Pre-select filter

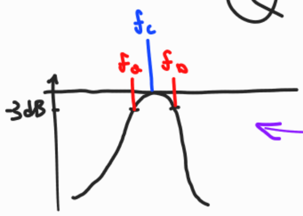
- Bandpass filter
- Allows only band of interest to pass → FM band 88 MHz − 108 MHz

What Q should the filter have?

$$Q = \frac{f_c}{f_b - f_a} = \frac{98\ \text{MHz}}{108\ \text{MHz} - 88\ \text{MHz}} = \frac{98}{20} = 4.9$$

center freq

−3dB bandwidth

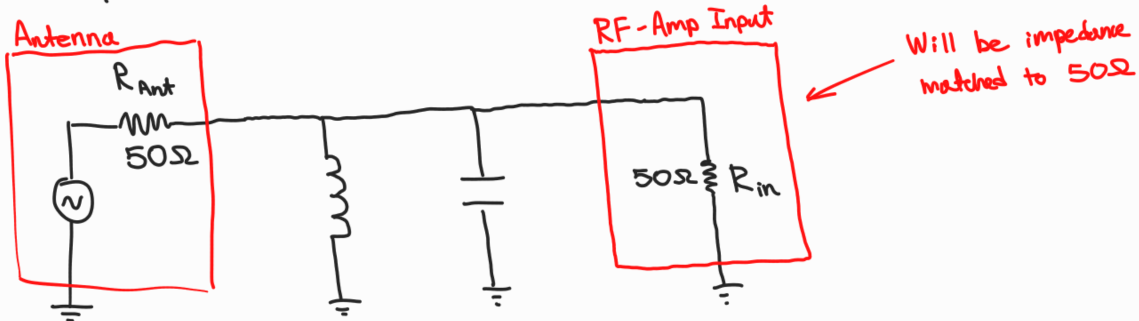


We are making a parallel bandpass filter, so the formula for the reactances @ resonance is

$$X_o = \frac{R_p}{Q}$$

← parallel resistance

Our circuitry looks like



Video Series
Ep 1, Pt 2
~10:30

The total parallel impedance of this circuit is $50 \parallel 50 = 25\ \Omega$

(My explanation for this making sense is that the "additional", source, series resistance increases the damping of the filter, and hence reduces $Q \propto R_p$)

$$X_0 = \frac{25}{4.9}$$

$$X_0 = \omega L$$
$$= 2\pi f_c \cdot L$$
$$\frac{X_0}{2\pi f_c} = L$$

$$8.28587 \times 10^{-9}\ H = L$$
$$\underline{8.28587\ nH = L}$$

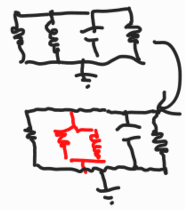
$$X_0 = \frac{1}{\omega C}$$
$$= \frac{1}{2\pi f_c \cdot C}$$
$$C = \frac{1}{2\pi f_c \cdot X_0}$$

@ $f_c = 98$ MHz

$$C = 3.1831 \times 10^{-10}\ H$$
$$\underline{C = 318.31\ pF}$$

this capacitance value will be used for impedance transformation

Also, these components have parasitics, most notably the inductor's equiv. series resistance (ESR)



Ep 1
pt 2
14:21

This clearly lowers Q $\left(Q = \frac{R_p}{X_0}\right)$, and increases insertion loss (lowers the equivalent load resistance).

