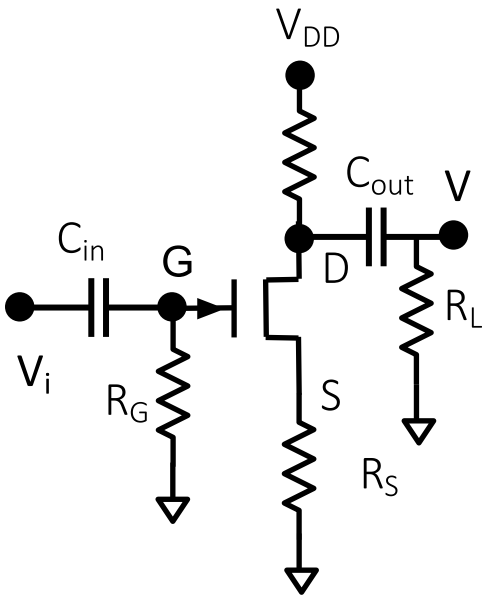
- $A_V = 1, Z_{out} = R_S || \frac{1}{g_m} \approx \frac{1}{g_m}$ (if $g_m R_S \gg 1$)



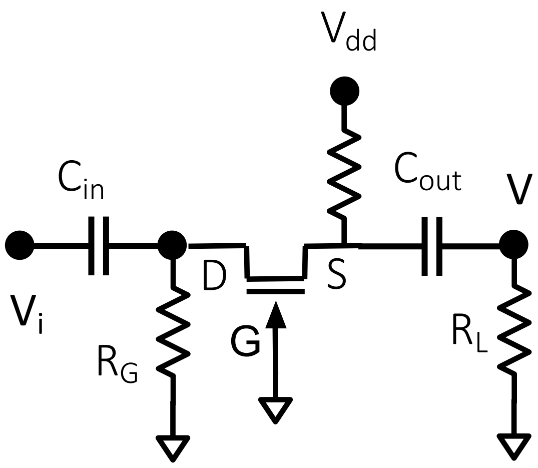
JFET common source amplifier

- $A_V = -\frac{g_m R_D}{1+g_m R_S} = -\frac{R_D}{R_L}, Z_{in} = R_G, Z_{out} \approx R_D$
- $R_S = \frac{-V_P}{i_{dd}}(1 - \sqrt{\frac{i_{dq}}{i_{dss}}}).\ \ g_m \approx 15mA/V, Z_0 = \frac{1}{g_m}$

1. $V_{dd} = 12V, i_{dss} = 35mA, V_P = 3.0V, A_V = 10, i_{dd} = 10mA$
2. From equation above, $R_S = 139\Omega$, $R_D = 10R_S = 1390\Omega$
3. $A_V = -g_m(R_D||R_L)$



JFET common gate amplifier

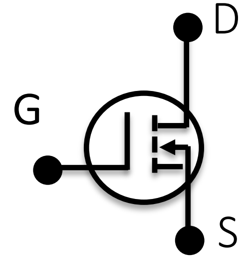


- $A_V = g_m(R_D||R_L), Z_{out} \approx r_0(g_m R_S + 1)||R_D, Z_{in} = R_S||\frac{1}{g_m}$
- $V_{DD} = 12V, i_{dss} = 60mA, V_P = -6, A_V = 10, R_L = 1k\Omega, R_S = 50\Omega$
- $i_{d,q} = \frac{V_P}{2{R_S}^2 i_{dss}}\left(V_P + \sqrt{{V_P}^2 - 4R_S i_{dss} V_P}\right) - \frac{V_P}{R_S}$

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

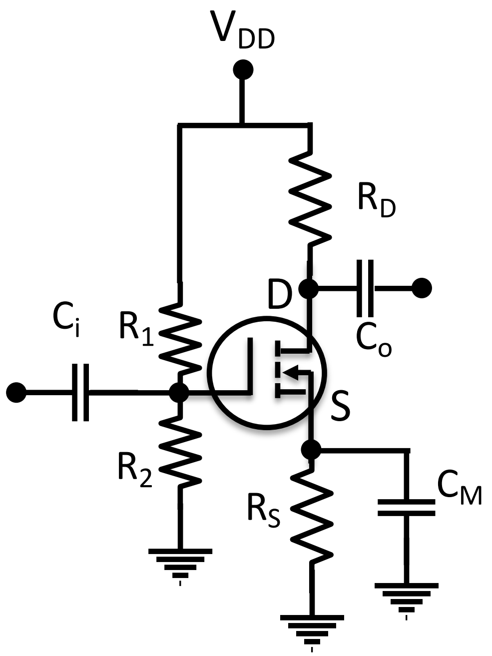
MOSFETs

- MOSFET circuit model, V_{Th} is threshold voltage, W is channel width, L is channel length
  - Ohmic: $I_D = k\left(\frac{W}{L}\right)[2(V_{GS} - V_T)V_{DS} - {V_{DS}}^2]$
  - Saturation: $I_D = k\left(\frac{W}{L}\right)(V_{GS} - V_T)^2 = I_{DS}$



N channel MOSFET

CMOS common emitter amplifier



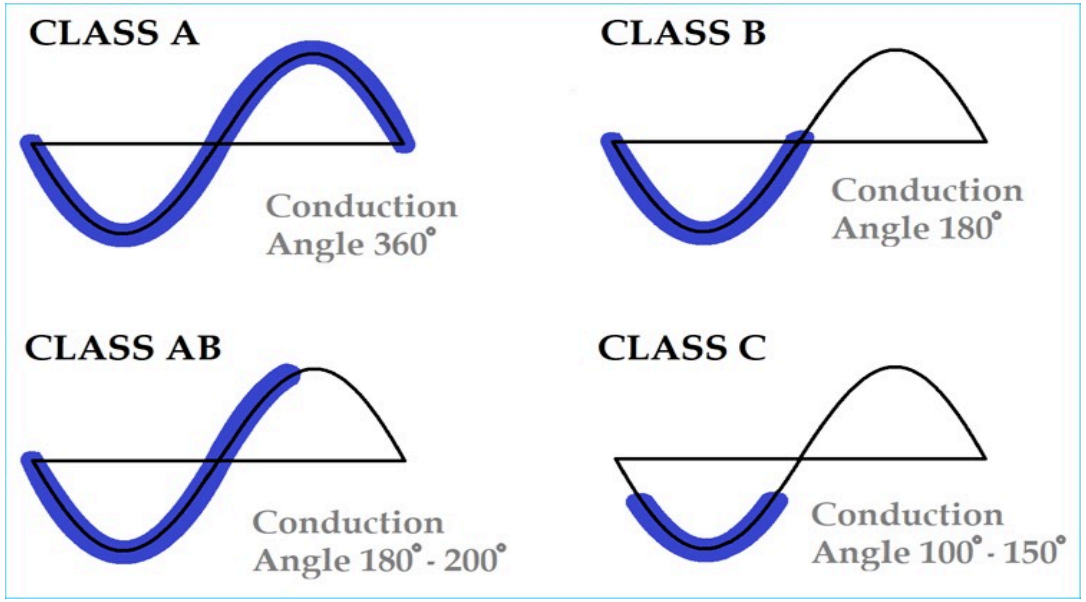
- Pick power
- $V_{DD} = i_D R_D + V_{DS} + i_D R_S$
- $V_{GS} = V_G - i_S R_S$
- $V_G = V_{DD} \frac{R_1}{R_1 + R_2}$
- $i_D = k(V_G - V_{TH})^2$
- Bias around $\frac{V_{DD}}{3}$
- Pick gain, $A = \frac{R_D}{R_S + \frac{1}{g_m}}$

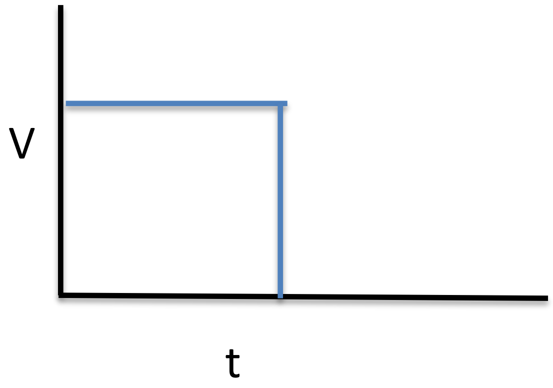
Amplifier classes

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

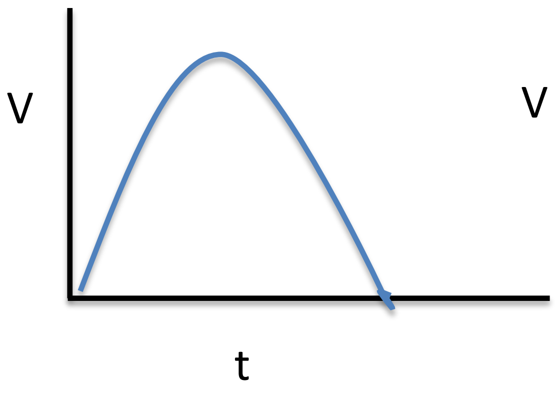
$$\eta = \frac{P}{P_0}, P_d = P_0 - P_i$$

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

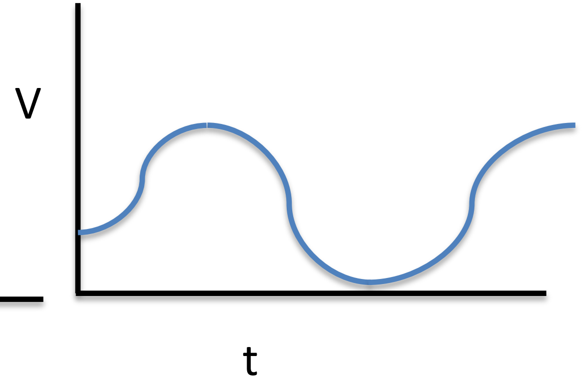




Class D



Class E



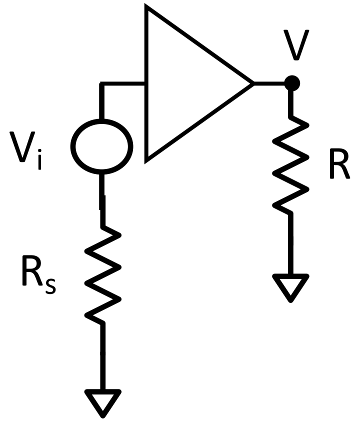
Class F

Efficiency of class A amplifiers

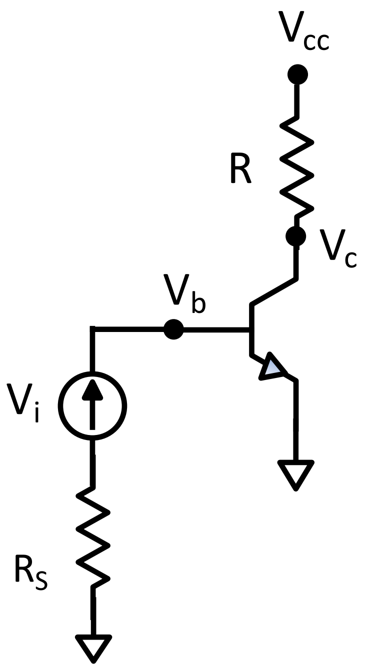
- Here, P_o is the power from the supply, P_d is the dissipated power, P is the output power. R is the collector resistor.
- $\eta = \frac{P}{P_0}$, P_0 is DC power
- $P_0 = V_{cc} I_0$, where $I_0 = < i_c >$, so $I_0 = \frac{V_{cc}}{2R}$ (R is the collector resistance). Thus, $P_o = \frac{{V_{cc}}^2}{2R}$.
- 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. $V_+ = \frac{V_0}{2}$
- $G = 10\log(\frac{P}{P_+})$
- $P = \frac{{V_{pp}}^2}{8R}$
- $P_+ = \frac{{V_{+,pp}}^2}{8R_S}$
- $G_S = 10\log(\frac{V}{V_+})$

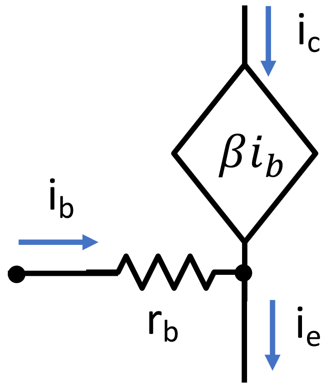


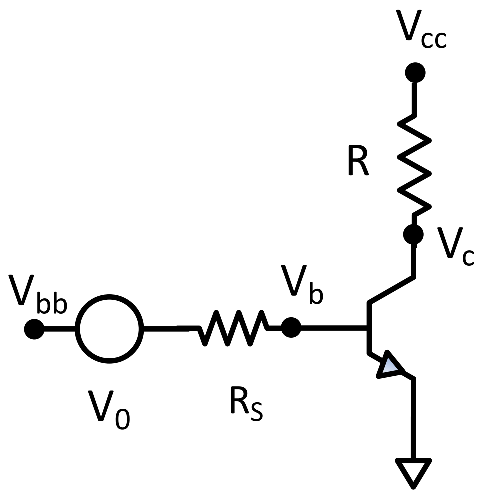
Bipolar transistors - IV



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

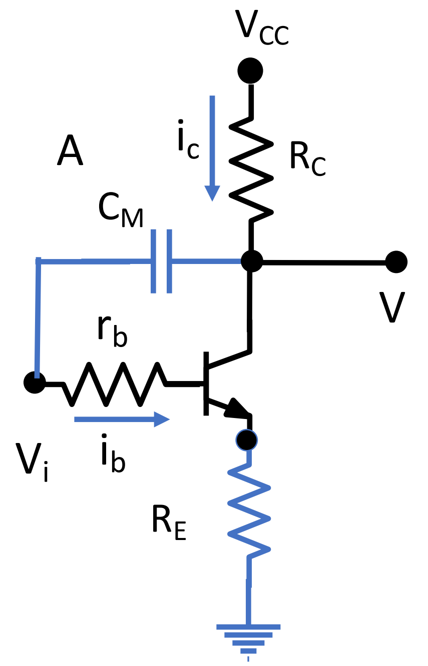


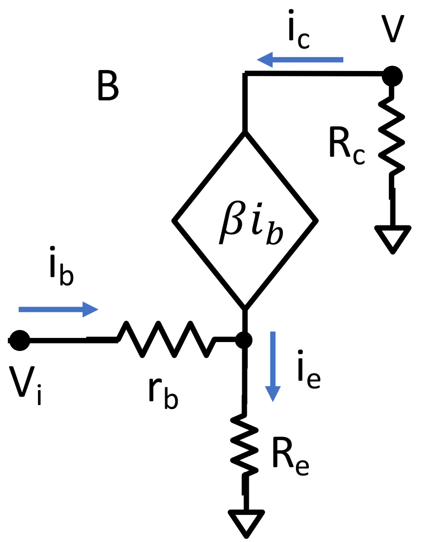


Emitter degeneration

- 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 $Z_i = \beta R_E$.

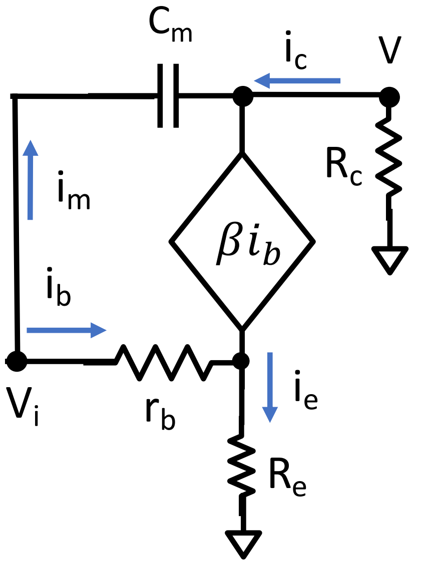


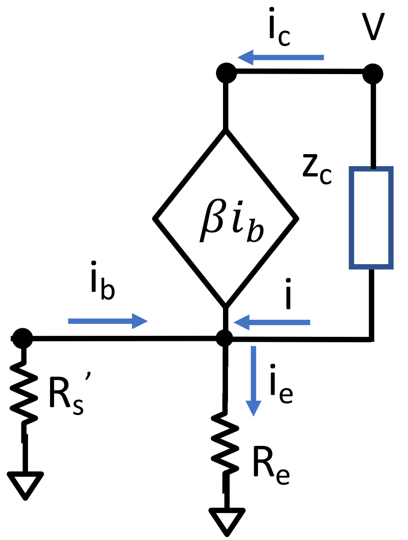


Emitter degeneration

- 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.