I chose an inductor with a Q of 32, meaning that...
@ 250MHz

$$X_0 = \frac{25}{4.9}$$
@ 98MHz

$$X_0 = \frac{25}{4.9} \cdot \frac{250}{98}$$
$$= 13.0154$$
@ 250 MHz

$$Q_{inductor} = \frac{R_{P\,ind}}{X_0}$$

$$Q \cdot X_0 = R_{P\,ind}$$

$$32 \cdot 13.0154 = R_{P\,ind} = 416.5\ \Omega$$

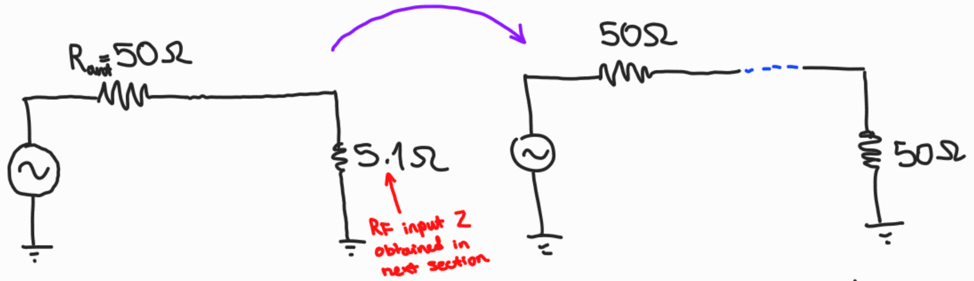
So the equivalent load resistance is slightly smaller than 50Ω, as $50 \,//\, 163.265 = 44.64\Omega$

I won't worry about this effect.

From the video series, ≤0603 caps are recommended for low ESL.

At frequencies $> f_c$, we want the filter to bypass/short all signals to ground. Because of ESL, at some point, the capacitor component's impedance becomes dominated by the ESL's reactance, increasing the impedance, and degrading the filter.

Since the amplifier input impedance will need to be transformed to a higher value, the parallel capacitance produced from this transformation can be used in the filter.



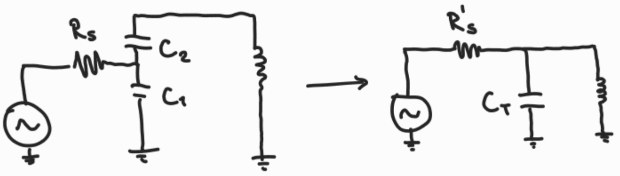
RF input Z obtained in next section

Why don't we want to bring the 50Ω antenna impedance down to 5.1Ω?

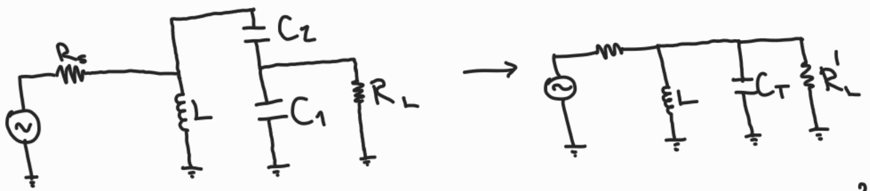
- Hard to design filters with good (high) Q with low load (and/or source) resistance
- Won't be able to use a 50Ω transmission line. ← more complicated than I expect... don't worry about it here.
  - loss of gain and reflections due to a mismatch between the antenna and the transmission line may or may not be significant.
    - Better to avoid.
    - Though, I've seen radios built with general purpose RF boards (2 layers – 1 sig; 1 gnd plane) and there clearly weren't 50Ω transmission lines.
- Will power still be transfered efficiently if source is matched?

Impedance transformation

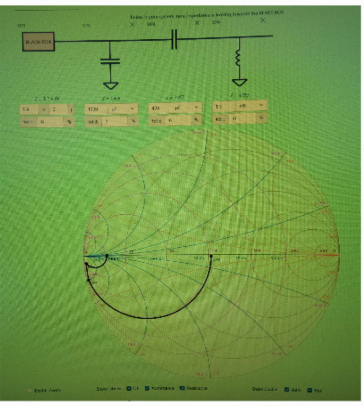
Pg. 36 Christopher Bowick



I want to increase the load resistance, so I will do the following



$$C_T = 318.31 \times 10^{-12} = \frac{C_1 C_2}{C_1 + C_2} \qquad R'_L = 50 = 5.1\left(1 + \frac{C_1}{C_2}\right)^2$$



$$C_1 = 9.96667 \times 10^{-10} \qquad C_2 = 4.67673 \times 10^{-10}$$

$$= 997\text{ pF} \qquad = 468\text{ pF}$$

These capacitances will be used (as a reference) for the filter + impedance transformation!

Practical values used were $C_1 = 1\text{nF}$, $C_2 = 470\text{pF}$, 8.1nH resulting in an impedance of $52.2 - j1.43$ and filter with $f_c = 98.8\text{ MHz}$

The amplifier input now looks like 50 Ω.

RF Amp

Common-Collector/Source Amp.

↳ Why?

- Lower input impedence → easier to match to antenna
- Medium output impedance → higher Q output filter

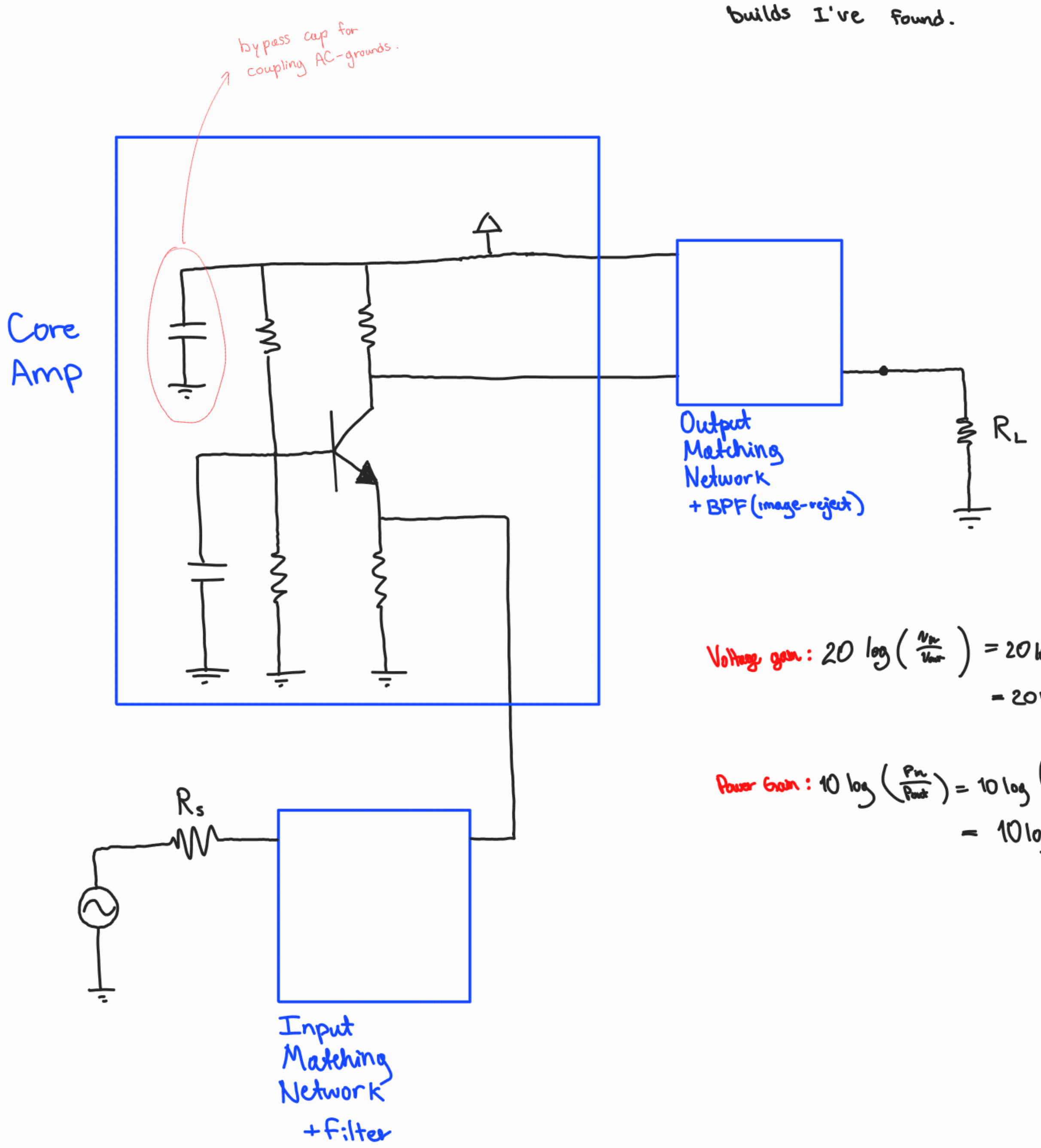
→ MOSFET vs. BJT

MOSFET

- Has gate capacitance
  - ↳ doesn't matter for common-source config?

BJT

- has r_π component
  - ↳ β varies a lot... though, if it's high enough, base current is negligible, and barely affects output
- Found in many FM receiver builds I've found.



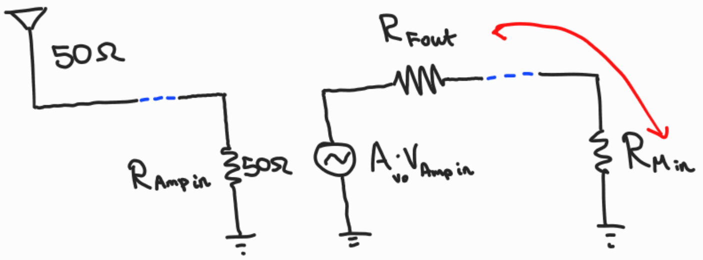
Voltage gain: $20\log\left(\frac{V_{in}}{V_{out}}\right) = 20\log\left(\frac{I R_{in}}{I R_{out}}\right) = 20\log\left(\frac{R_{in}}{R_{out}}\right)$