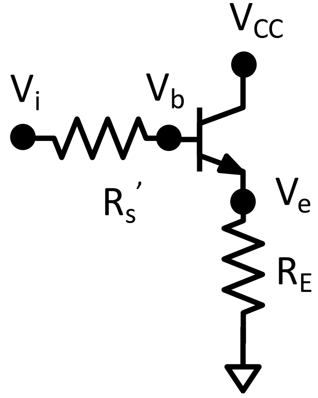
- With the Miller capacitor, $Z_i = \beta R_E || (1+|G_v|)C_M$
- $r_c \approx \frac{V_{early}}{i_c}$, r_c is the collector resistanc.e
- $R_s{}' = R_S + r_b$, r_b is the base resistance. $R_S{}'$ is the combined source resistance.
- z_C is called the collector impedance and $z_c = r_c || C_C$, C_c is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





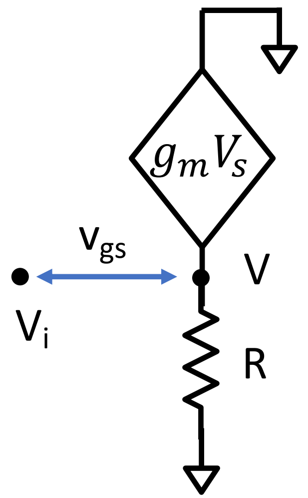
More on emitter follower

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



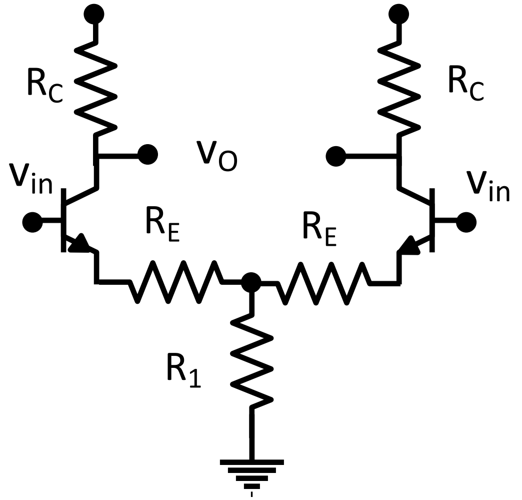
JFET follower

- $V = Rg_m v_{gs}$
- $v_{gs} = V_i - V$
- $V = \frac{RV_i}{R+\frac{1}{g_m}}$
- $Z_0 = \frac{1}{g_m}$
- $G_v = \frac{V}{V_i} = \frac{Rg_m}{1+Rg_m} \approx 1$



Differential Amplifier

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

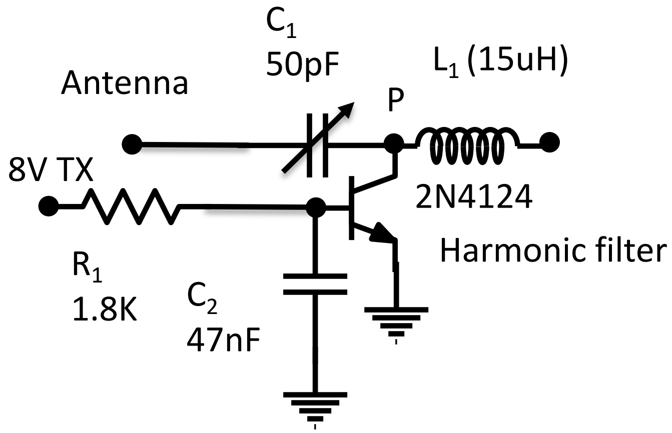


Differential amplifier

- 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

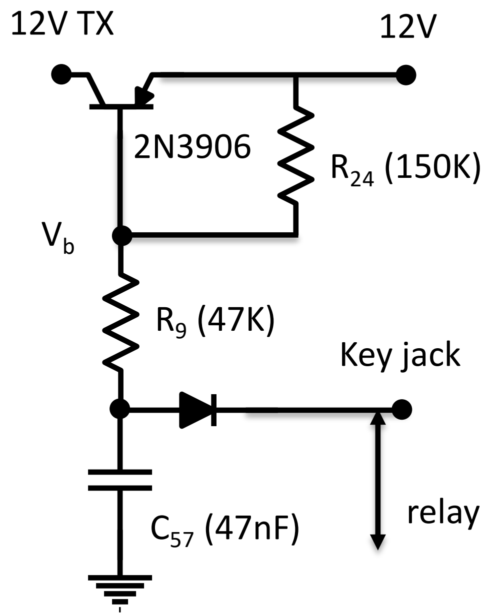
1. Consider the rising part of the base voltage waveform. Calculate slope.
2. Do the same for the falling part for voltage below .6V. Calculate t_2.
3. Measure the switch attenuation
   - When the transistor is saturated, the drop across ce is 1.4V. At full power, $P = \frac{{V_m}^2}{8R}$ and $V_m = 33.9V$. $\frac{P_{new}}{P_{original}} = \frac{1}{33.9^2}$, 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, $i_b = \frac{2i}{100} = 140\mu A$.

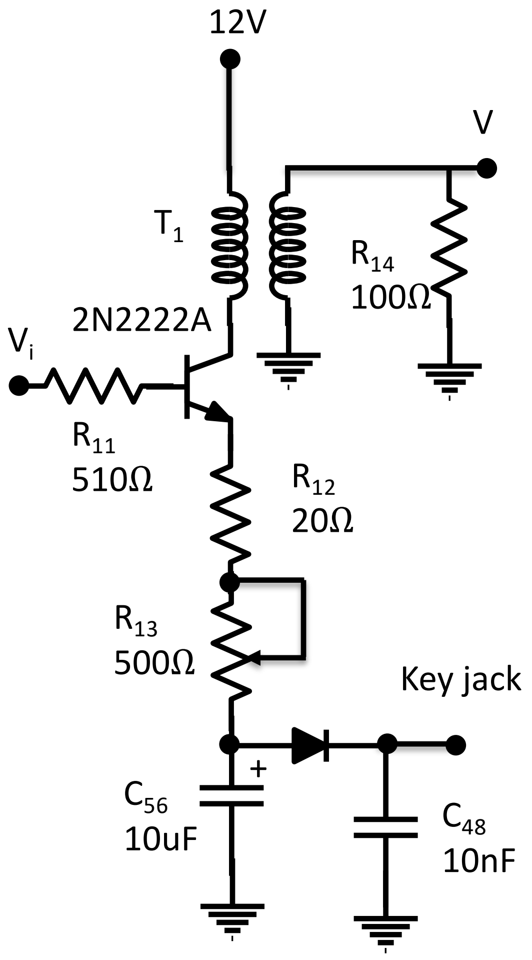
1. Calculate voltage on C_{57}. Measure time for capacitor to charge half-way. Calculate what the time should be.
   - $\tau = 197\times10^4\times47\times10^{-9} = 9.2\times10^{-2}sec = 92msec.$
2. Calculate the approximate current i_c when Q_4 is on. Assume base voltage on Q_1 is 700 mV. Neglect saturation 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

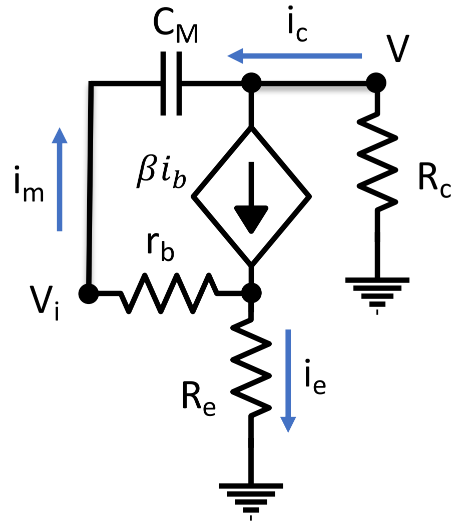
1. Measure the output voltage and calculate the power, P.
2. Calculate the power from the power supply.
3. Measure the voltage gain $G_v = \frac{V}{V_i}$ with R13 at minimum and maximum gain.

- $\omega = 4.4\times10^7$, $L_{p,T1} = 68.6\mu H$. This is a class A amplifier.
- $R' = n^2R, n = \frac{14}{4}, \; R' = 1225\Omega$
- $Z_{eq}(R) = (20 + R) + j\omega L_{p,T1}, 0 \leq R \leq 500. R = R_{13}$
- $Z_{eq}(0) = 20 + 2992j,$
- $i_c = \frac{V_{cc} - V_{ce}}{20 + R'}, i_c = \beta i_b, V_e = 20 i_c$
- 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

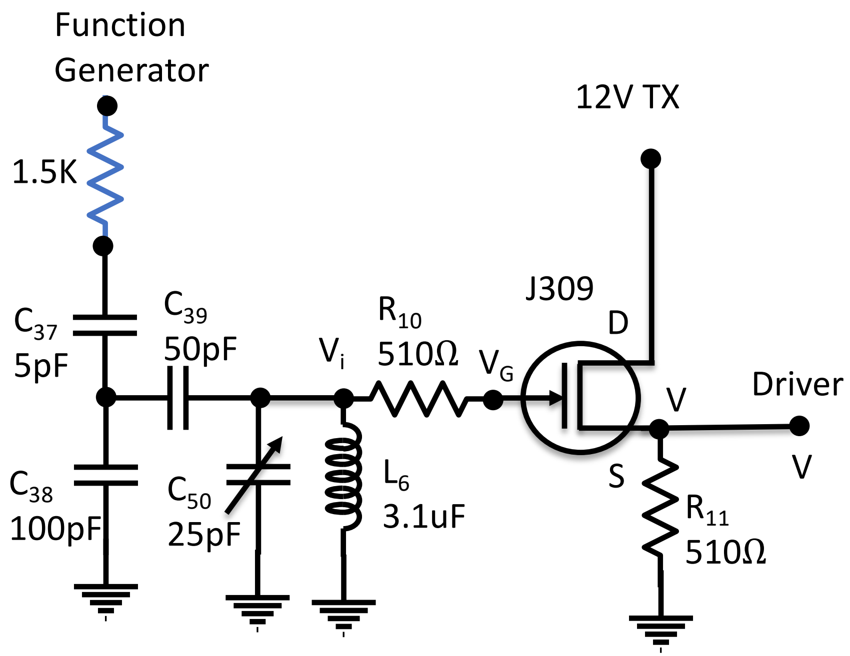
- In Driver amplifier, add probe to R_{11}, this allows us to measure the AC voltage, V_i



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 $\frac{V}{V_i} = -\frac{1225}{Z_i} \cdot \frac{i_c}{i_b} = -\beta \frac{1225}{Z_i}$

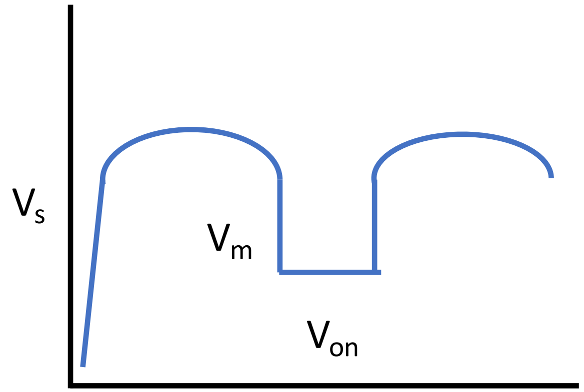
Exercise 23: Norcal Buffer amplifier

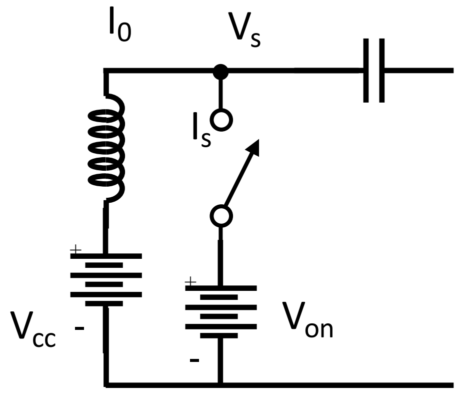
1. R_{11} determines the bias. Measure the DC voltage at source of the JFET (V).
2. Calculate the drain bias current. Calculate the source and drain voltages you should expect ($R = R_{11}$)
   - $V_{gs} = V_i - V,\ i_d = i_{dss}(1 - \frac{V_{gs}}{V_C})^2,\ V = g_m V_{gs} R$
   - $g_m \approx 12,\ i_{dss}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



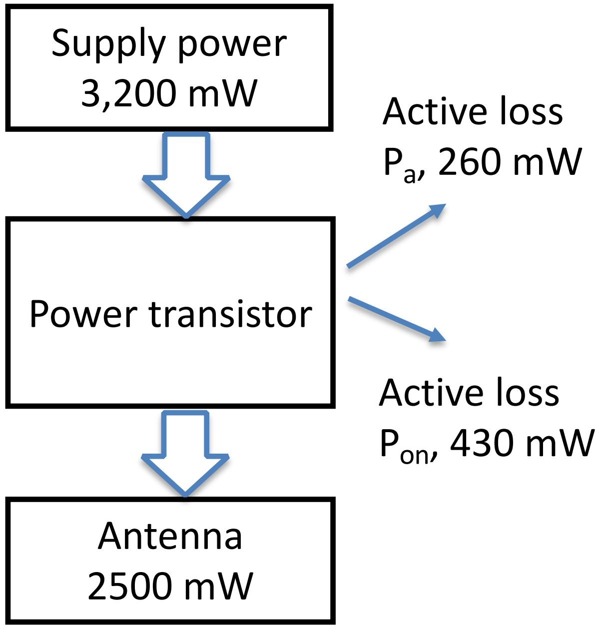
Class C amplifiers and Norcal 40 Power amp

- 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: $E = \frac{CV^2}{8R} = 37nJ$
- $P_a = Ef = 260mW$
- $i_c = i_0 - i_c = 215mA$
- $P_{on} = V_{on} i_{on} = 430mW$
- $P_d = P_a + P_{on} = 690mW$



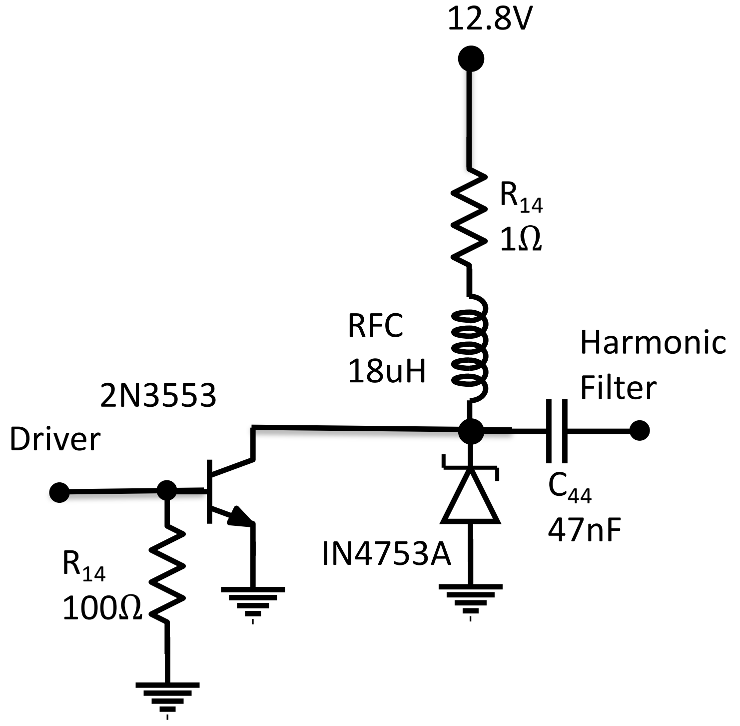


Switch model for class C amp



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

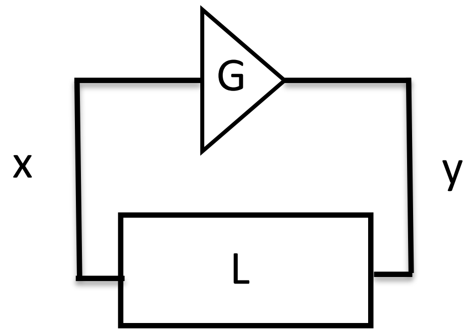
- For Motorola 2N3553, $T_j = 25\,{}^{C}/_{W}$

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

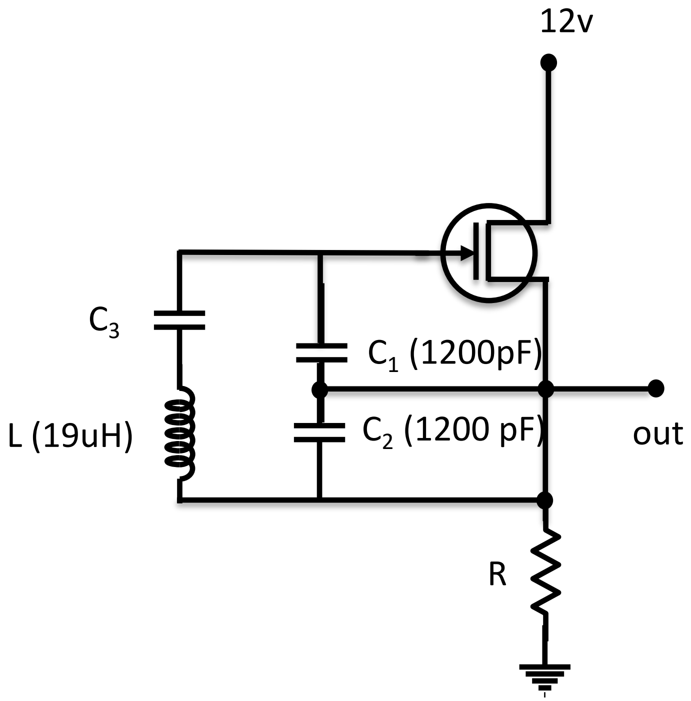
- Need measurements

Clapp oscillator

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

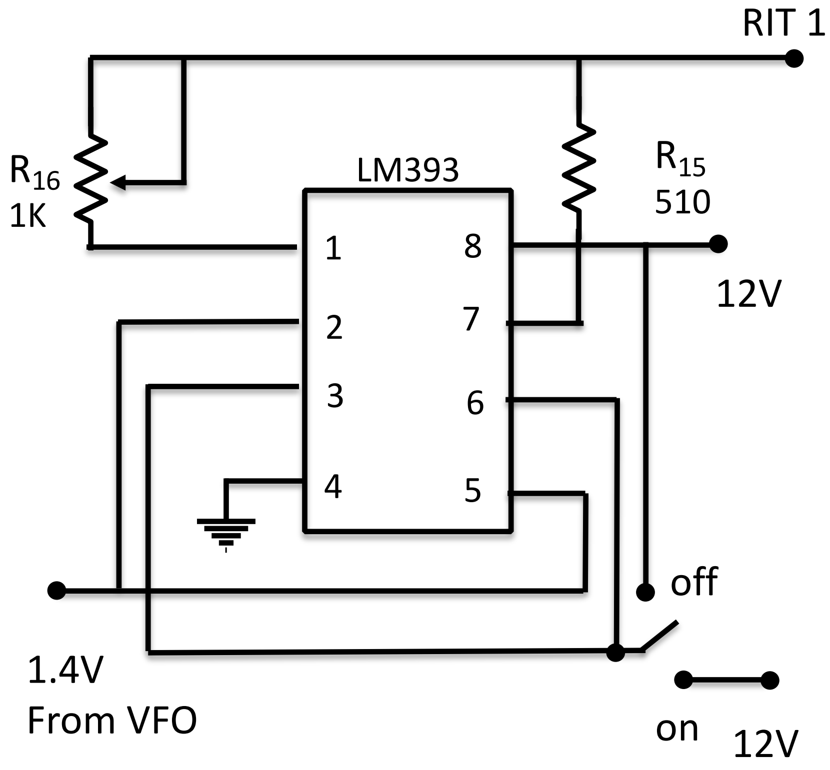


- Clapp (circuit on right)
  - $i_d = g_m v_{gs}$
  - Resonance: $-\frac{1}{j\omega_0 C_2} = j\omega_0 L + \frac{1}{j\omega_0 C_3} + \frac{1}{j\omega_0 C_1}$
  - $\omega_0 = \frac{1}{\sqrt{LC}}, C = C_1||C_2||C_3$
  - At resonance, $v_{gs} = Ri_d \frac{C_1}{C_2}, L = \frac{C_1}{RC_2}$
  - Oscillation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



Norcal Receiver Incremental Tuning (RIT)

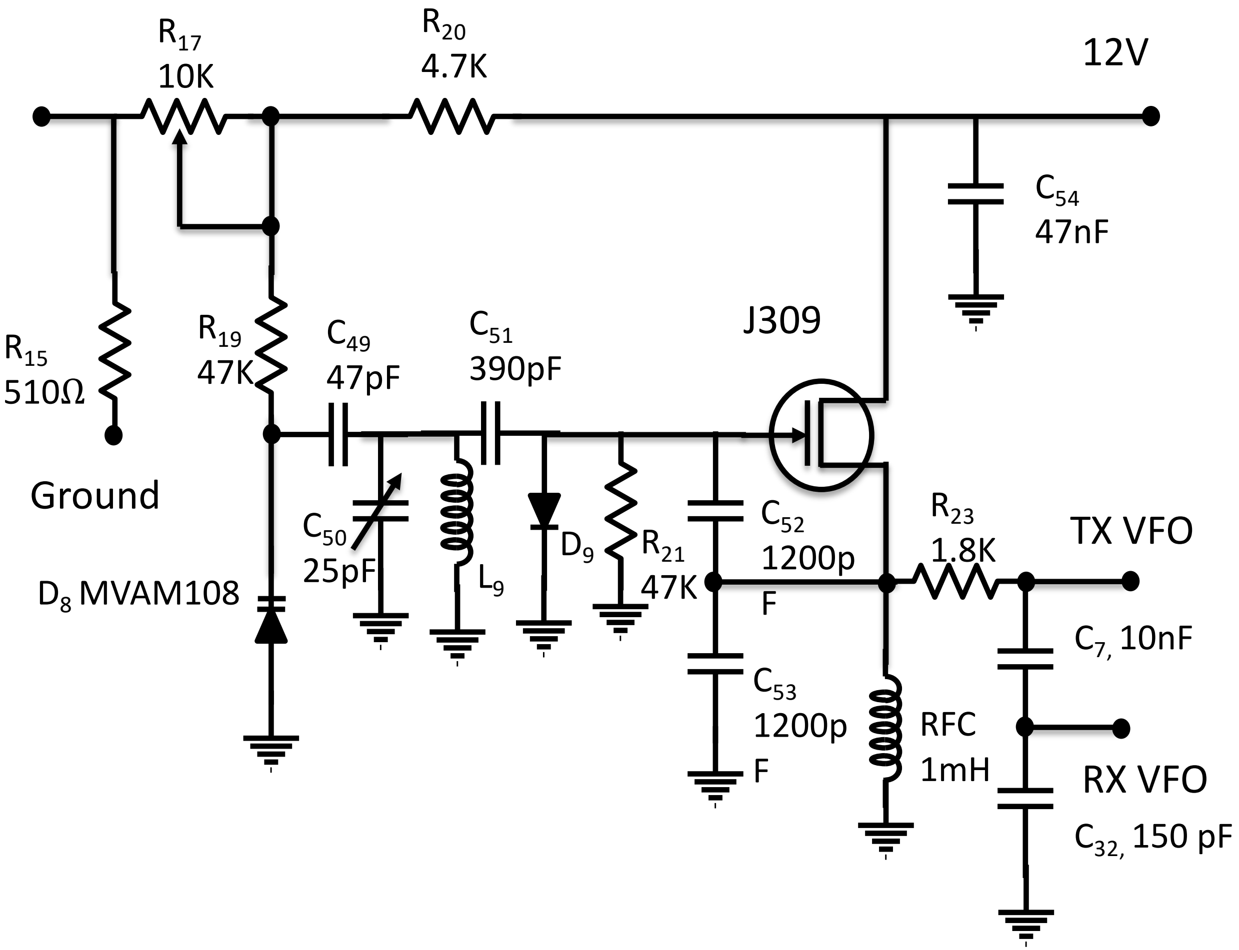
- LM393 is a comparator
- RIT allows transmit and receive frequency to be offset.
- If transmitter is on, TX will be 8V and the left comparator will be off, the right one on and R_{15} will be grounded.
- For receiving, TX is <1.4V, disconnecting R_{15} and shorting R_{16} to ground.



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 $g_m>1/R$

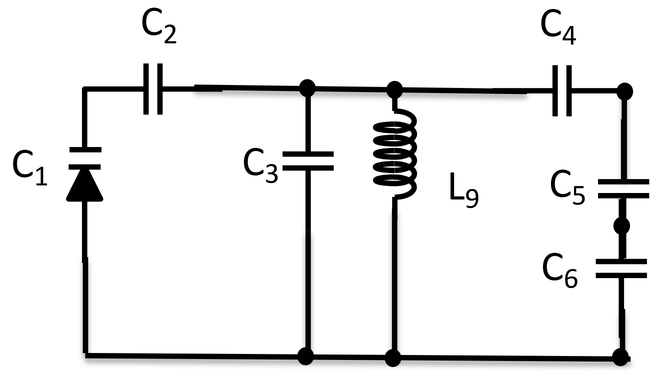
1. Measure p-p voltage, V. What should you expect?
2. Measure DC voltage across wiper in R_{17}
3. Calculate expected V for large signal oscillation
4. How does the frequency change as R_{17} changes?
5. Calculate the oscillation frequency and the loss ratio $|V/V_l|$
6. How would this change if you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



VFO Problem

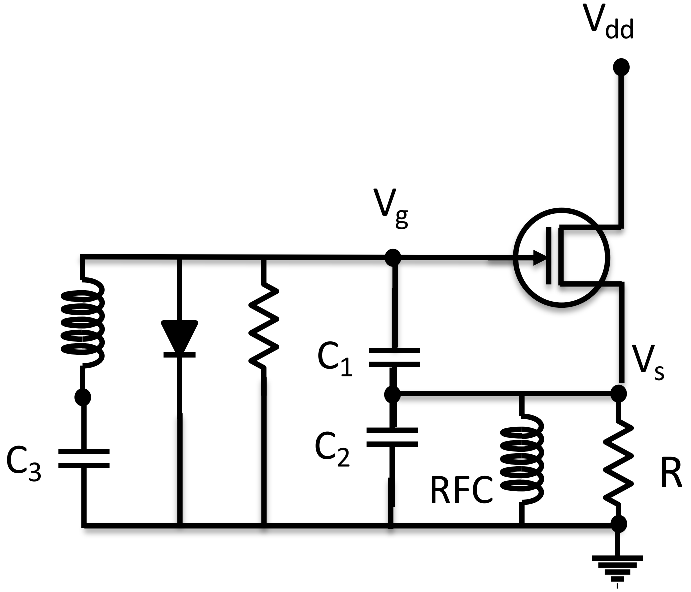
- The figure on the right is an equivalent circuit for the oscillator. The varactor, C_1 varies from 30 to 600 pF depending on the voltage.
- $C_2 = 47pF, C_3 = 7pF, C_4 = 390pF, C_5 = C_6 = 1200pF, L_9 = 19.2\mu H$
- The equivalent capacitance for $C_4 - C_5 - C_6$ is $C_{R,eq} = 236pF.$
- When $C_1 = 187pF$, 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

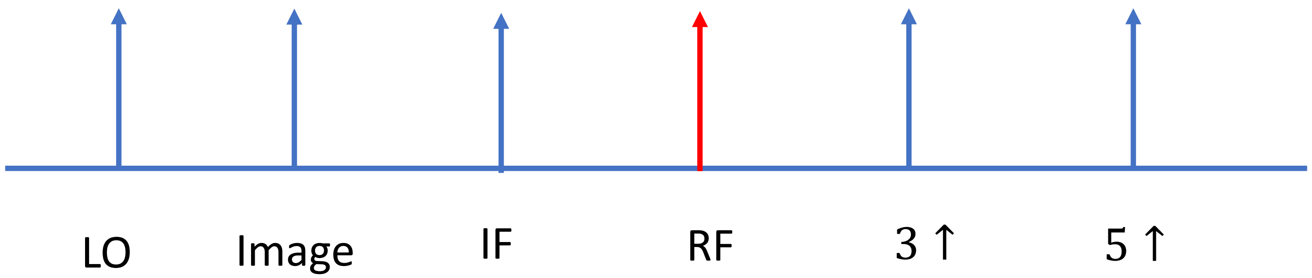
1. Measure the voltage, V, on R_{23}
2. In deriving the oscillation condition, we neglected the inductor resistance and drain source resistance, r_d. How does this affect the conditions. L_9 has a Q of 250 and $r_d = 5k\Omega$, now what is the predicted V.
3. Find the loss ratio $|\frac{V}{V_i}|$ and calculate what it should be.
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

Mixers

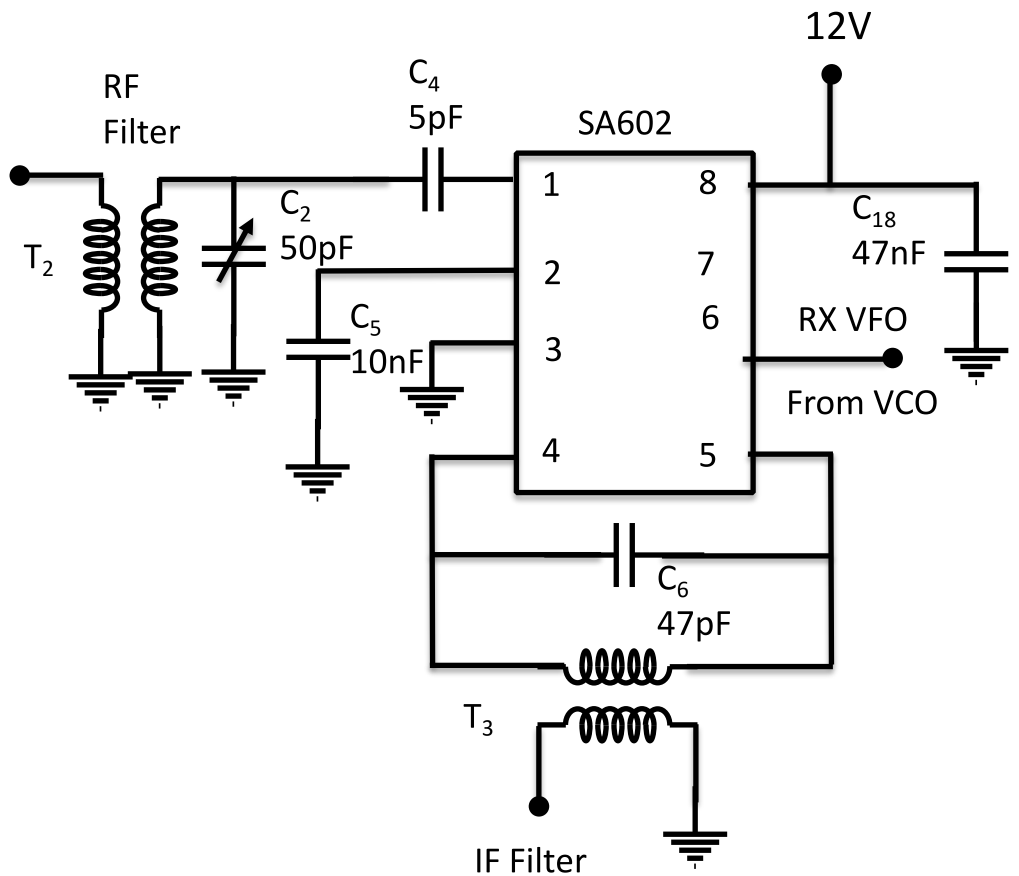
- $V_{lo}(t)$ is a square wave with period ω_{lo}. Expanding this in a Fourier series, we get:
- $V_{lo}(t) = \frac{4}{\pi}(\cos(\omega_{lo}t) - \frac{\cos(3\omega_{lo}t)}{3} + \frac{\cos(5\omega_{lo}t)}{5} \ldots), V_{rf}(t) = V_{rf}\cos(\omega_{rf}t)$
- $V_{lo}(t)V_{rf}(t) = \frac{2V_{rf}}{\pi}(\cos(\omega_{-}t) - \frac{\cos(3\omega_{-}t)}{3} + \frac{\cos(5\omega_{-}t)}{5} \ldots) + \frac{2V_{rf}}{\pi}(\cos(\omega_{+}t) - \frac{\cos(3\omega_{+}t)}{3} + \frac{\cos(5\omega_{+}t)}{5} \ldots)$
- $\omega_{+} = \omega_{lo} + \omega_{rf}$ and $\omega_{-} = |\omega_{lo} - \omega_{rf}|$
- We define $\omega_{k+} = (k\omega_{lo}+\omega_{rf})$ and $\omega_{k-} = |k\omega_{lo} - \omega_{rf}|$ and $V_{k+}(t) = \frac{2V_{rf}}{k\pi}\cos(\omega_{k+}t)$ and $V_{k-}(t) = \frac{2V_{rf}}{k\pi}\cos(\omega_{k-}t)$
- $\omega_i = \omega_{if} - \omega_{lo}$ and $\omega_{if} = \omega_{if} + \omega_i$, ω_i is a spurious signal. ω_{k+} and ω_{k-} are the spurs from the kth harmonic