Power Gain: $10\log\left(\frac{P_{in}}{P_{out}}\right) = 10\log\left(\frac{I^2 R_{in}}{I^2 R_{out}}\right) = 10\log\left(\frac{R_{in}}{R_{out}}\right)$

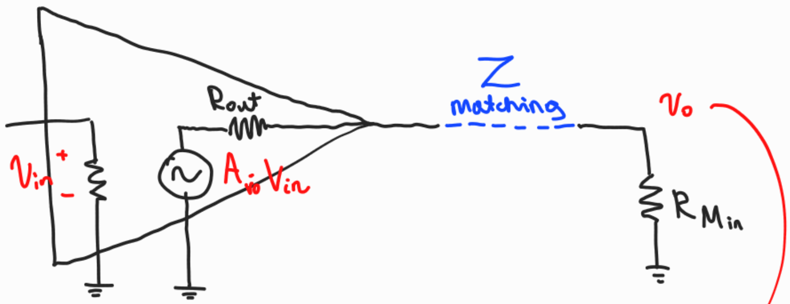
I've seen BJT amplifiers to be found more commonly. Thus I will design one.

Let's aim for +20dB of overall gain from the antenna to the mixer input, so, calculate the needed amplifier A_{vo}:

ignoring filters since they are ~infinite Z.



→ equal because of impedance-matching circuitry.



R_{Min} : mixer input resistance

$$V_o = A_{vo} \cdot V_{in} \cdot \underbrace{\frac{R_{Min}}{R_{Min} + R_{out}}}_{\frac{1}{2}}$$

$$V_o = \frac{1}{2} A_{vo} \cdot V_{in}$$

$$20\text{dB} = 20 \log\left(\frac{V_o}{V_{in}}\right)$$

$$= 20 \log\left(\frac{1}{2} \frac{A_{vo} \cdot \cancel{V_{in}}}{\cancel{V_{in}}}\right)$$