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 $f_{\downarrow 5}$

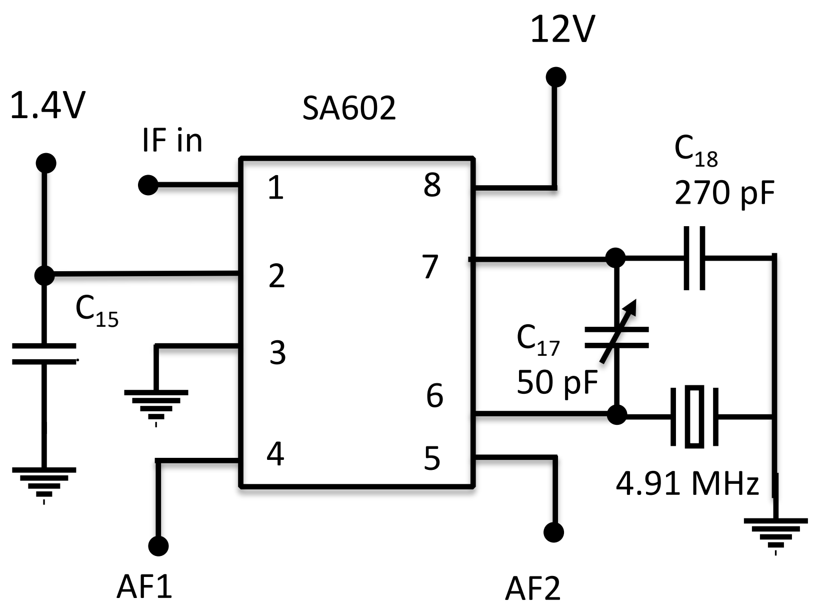
- Need measurements



Exercise 29: Norcal Product Detector

1. 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

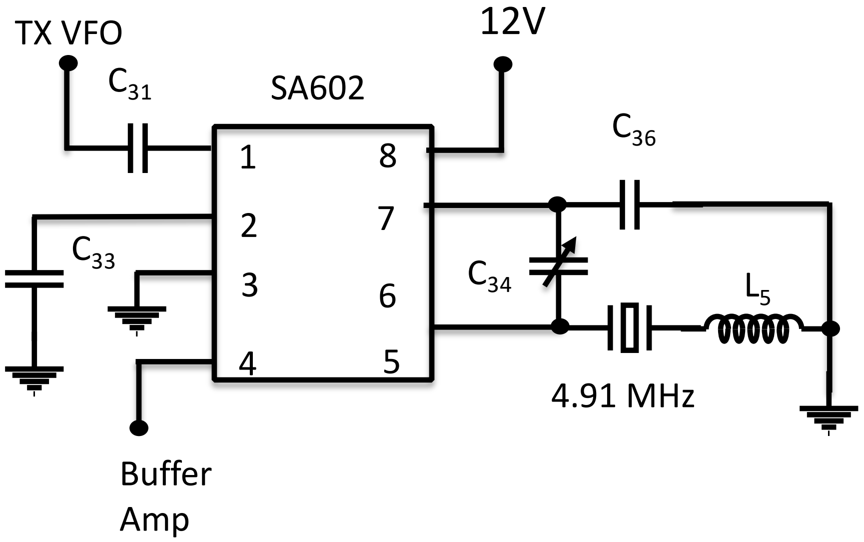
Product detector

x.

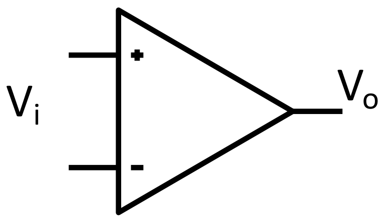
Exercise 30: Norcal transmit mixer and oscillator

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 $f_{mn}=mf_{vfo}+nf_{to}$.

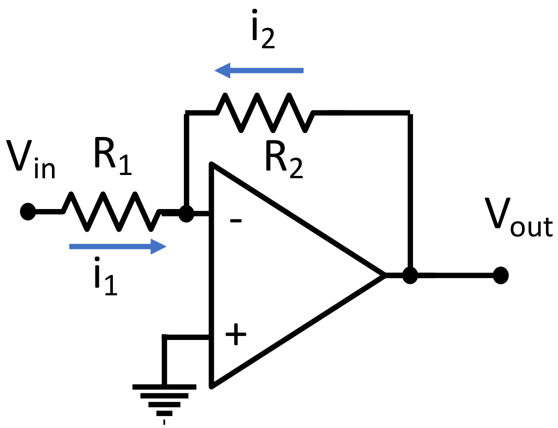
- Need measurements



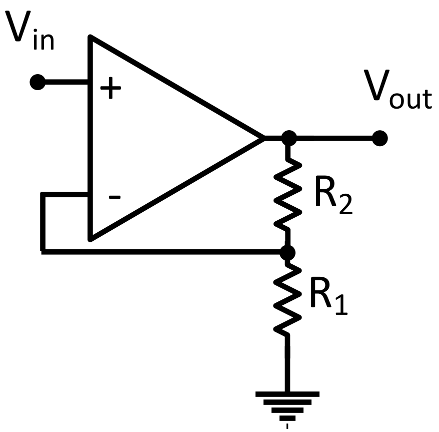
Op Amps



An op amp



Inverting amp



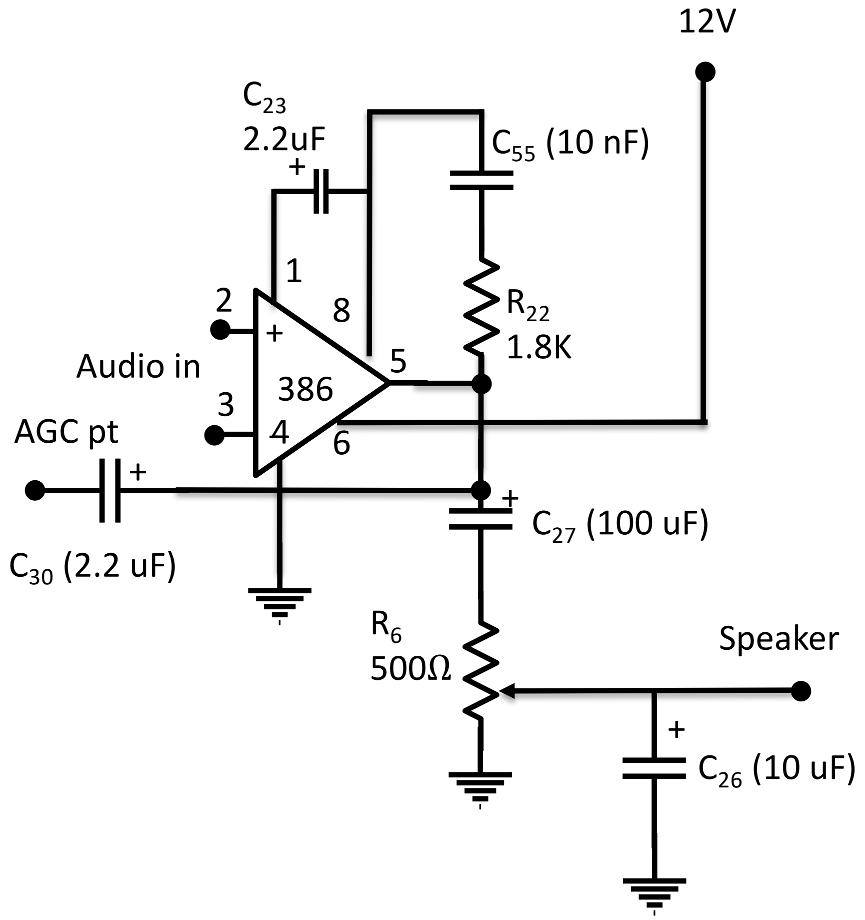
Noninverting amp

- Ideal op amp
  - $Z_{in} = \infty$
  - $A_V = \frac{V_o}{V_{i,+} - V_{i,-}}$, A_V is an op amp parameter between 10^4 and 10^6.
  - $V_- = V_+$ for negative feedback
- Example 1: Inverting amp, we'll show the gain is $\frac{R_2}{R_1}$
  - $i_1 = \frac{V_{in}}{R_1}$, $i_2 = \frac{V_{out}}{R_2}$ since the op amp has infinite input impedance
  - By Kirchhoff, $i_1 = -i_2$, so $V_{out} = \frac{R_2}{R_1} V_{in}$
  - $Z_{in} = R_1$
  - Z_{out} is same as non-inverting.
- Example 2: Non-inverting amp
  - $V_- = V_+$, so $V_{out} = (1 + \frac{R_2}{R_1}) V_{in}$
  - $Z_{in} > 10^6 \Omega$
  - $Z_{out} = \frac{R_{o,Th}}{1 + A_V \beta}$, $\beta = \frac{R_2}{R_1 + R_2}$, $R_{o,Th}$ is the Thevenin resistance of the op amp

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 $G = 2\frac{15}{.15} = 200$.

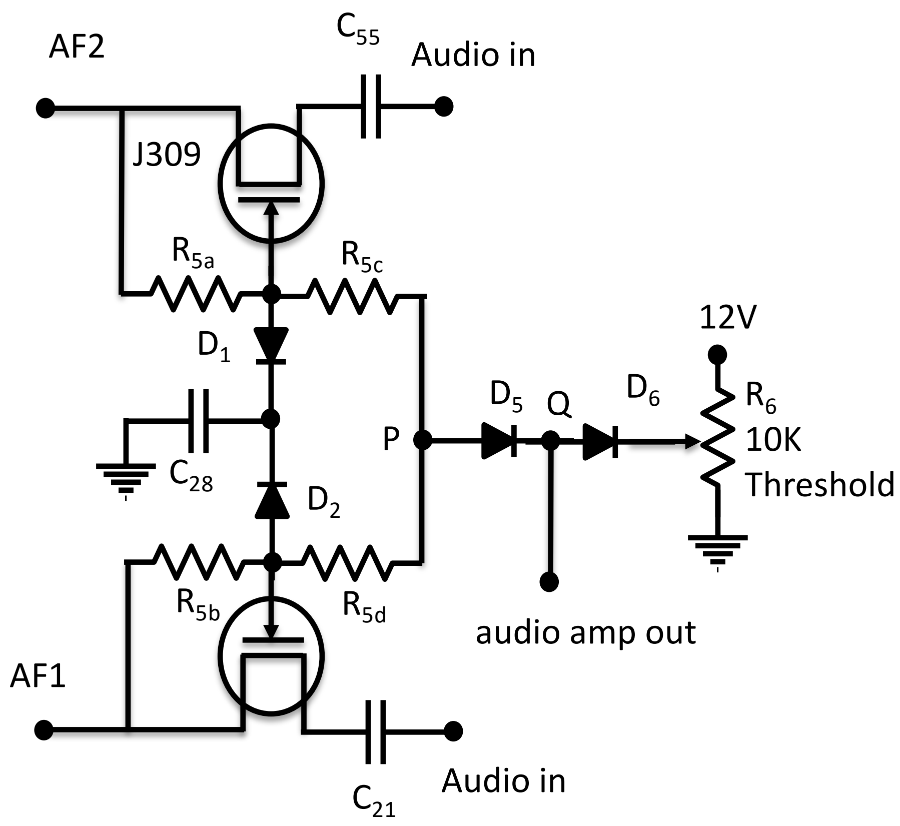


Exercise 32: Norcal AGC

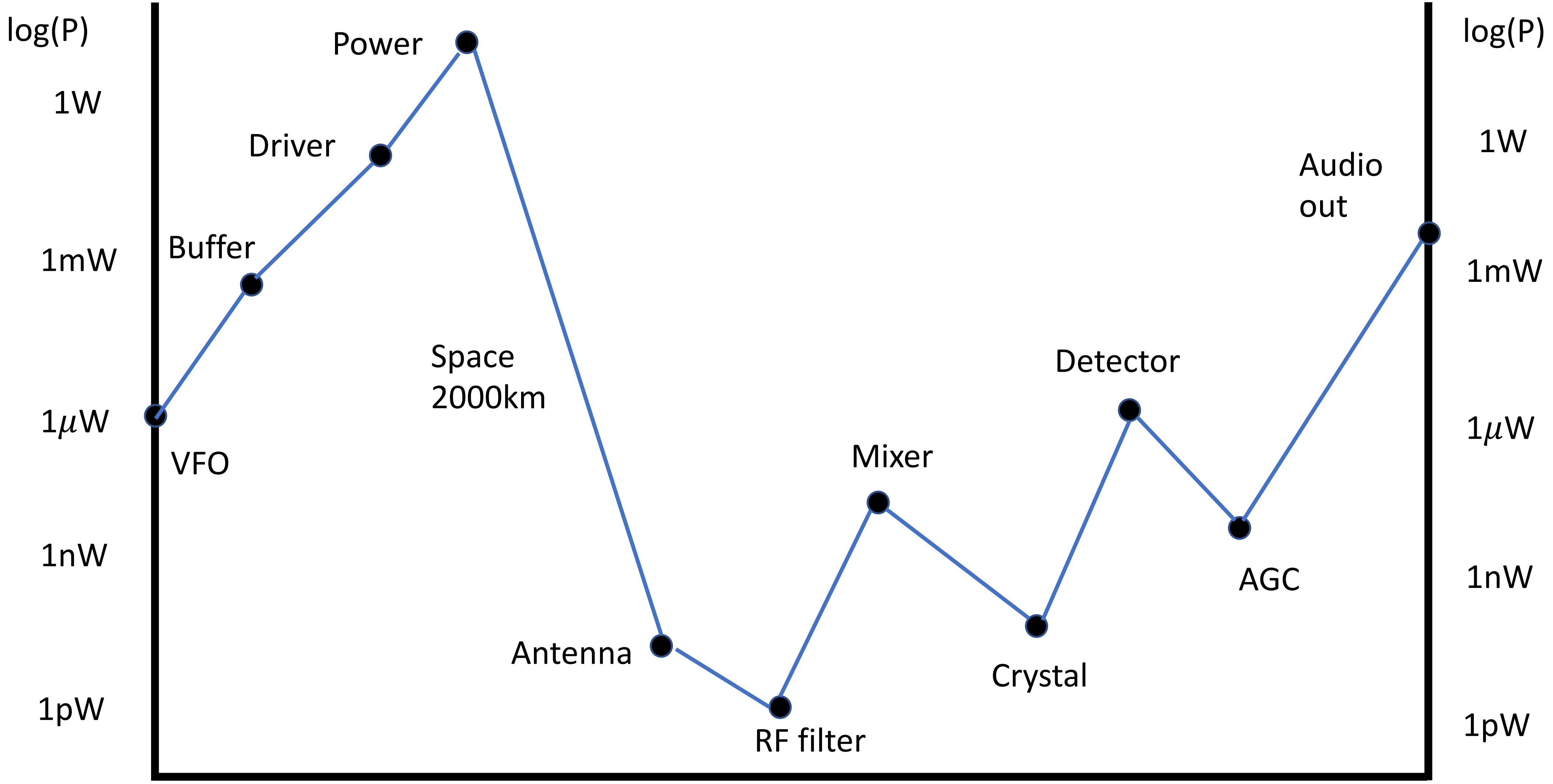
- Connect function generator through 300K resistor to AF2 (620Hz sine, R_6 fully counterclockwise) and oscilloscope to audio output. Adjust input so output is 1V rms. Connect multimeter to P.

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

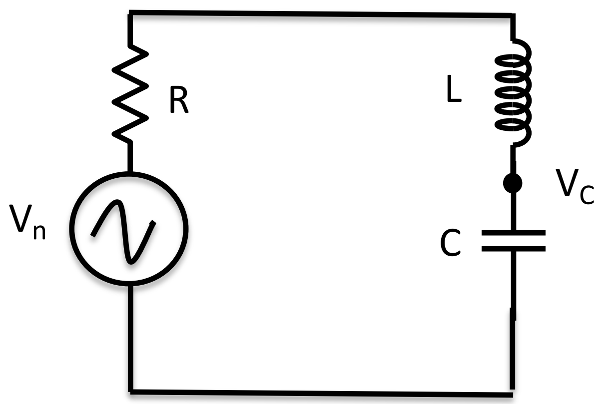


NorCal power levels



Noise

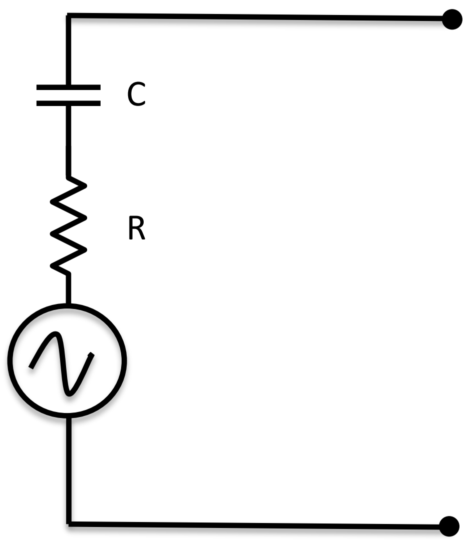
- $V_{n(rms)} = \sqrt{\frac{1}{\tau}\int_0^{\tau} V(t)^2\ dt}$
- $P_n = \frac{{V_{n(rms)}}^2}{R}$, R is load resistance
- $SNR = \frac{P}{P_n}$
- $MDS = \frac{P_n}{G}$
- $P_n = NB$, N is noise power density, B is bandwidth
- $NEP = \frac{N}{G}$



- Nyquist
  - $V_C = \frac{1}{j\omega C}\frac{V_n}{R+j\omega L+\frac{1}{j\omega C}}$
  - $\overline{|V_c|^2} = \frac{\overline{|V_n|^2}}{|1-\omega^2 LC+j\omega RC|^2}$
  - Expected energy at resonance is $kT = \frac{C}{2}\int_0^{\infty}|V_c|^2 df$, by equipartition theorem
  - $\int_0^{\infty}\frac{1}{|1-\omega^2 LC+j\omega RC|^2}df = \frac{1}{4RC}$
  - So, $\overline{|V_n|^2} = 8\mathrm{kTR}$
  - $N = kT = \frac{|\frac{V_n}{2R}|^2}{2R}$
  - $T_c = \frac{N}{k}, T_n = \frac{NEP}{k}, V_{rms} = \sqrt{4kTR}$

Antennas

- From Maxwell, for a plane wave (E in x direction, H in y direction), wave is of form $\exp(j\omega t - j\beta z)$
- $\nabla \times E = -j\mu_0 \omega H$
- $\nabla \times B = j\epsilon_0 \omega E$
- $\beta \hat{z} \times E = \mu_0 \omega H, \beta E_x \hat{y} = \mu_0 \omega H$
- Substituting and taking the restricted cross products, we get: $\beta E_x = \omega \mu_0 \frac{\omega \epsilon_0}{\beta}$, so $\beta = \omega \sqrt{\mu_0 \epsilon_0}$
- Power density: $S = Re\left(\frac{E_x \overline{H_y}}{2}\right) = \frac{(|E_x|)^2}{2\eta_0}$
- $\eta_0 = \frac{E_x}{H_y} = \sqrt{\frac{\mu_0}{\epsilon_0}} = 377\Omega$
- Impedance: $P_t = \frac{R|I|^2}{2}$, R is real part of Z, $R = R_r + R_l, \eta = \frac{R_r}{R}$
- Power density for isotropic antenna: $S_i = \frac{P_t}{4\pi r^2}$
- Define $G(\theta, \phi) = \frac{S(\theta,\phi)}{S_r}$. $S(\theta, \phi)$ is just the Poynting vector



Receiving antenna Thevenin

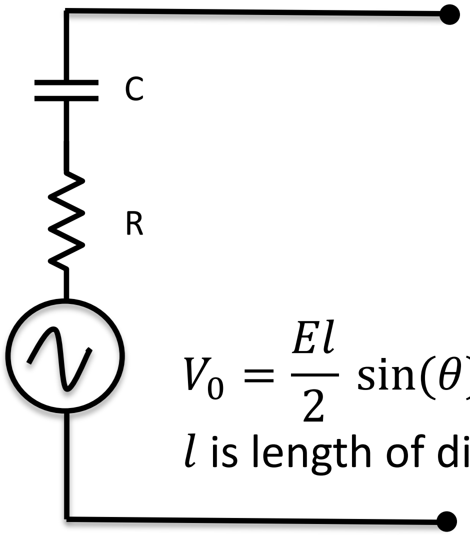
Transmitting Antenna

- Define $G(\theta, \phi) = \frac{S(\theta,\phi)}{S_r}$. $S(\theta, \phi)$ is just the Poynting vector
- For isotropic reference: $S_i = \frac{P_t}{4\pi r^2}$, $G = \frac{4\pi r^2 S}{P_t}$
- $\int G \, d\Omega = 4\pi$

Receiving Antenna

- $V_0 = hE$, h is effective antenna length ($h = \frac{l}{2}$ for short antenna)
- For dipole: $V_0 = \frac{l}{2}\, E \sin(\theta)$
- $A(\theta, \phi) = \frac{P_r}{S(\theta,\phi)}$. This is the definition of the effective area, A.
- By reciprocity, $A(\theta, \phi) = \frac{\lambda^2}{4\pi} G(\theta, \phi)$
- $P_r = \frac{|V_0|^2}{8R_a} = \frac{|hE|^2}{8R_a}$, so
- $P_r = \frac{h^2 S \eta_0}{4R}$
- $A = \frac{h^2 \eta_0}{4R}$

Dipole Thevenin equivalent circuit



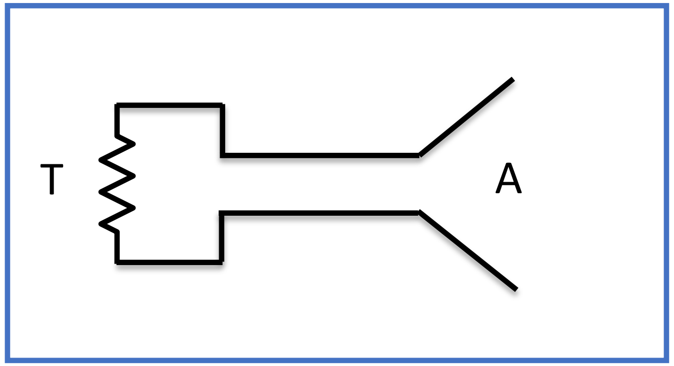
$$V_0 = \frac{El}{2} \sin(\theta)$$

l is length of dipole

Friis, blackbody and Antenna Theorem

- For transmitting/receiving antenna pairs: $G_1 A_2 = \frac{|V|^2 \pi r^2}{|I|^2 R_1 R_2} = G_2 A_1$. So $\frac{G_1}{A_1} = \frac{G_2}{A_2} = \frac{4\pi}{\lambda^2}$
- $S = \frac{P_t G}{4\pi r^2}$
- $P_r = SA = \frac{P_t G A}{4\pi r^2}$. --- Friis radiation formula
- For us, $G = 1, A = 150\ m^2, r = 2000km, P_t = 2W$
- $P_r = 6pW$
- Antenna theorem: $\oint A\, d\Omega = \lambda^2$
- For cavity on right, T is constant at thermodynamic equilibrium and the same power is transmitted and emitted, the Johnson noise is kT. The energy received is
  - $E = \frac{4\pi kT}{c\lambda^2}$.
  - Set $B = \frac{kT}{\lambda^2}$.
  - $kT = \oint BA\, d\Omega = \oint A \frac{kT}{\lambda^2}\, d\Omega$, which gives the antenna theorem

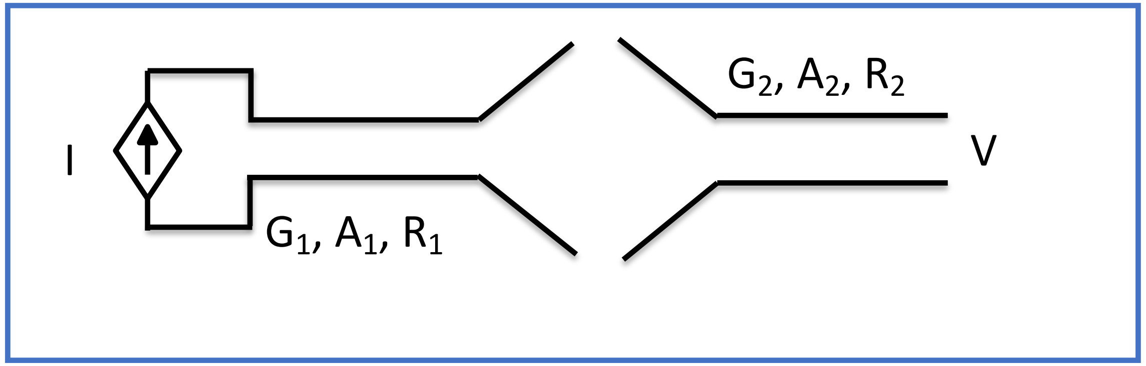
Insulated cavity



Reciprocity

- *Reciprocity:* The position of an ideal voltmeter and ideal current source can be interchanged without changing the voltmeter reading.

- $\frac{G}{A} = \frac{4\pi}{\lambda^2}$
- $\frac{G_1}{A_1} = \frac{G_2}{A_2}$



Reciprocity and dipoles

- For dipole (Length: $l = \frac{\lambda}{2}$)
- $\lambda^2 = \int A \, d\Omega = \int \frac{h^2 \eta_0}{4R_r} \, d\Omega$, so
- $R_r = \frac{l^2 \eta_0}{16\lambda^2} \int sin^2(\theta) d\Omega = \eta_0 \frac{\pi}{6} \, (\frac{l}{\lambda})^2$
- $A = \frac{3\lambda^2}{8\pi} \, sin^2(\theta)$ and $G = 1.5 \, sin^2(\theta)$. Note we used $h = \frac{l}{2} \sin(\theta)$
- $\frac{|V|^2}{8R_2} = \frac{|I|^2 R_1 G_1 A_2}{8\pi r^2}, G_1 A_2 = G_2 A_1$
- $P_t = \frac{|I|^2 R_1}{4\pi r^2}, P_t = \frac{|V|^2}{8R_2}$
- $P_r = \frac{P_t G_1 A_2}{4\pi r^2}$

Exercise 35: Intermodulation

- Only $f_{3\uparrow} = 2f_1 - f_2, f_{3\downarrow} = 2f_2 - f_1, f_{5\uparrow} = 3f_1 - 2f_2\ and\ f_{5\downarrow} = 3f_2 + 2f_1$ are close enough to the rf frequency to matter for intermodulation

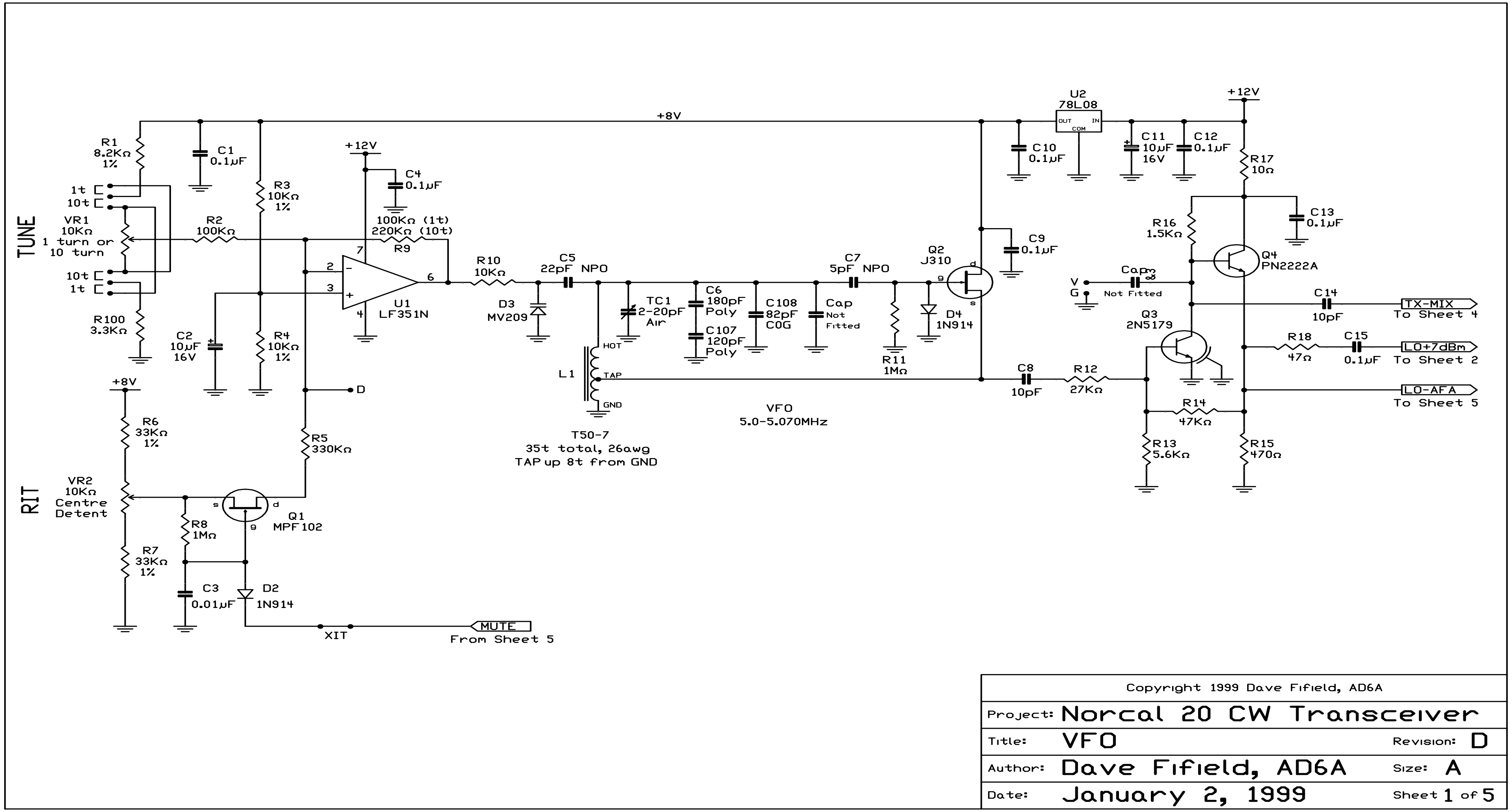
1. Find coefficients and frequencies for $[\cos(\omega_1 t) + [\cos(\omega_2 t)]^5$
2. Find $f_{3\uparrow}$, $f_{3\downarrow}$, $f_{5\uparrow}$ and f_1
3. Find the MDS and the antenna limited MDR

- $MDS = \frac{P_n}{G}$, MDI is input that gives output tone + $2P_n$
- $MDR = MDI - MDS$
- This needs measurements to measure the minimum detectable signal.

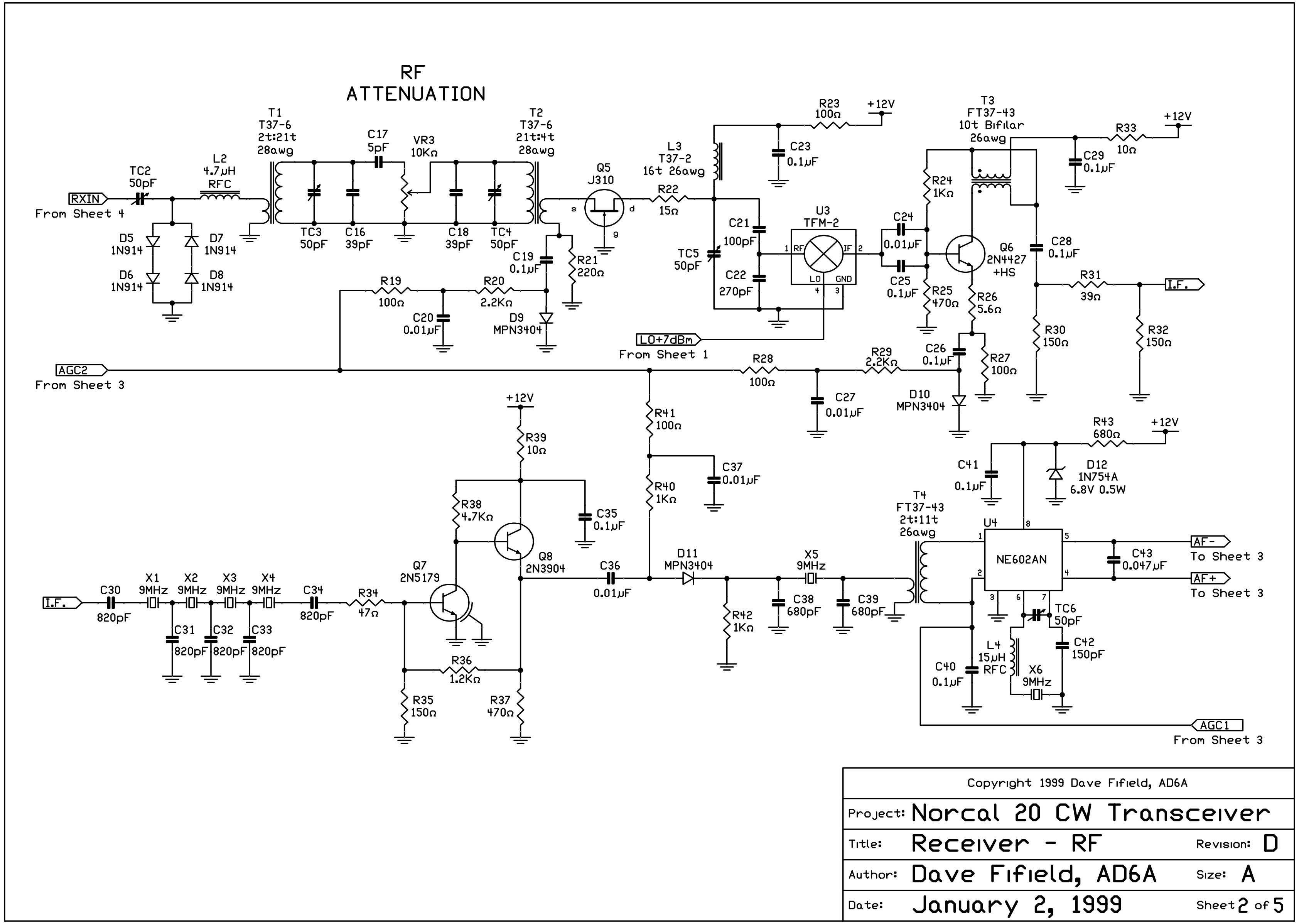
Exercise 37: Antennas

1. Use the relation between gain and effective area to rewrite the Friis transmission formula in terms of gain only. Consider UHF for airplanes. If the frequency makes the quarter length stub antenna have gain 2, find the maximum possible LOS at 10km height. Required receiver power is –90 dBm. Find the minimum transmission power.
   - Friis radiation formula is $P_r = \frac{P_t G A}{4\pi r^2}$, $\lambda = \frac{c}{f}$ and $\frac{G}{A} = \frac{4\pi}{\lambda^2}$, so $P_r = P_t(\frac{Gc}{4\pi r f})^2$
   - P_r requires $-90dBm$ power. $-90 = 10\log\left(\frac{P_r}{1mW}\right)$, so P_r must be at least $10^{-12}\ W$.
   - $P_t(f) = (\frac{4\pi r f}{Gc})^2 \times 10^{-12} W$. $P_t(10^8) = 4.4mW$, $P_t(2\times10^8) = 19.24mW$, $P_t(3\times10^8) = 40mW$ and $P_t(10^9) = 440mW$.
2. Find the inductance to resonate with a 3m whip. Assuming the Q of the coil is 200, find the turns ratio required to give a transceiver a 50 ohm load. What is the radiation efficiency?
   - For whip, $R_r = 160\pi^2(\frac{l}{\lambda})^2 \approx 7.8\Omega$. $Q = 200 = \frac{L}{7.8}$. $L = 1551\Omega$.
   - $Z_s = 7.8 + 1551j$. $L = 3.6\mu H$
   - $\frac{1551}{50} = \frac{Z_p}{Z_s} = (\frac{N_p}{N_s})^2$, $\frac{N_p}{N_s} = \sqrt{31} = 5.2$.
   - Radiation efficiency is $\frac{P_{radiated}}{P_{input}} = \frac{R_r}{R_r + R_L} = \frac{7.8}{57.8} \approx .14$

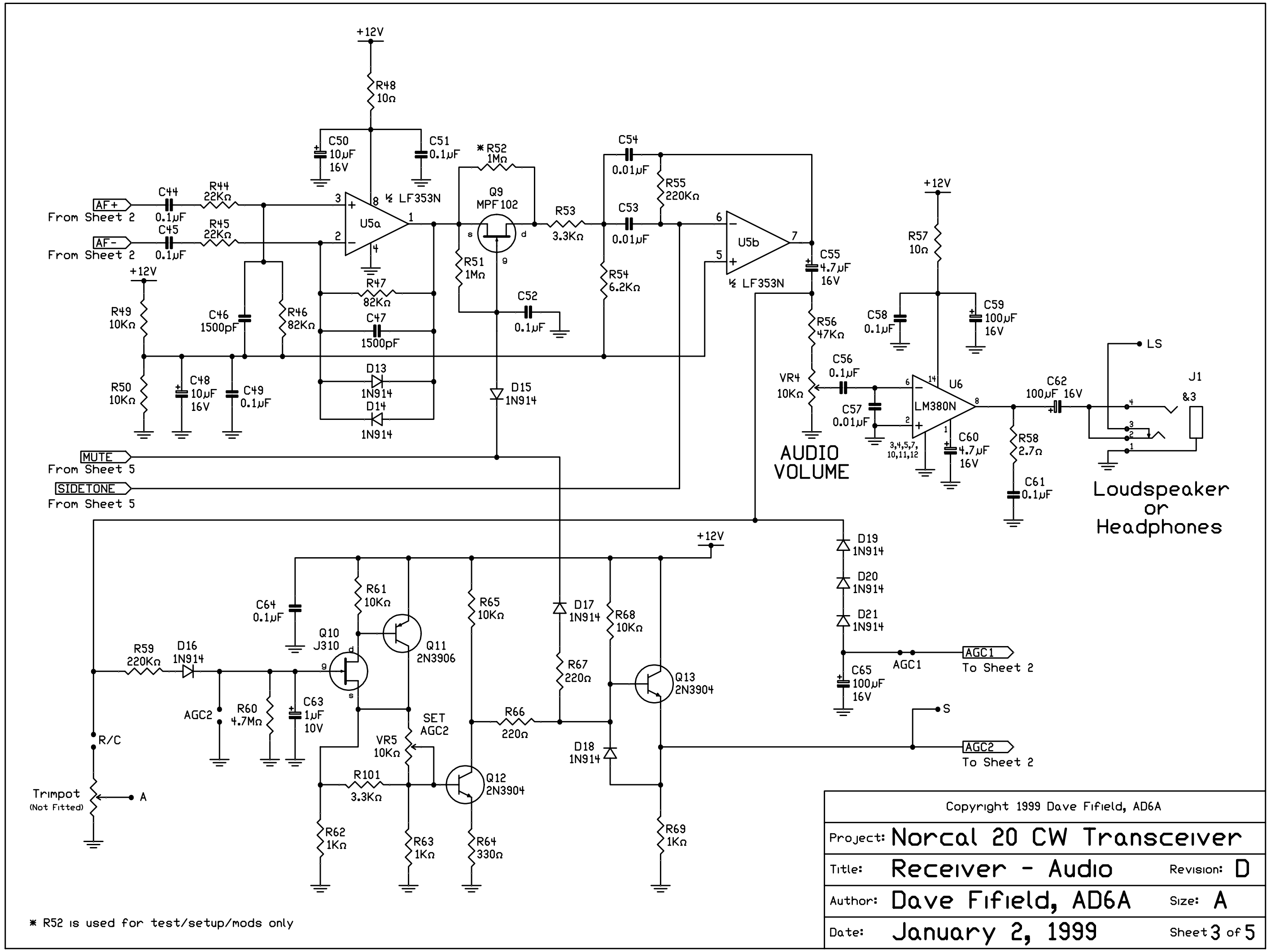
Norcal circuit diagram, 1



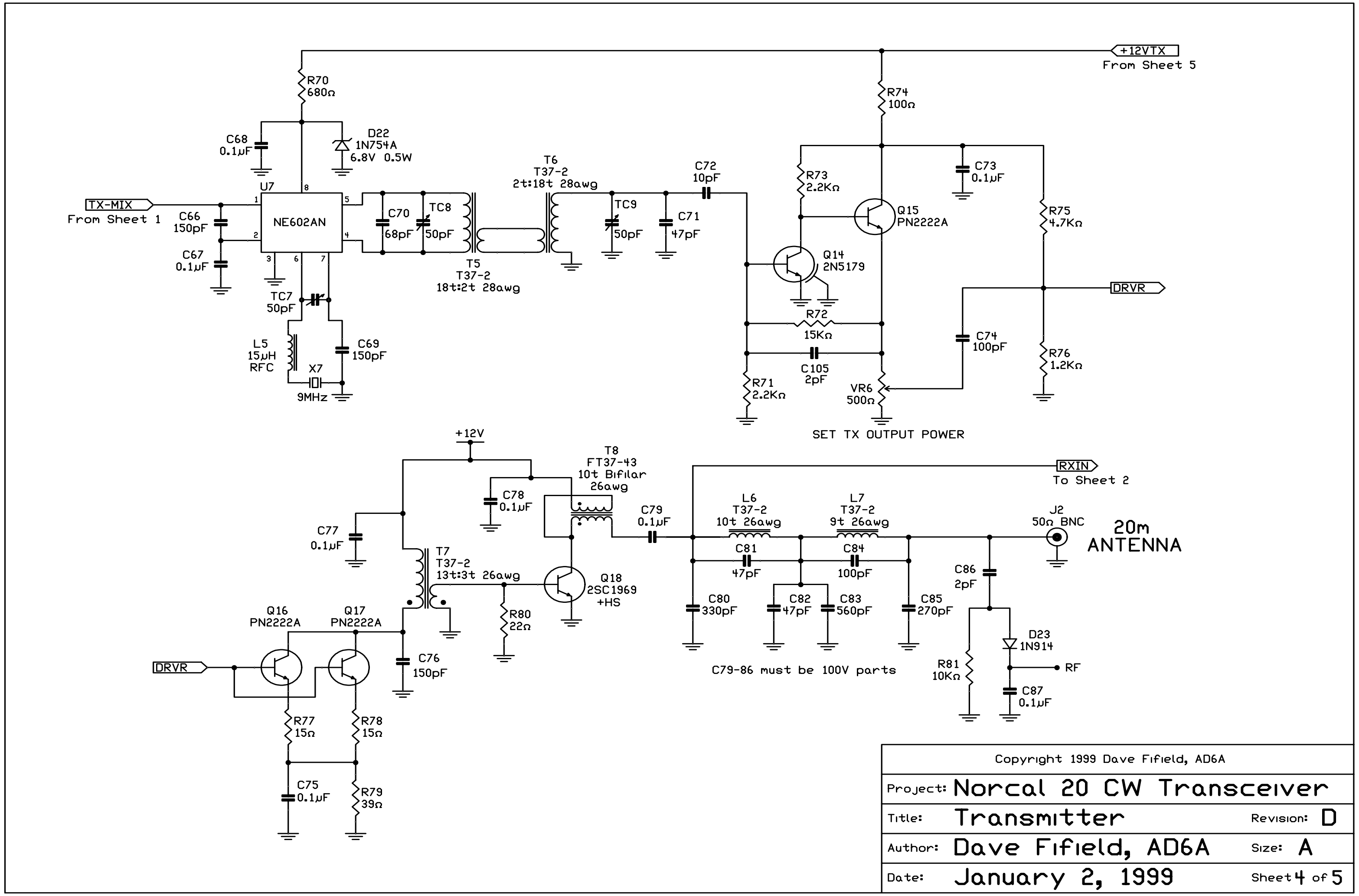
Norcal circuit diagram, 2



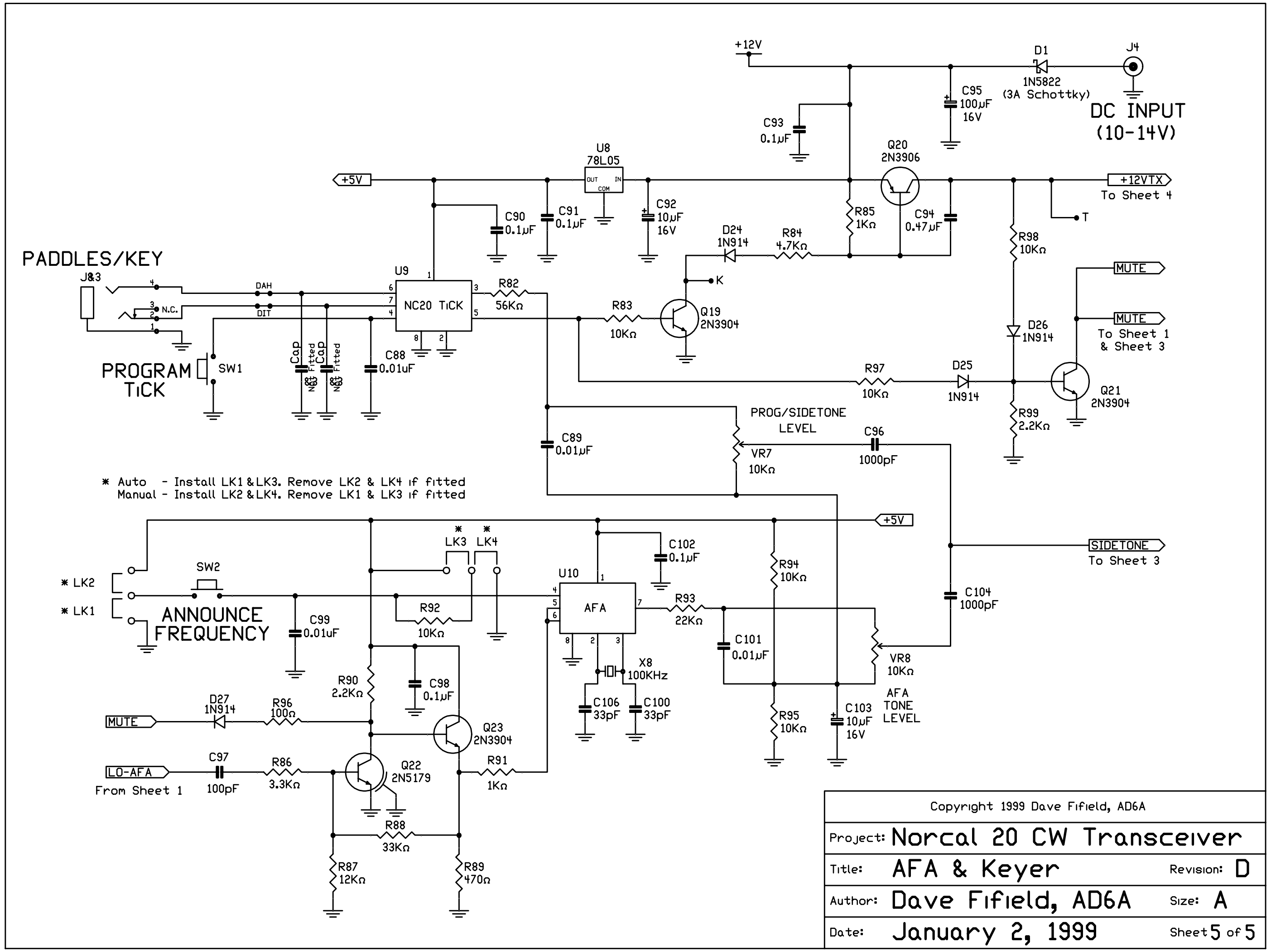
Norcal circuit diagram, 3



Norcal circuit diagram, 4



Norcal circuit diagram, 5



Appendix

Exercise 17: Tuned Speaker

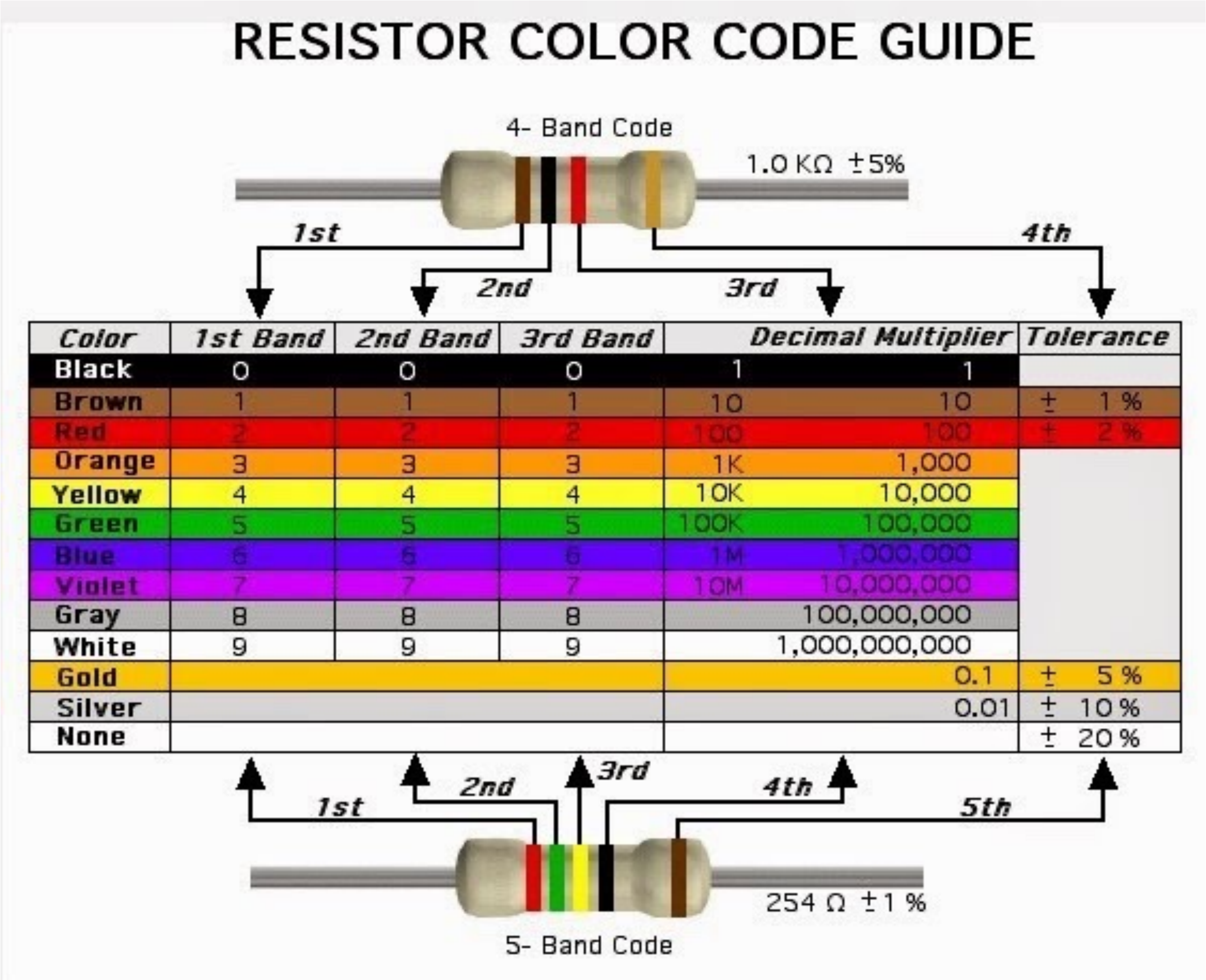
- Connect speaker to function generator 600Hz, 25mVrms.

1. Sound peaks at resonance. Find resonant frequency L_p.
2. Measure f_l, f_u by noting the 3dB loss. Calculate Q.
3. Use voltmeter to find resonance with speaker (nominally 8ohm) to calculate impedance
4. Calculate the resonant frequency from a transmission line equivalent circuit.

Morse

Symbol	Code	Symbol	Code	Symbol	Code
a	._	m	__	y	_.__
b	_...	n	_.	z	__..
c	_._.	o	___	0	_____
d	_..	p	.__.	1	.____
e	.	q	__._	2	..___
f	.._.	r	-_.	3	...__
g	__.	s	...	4	...._
h	....	t	_	5	.....
i	..	u	.._	6	_....
j	.___	v	..._	7	__...
k	_._	w	.__	8	___..
l	._..	x	_.._	9	____.

Color codes



Color	1st Band	2nd Band	3rd Band	Decimal Multiplier		Tolerance
Black	0	0	0	1	1	
Brown	1	1	1	10	10	± 1 %
Red	2	2	2	100	100	± 2 %
Orange	3	3	3	1K	1,000	
Yellow	4	4	4	10K	10,000	
Green	5	5	5	100K	100,000	
Blue	6	6	6	1M	1,000,000	
Violet	7	7	7	10M	10,000,000	
Gray	8	8	8		100,000,000	
White	9	9	9		1,000,000,000	
Gold					0.1	± 5 %
Silver					0.01	± 10 %
None						± 20 %

- Resistors markings in ohms
- Capacitors markings in picoFarads
- Inductors markings in microHenries

Component data

Core Size	26	3	15	1	2	6	10	12/17	0
T-12-( )	*	60	50	48	20	17	12	7.5	2.4
T-16-( )	145	61	55	44	22	19	13	8	3
T-20-( )	185	76	65	52	25	22	16	10	3.5
T-25-( )	245	100	85	70	34	27	19	12	4.5
T-30-( )	335	140	93	85	43	36	25	16	6
T-37-( )	285	120	90	80	40	30	25	15	4.9
T-44-( )	370	180	160	105	52	42	33	18.5	6.5
T-50-( )	330	175	135	100	49	40	31	18	6.4
T-68-( )	435	195	180	115	57	47	32	21	7.5
T-80-( )	460	180	170	115	55	45	32	22	8.5
T-94-( )	600	248	200	160	84	70	58	*	10.6
T-106-( )	930	450	345	325	135	116	*	*	19
T-130-( )	810	350	250	200	110	96	*	*	15
T-157-( )	1000	420	*	320	140	115	*	*	*
T-184-( )	1690	720	*	500	240	195	*	*	*
T-200-( )	920	425	*	250	120	100	*	*	*
T-200A-( )	1600	*	*	*	218	*	*	*	*

IRON POWDER TOROIDS - A_l Values **

* size not available in this material

** L= µH/100 turns

In the beginning…

- The laws of EM according to Clerk Maxwell are:
  1. $\nabla \cdot \boldsymbol{E} = \frac{\rho}{\epsilon_0}$
  2. $\nabla \cdot \boldsymbol{B} = 0$
  3. $\nabla \times \boldsymbol{E} = -\frac{\partial \boldsymbol{B}}{\partial t}$
  4. $c^2\, \nabla \times \boldsymbol{B} = \frac{\boldsymbol{j}}{\epsilon_0} + \frac{\partial \boldsymbol{E}}{\partial t}$, $\epsilon_0 = 8.854 \times 10^{-12}\ \frac{C^2}{N-m^2}$, $\frac{1}{c^2} = \epsilon_0 \mu_0$
  5. $\nabla \cdot \boldsymbol{j} = -\frac{\partial \rho}{\partial t}$
     - Here **E** is the electric field, **B** is the magnetic field, **j** is the current density through a closed surface, c is the speed of light and ρ is the charge density at a point.
     - In non-dispersive matter, $B = \mu H = \mu_0(H + M), \mu = \mu_0(1 + \chi_m)$,
     - $D = \epsilon E = \epsilon_0 E + P, \epsilon = \kappa \epsilon_0$ and (1) becomes $\nabla \cdot \boldsymbol{D} = \rho_f$, (4) becomes $\nabla \times \boldsymbol{H} = \boldsymbol{j_f} + \frac{\partial \boldsymbol{D}}{\partial t}$

- The rest of classical physics, including special relativity, is:
  - Newton-Einstein: $\boldsymbol{p} = m\boldsymbol{v}$, $m = \frac{m_0}{\sqrt{1-(\frac{v}{c})^2}}$, $\boldsymbol{F} = m\frac{d\boldsymbol{p}}{dt}$.
  - Gravity: $\boldsymbol{F} = -\frac{Gm_1 m_2}{r^2}\ \boldsymbol{u}_r$ where $\mathbf{u_r}$ is the unit vector from m_1 to m_2 and **F** is the force on m_2.

Solutions to the wave equation

- The solution of $\nabla^2\psi - \frac{1}{c^2}\frac{\partial^2 \psi}{\partial t^2} = -s$ is $\psi(x,y,z,t) = \frac{S(t-\frac{r}{c})}{4\pi r}$ where $S = \int_V \; s\, dV$
- Later, we will use this to find the ”general” solution to Maxwell’s equations
  - $\phi(r_1,t) = \int_{V_2} \frac{\rho(r_2,t-\frac{|r_1-r_2|}{c})}{4\pi\epsilon_0|r_1-r_2|} dV_2$ and $\mathbf{A}(r_1,t) = \int_{V_2} \frac{\mathbf{j}(r_2,\; t-\frac{|r_1-r_2|}{c})}{4\pi\epsilon_0 c^2|r_1-r_2|} dV_2$, where
  - $B = \nabla\times A$, $E = -\nabla\phi - \frac{\partial A}{\partial t}$, and $c^2\nabla\cdot A = -\frac{\partial\phi}{\partial t}$
- You are not expected to have guessed this answer
- To do this, we’ll need the “BAC-CAB” identity: $A\times(B\times C) = B(A\cdot C) - C(A\cdot B)$
- When we apply this to $\nabla\times(\nabla\times\mathbf{A})$, we get $\nabla\times(\nabla\times\mathbf{A}) = \nabla(\nabla\cdot A) - \nabla^2 A$

General solution to Maxwell's equations

- Returning to the general Maxwell equations, from $\nabla \cdot B = 0$, we get $B = \nabla \times A$
- Substituting into $c^2 \nabla \times \mathbf{B} = \frac{\boldsymbol{j}}{\epsilon_0} + \frac{\partial \mathbf{E}}{\partial t}$, we get $c^2 \nabla \times (\nabla \times A) = \frac{j}{\epsilon_0} + \frac{\partial E}{\partial t}$
- Applying "BAC-CAB", we get $\nabla (\nabla \cdot A) - \nabla^2 A = \frac{j}{c^2 \epsilon_0} + \frac{1}{c^2} \frac{\partial E}{\partial t}$ (Equation 1)
- Now, $\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t}$, so substituting for B, we get $\nabla \times \left(\mathrm{E} + \frac{\partial A}{\partial t} \right) = 0$ and so $E = -\nabla \phi - \frac{\partial A}{\partial t}$
- Substituting into equation 1, $\nabla (\nabla \cdot A) - \nabla^2 A = \frac{j}{c^2 \epsilon_0} + \frac{1}{c^2} \frac{\partial}{\partial t} (-\nabla \phi - \frac{\partial A}{\partial t})$, or
- $\nabla (\nabla \cdot A) - \nabla^2 A = \frac{j}{c^2 \epsilon_0} - \frac{1}{c^2} \nabla \frac{\partial \phi}{\partial t} - \frac{1}{c^2} \frac{\partial^2 A}{\partial t^2}$.
- $\nabla^2 A - \frac{1}{c^2} \frac{\partial^2 A}{\partial t^2} = -\frac{j}{c^2 \epsilon_0} + \nabla [\frac{1}{c^2} \frac{\partial \phi}{\partial t} + (\nabla \cdot A)]$
- Now if A and ϕ give $B = \nabla \times A$ and $E = -\nabla \phi - \frac{\partial A}{\partial t}$, then $A' = A + \nabla \varphi$ and $\phi' = \phi - \frac{\partial \varphi}{\partial t}$ give $B = \nabla \times A'$ and $E = -\nabla \phi' - \frac{\partial A'}{\partial t}$, for any function φ

General solution to Maxwell's equations

- Thus, we can pick a solution $(A, \phi)$with $\nabla \cdot A = -\frac{1}{c^2}\frac{\partial \phi}{\partial t}$. Then we get
- $\nabla^2 A$ - $\frac{1}{c^2}\frac{\partial^2 A}{\partial t^2} = -\frac{j}{c^2 \epsilon_0}$
- Substituting for $E = -\nabla\phi - \frac{\partial A}{\partial t}$ into $\nabla \cdot \mathrm{E} = \frac{\rho}{\epsilon_0}$, we get
- $\nabla \cdot (\nabla\phi) + \frac{\partial\, \nabla \cdot A}{\partial t} = -\frac{\rho}{\epsilon_0}$, or $\nabla \cdot (\nabla\phi) - \frac{1}{c^2}\frac{\partial^2 \phi}{\partial t^2} = -\frac{\rho}{\epsilon_0}$, or $\nabla^2\phi$ - $\frac{1}{c^2}\frac{\partial^2 \phi}{\partial t^2} = -\frac{\rho}{\epsilon_0}$
- The solutions are $\phi(r_1, t) = \int_{V_2} \frac{\rho(r_2, t - \frac{|r_1 - r_2|}{c})}{4\pi\epsilon_0 |r_1 - r_2|} dV_2$ and
- $\mathbf{A}(r_1, t) = \int_{V_2} \frac{\mathbf{j}(r_2,\ t - \frac{|r_1 - r_2|}{c})}{4\pi\epsilon_0 c^2 |r_1 - r_2|} dV_2$ with $\nabla \cdot A = -\frac{1}{c^2}\frac{\partial \phi}{\partial t}$, with $\nabla \cdot A = -\frac{1}{c^2}\frac{\partial \phi}{\partial t}$.

Solution to Maxwell's equations in free space

- Free space is defined by $\rho = 0$ and $j = 0$, so our potentials satisfy
- $\nabla^2 A$ - $\frac{1}{c^2}\frac{\partial^2 A}{\partial t^2}$ =0 and $\nabla^2 \phi$ - $\frac{1}{c^2}\frac{\partial^2 \phi}{\partial t^2} = 0$
- These have the usual wave equation solutions $\phi(x, y, z, t) = \ f(k \cdot r \ - \ \omega t)$, etc
- Thus, in free space ϕ and **A** and hence E and B propagate as waves.

Solution to Maxwell's equations in conductors

- In conductors, $\mathbf{j} = \sigma\mathbf{E}$
  - $c^2\,\nabla\times\mathbf{B} = \frac{\boldsymbol{j}}{\epsilon} + \frac{\partial\mathbf{E}}{\partial t} = \frac{\sigma}{\epsilon}\mathbf{E} + \frac{\partial\mathbf{E}}{\partial t}$
  - This becomes $c^2\,\frac{\partial(\nabla\times\mathbf{B})}{\partial t} = \frac{\sigma}{\epsilon}\frac{\partial\mathbf{E}}{\partial t} + \frac{\partial^2\mathbf{E}}{\partial t^2}$
  - $\nabla\times\mathbf{E} = -\frac{\partial\mathbf{B}}{\partial t}$, so we get $c^2\nabla\times\frac{\partial B}{\partial t} = -c^2\nabla\times(\nabla\times E) = \frac{\sigma}{\epsilon}\frac{\partial\mathbf{E}}{\partial t} + \frac{\partial^2\mathbf{E}}{\partial t^2} = -c^2[\nabla(\boldsymbol{\nabla}\cdot\mathbf{E}) - \nabla^2\mathbf{E}] = \boldsymbol{c^2\nabla^2}\mathbf{E}$ (since $\rho = 0$ in a conductor)
  - Applying the trial solution E= E_0 exp(ωt-kr), we get $-k^2$-i$\omega\mu\sigma$ + $\omega^2\mu\epsilon$ = 0.
  - Putting k = α-βi, $\alpha = \frac{\omega}{2}\sqrt{\mu\epsilon}\,(1 + \sqrt{1 + \frac{\sigma^2}{\omega^2\epsilon^2}})$ and $\beta = \frac{\omega\mu\sigma}{2\alpha}$.
  - For copper, σ = 5.78 × 107 $\boldsymbol{\Omega}$-m. This explains the "skin effect" in conductors.

Radiation, antennas

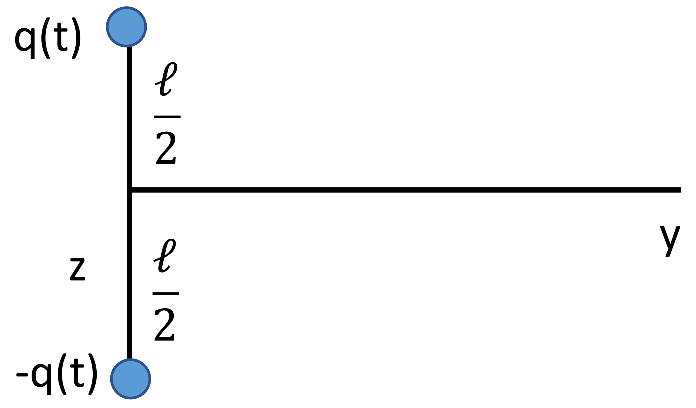
- Accelerating charges radiate energy in the form of electromagnetic waves (companion E and B fields).
- The radiation from accelerating charge q is $\mathbf{E}_{rad} = -\frac{1}{4\pi\epsilon_0 c^2}\frac{q}{r}\,\boldsymbol{a}_{\perp}(t - \frac{r_{12}}{c})$.
- Here, $\mathbf{a}_{\perp}$ is the acceleration $\perp$ to the line from r_1 to r_2.
- For example, applying a time varying potential $V_0\sin(\omega t)$ to an antenna will cause the antenna to radiate power since the voltage and hence charges affected accelerate within the antenna, that is, their positions have a non-zero second derivative. That's how a transmitter "couples" to the antenna of a receiver. In the receiver, the radiated wave accelerates charges in the antenna replicating the original wave (at much reduced power).
- These simple radio waves are carrier waves of frequency $\frac{\omega}{2\pi}$. To transfer information (voice, images, binary data), we modulate carrier waves combining them with an "information source" signal. Receivers demodulate the incoming wave and recreate the original "information source" signal.

Maxwell's equations in a non-dispersive media

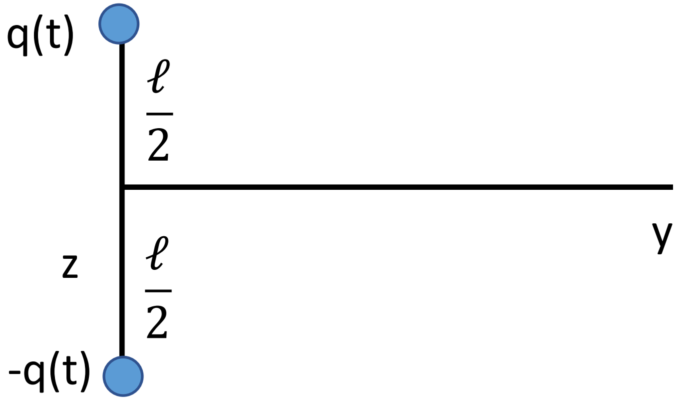
- $\mathrm{B} = \mu\mathrm{H}, \mathrm{D} = \epsilon\mathrm{E}$
- $\nabla \cdot D = \rho$
- $\nabla \cdot B = 0$
- $\nabla \times \mathrm{E} = -\frac{\partial \mathrm{B}}{\partial \mathrm{t}}$
- $\nabla \times \mathrm{H} = \mathrm{j} + \frac{\partial \mathrm{E}}{\partial \mathrm{t}}$
- $\nabla \cdot j = -\frac{\partial \rho}{\partial t}$

Radiation from a small dipole

- $A_z(r,t) = \frac{\mu_0}{4\pi} \int_{[\frac{-l}{2},\frac{l}{2}]} \frac{I(z',t-\frac{z'}{c}k)}{|r-z'k|}\, dz'$
- If $l \ll cT = \lambda$
- $A_z(r,t) = \frac{\mu_0}{4\pi} \frac{l}{r} I(z', t - \frac{r}{c})$
- Choosing gauge, $\nabla \cdot \mathrm{A} + \frac{1}{c^2} \frac{\partial \varphi}{\partial t} = 0$
- $\frac{\partial \varphi}{\partial t} = -\frac{l}{4\pi\epsilon_0} \frac{\partial}{\partial z} (\frac{1}{r} I\left(t - \frac{r}{c}\right)) = \frac{z}{r^2} (\frac{q(t-\frac{r}{c})}{r} - \frac{I((t-\frac{r}{c})}{c})$
- $q\left(t - \frac{r}{c}\right) = q_0 \cos\left(\omega\left[t - \frac{r}{c}\right]\right),\ I\left(t - \frac{r}{c}\right) = I_0 \sin(\omega\left[t - \frac{r}{c}\right])$



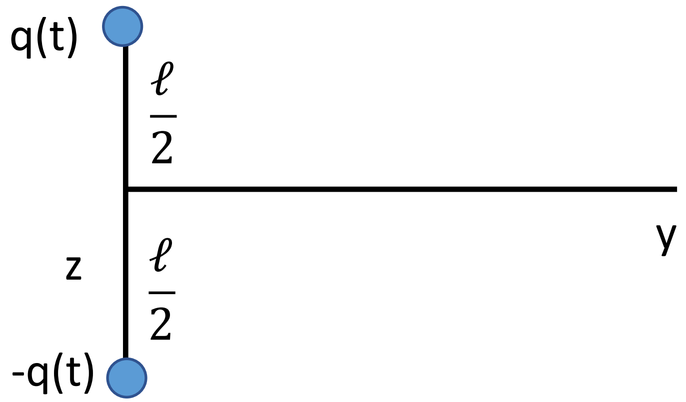
Radiation from a small dipole



- $\nabla^2 H - \epsilon\mu \frac{\partial^2 H}{\partial t^2} - \sigma\mu \frac{\partial H}{\partial t} = 0$
- $\nabla^2 E - \epsilon\mu \frac{\partial^2 E}{\partial t^2} - \sigma\mu \frac{\partial E}{\partial t} = 0$
- $A_r = \frac{\mu_0}{4\pi} \frac{I_0 l}{r} \cos(\theta) \sin(\omega \left[t - \frac{r}{c}\right])$
- $A_\phi = 0, \; A_\theta = -\frac{\mu_0}{4\pi} \frac{I_0 l}{r} \cos(\theta) \sin(\omega \left[t - \frac{r}{c}\right])$
- $B_\phi = \frac{1}{r} \frac{\partial}{\partial r}(rA_\theta) - \frac{1}{r} \frac{\partial A_r}{\partial \theta} = \frac{\mu_0}{4\pi} \frac{I_0 l}{r} \sin(\theta)[\frac{\omega}{r} \cos\left(\omega \left[t - \frac{r}{c}\right]\right) + \frac{1}{r} \sin(\omega[t - \frac{r}{c}])]$
- $E_r = -\frac{\partial \phi}{\partial t} - \frac{\partial A_r}{\partial t} = \frac{2lI_0 \cos(\theta)}{4\pi\epsilon_0} [\frac{\sin(\omega\left[t - \frac{r}{c}\right])}{r^2 c} - \frac{\cos(\omega\left[t - \frac{r}{c}\right])}{\omega r^3}]$
- $E_\theta = \frac{-I_0 \, \mathrm{l}\sin(\theta)}{4\pi\epsilon_0} (\left[\frac{1}{r^3 \omega} - \frac{\omega}{rc^2}\right] \cos\left(\omega \left[t - \frac{r}{c}\right]\right) - \frac{1}{cr^2} \sin\left(\omega \left[t - \frac{r}{c}\right]\right))$
- $E_\phi = -\frac{1}{r sin(\theta)} \frac{\partial \varphi}{\partial \phi} - \frac{\partial A_\phi}{\partial t} = 0$

Radiation from a small dipole

- $\nabla \cdot (\mathbf{E} \times \mathbf{H}) = -\mathbf{H}\frac{\partial \mathbf{B}}{\partial t} - \mathbf{E}\frac{\partial \mathbf{D}}{\partial t} - \mathbf{E} \cdot \mathbf{j}, \mathbf{S} = \mathbf{E} \times \boldsymbol{H}, \nabla \cdot \mathbf{S} + \frac{\partial u}{\partial t} = -\mathbf{E} \cdot \mathbf{j}$ (u is energy density)
- $\int S \cdot dA = \frac{(l\, I_0\, \omega)^2}{6\pi\epsilon_0 c^3} \cos(\omega \left[t - \frac{r}{c}\right])^2$
- $P_{av} = \frac{(l\omega)^2}{6\pi\epsilon_0 c^3}\frac{{I_0}^2}{2} = \frac{2\pi}{3}\sqrt{\frac{\mu_0}{\epsilon_0}}\,(\frac{l}{\lambda})^2\,\frac{{I_0}^2}{2}$
- $R_r = \frac{2\pi}{3}\sqrt{\frac{\mu_0}{\epsilon_0}}\,(\frac{l}{\lambda})^2$



Large half wave dipole

- For large half wave, add small dipoles to produce half wave antenna.
- $dE_\theta = I_0 \frac{\sin(\theta)}{4\pi\epsilon_0 Rc^2}\ \omega \cos(\omega) \cos\left(\frac{2\pi z\prime}{\lambda}\right)\ dz'$
- $dB_\phi = I_0 \frac{\mu_0 \omega}{4\pi Rc}\ \omega \cos\left(\omega \left[\left[t - \frac{r}{c}\right]\right]\right) \cos\left(\frac{2\pi z\prime}{\lambda}\right)\ dz'$
- $K = \int_{[-\frac{\pi}{2},\frac{\pi}{2}]} \frac{1}{R} \cos\left(t - \frac{R}{c}\right) \cos(u)\, du = \frac{1}{2\pi\epsilon_0 rc} \cos\left(\omega \left[t - \frac{r}{c}\right]\right) \frac{\cos\left(\frac{\pi}{2}\cos(\theta)\right)}{sin^2(\theta)}, u = \frac{2\pi z\prime}{\lambda}$
- $E_\theta = I_0 \frac{1}{2\pi\epsilon_0 rc} \cos\left(\omega \left[t - \frac{r}{c}\right]\right) \frac{\cos\left(\frac{\pi}{2}\cos(\theta)\right)}{sin\,(\theta)}$
- $B_\phi = I_0 \frac{\mu_0}{2\pi r}\ \omega \cos\left(\omega \left[\left[t - \frac{r}{c}\right]\right]\right) \frac{\cos\left(\frac{\pi}{2}\cos(\theta)\right)}{sin\,(\theta)}$
- $P_{av} = \frac{1}{4\pi}\sqrt{\frac{\mu_0}{\epsilon_0}}\ {I_0}^2 \int_{[0,\pi]} \frac{cos^2(\frac{\pi}{2}\cos(\theta))}{sin^2(\theta)} \sin(\theta) d\theta = 73.1\Omega\ \frac{{I_0}^2}{2}$

Radiation from an accelerating charge

- $\mathrm{r}' + \mathrm{R} = \mathrm{r}$, R= |r-r'|
- $\varphi(r,t) = \frac{1}{4\pi\epsilon_0}\int_{V_1}\frac{\rho(r',t-\frac{R}{c})}{|r-r'|}dv' = \frac{1}{4\pi\epsilon_0}\left[\frac{Q}{r} + \frac{r\cdot p(t-\frac{r}{c})}{r^3} + \frac{r\cdot\frac{dp}{dt}(t-\frac{r}{c})}{cr^2}\right]$
- $\boldsymbol{A}(r,t) = \frac{\mu_0}{4\pi}\int_{V_1}\frac{\boldsymbol{j}(r',t-\frac{R}{c})}{|r-r'|}dv' = \frac{\mu_0}{4\pi r}\frac{d}{dt}\boldsymbol{p}(t-\frac{r}{c})$
- $\boldsymbol{E} = -\frac{\partial\boldsymbol{A}}{\partial t} - \nabla\varphi$
- $\boldsymbol{B}(r,t) = \frac{-\mu_0}{4\pi c r^2}\boldsymbol{r}\times\frac{d^2}{dt^2}\boldsymbol{p}(t-\frac{r}{c})$
- $\boldsymbol{E}(r,t) = -\frac{c}{r}\boldsymbol{r}\times\boldsymbol{B}(r,t)$
- $\frac{d\boldsymbol{p}}{dt} = q\frac{dr\prime}{dt} = qv,\ \frac{d^2}{dt^2}\boldsymbol{p}(t-\frac{r}{c}) = q\frac{dv}{dt}$
- $P_R = -\frac{dW}{dt} = \frac{2}{3}\frac{q^2}{4\pi\epsilon_0 c^3}(\frac{dv}{dt})^2$

Radiation from a single accelerating charge

- Near zone
  - $\varphi(r,t) = \frac{q}{4\pi\epsilon_0}\left[\frac{1}{R\prime(1+\frac{v\cdot n\prime}{c})}\right]$
  - $R^* = R' - \frac{v}{c}(x_0 - x'_1)$
  - $E(r,t) = \frac{1}{4\pi\epsilon_0}\frac{1}{R^{*3}}[\left(R' - \frac{R'\,v\prime}{c}\right)\left(1 - \frac{v^2}{c^2}\right) + \frac{R\prime}{c^2}\left(R' - \frac{R'v\prime}{c}\right)\times\frac{dv\prime}{dt}]$
  - $B = \frac{R' \times E}{R'c}$
  - $S = \frac{q^2}{16\pi^2 c^3\epsilon_0}\,\frac{R\prime(R'\times v')^2}{(R')^5}$

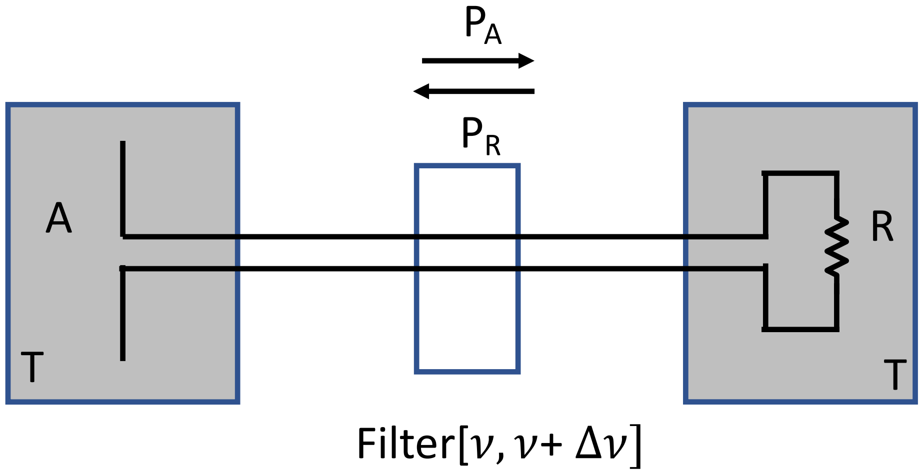
Radiation loss and antenna aperture

- Spreading loss: $L_s = 32 + \log(\mathrm{d}) + 20\log(\mathrm{f})$
  - d in kilometers
  - F in megahertz
- $W = A_e P_e, A_e = \frac{\lambda^2}{4\pi}$

Antenna aperture

- Deriving antenna aperture uses thermodynamic argument: black body equilibrium

- $P_A = \frac{A_e}{2} B_\upsilon \Delta\nu \int_{[0,4\pi]} d\Omega = 2\pi\, A_e B_\upsilon \Delta\nu$
- $B_\upsilon = \frac{2\nu^2 kT}{c^2} = \frac{2kT}{\lambda^2}$ (Rayleigh-Jeans)
- $P_R = kT\Delta\nu$
- $P_A = P_R$
- $A_e = \frac{\lambda^2}{4\pi}$



Impedance matching

- L networks, π and T networks
- Impedance and reflection coefficients

Blank

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