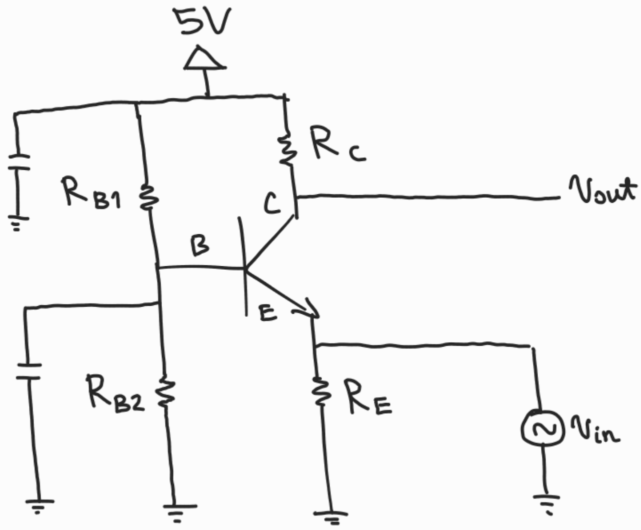
$$1 = \log\left(\frac{A_{vo}}{2}\right)$$

$$10 = \frac{A_{vo}}{2}$$

$$\underline{20 = A_{vo}}$$

The open circuit gain of the amp, A_{vo}, must be 20.

Common-Base Amplifier:



Going to choose an $I_C = 5\text{mA}$ ← quiescent / at the middle

(I've seen designs with 4mA; Ep 3 21:20 uses 5mA)

$$r_e = \frac{1}{g_m} = \frac{26\text{mV}}{I_E} \approx \frac{26\text{ mV}}{I_C} \approx \frac{26\text{ mV}}{5\text{mA}} \approx 5.2\,\Omega$$

pg. 399

$$A_{v_{oc}} = +\frac{1}{r_e}(R_c \| r_o) = 20$$

r_o can be treated as infinity (Ep 3, 40:20)

$$20 = \frac{1}{r_e} \cdot R_c$$

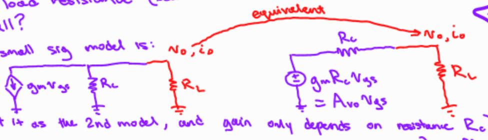
$$R_c = 104\,\Omega$$

will use $105\,\Omega$

Sources of confusion:

- Does the load resistance (because it is in parallel) affect this value at all?

  No, the small sig model is:



  (represent it as the 2nd model, and gain only depends on resistance R_c)

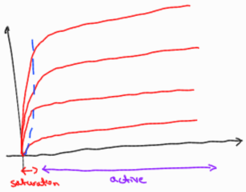
- Does the impedance matching circuitry (if matching source to load) affect what this value should be?
  - Sim showed lower gain with matching circuit + 28.5 Ohm load.
  - The load should be matched up to the source.

$$I_{C(SAT)} = 10\text{ mA}$$

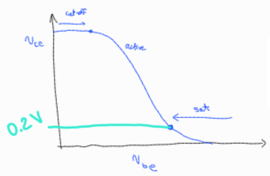
Typical $V_{CE(SAT)} = 0.2\text{V}$

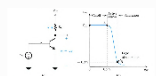
~typical value pg. 367 Sedra
4:10 CE Amp video

[Saturation region for BJT is where V_{BE} isn't ~linear with i_c]



Sedra ~pg. 326





Sedra pg. 380 pt Q

We have: $I_{C(SAT)} = 10\text{mA} = \dfrac{V_{DD} - V_{CE(SAT)}}{R_E + R_C} = \dfrac{5 - 0.2}{R_E + 104\,\Omega}$ $\quad R_E + 104 = 480$

$$R_E = 376\,\Omega$$

will use $374\,\Omega$

Next,

minimum β value of my transistor ~~"BFR360FH6327XTSA1"~~

$$R_{B2} \leqslant \frac{\beta \cdot R_E}{10}$$

$$R_{B2} \leqslant \frac{90 \cdot 376}{10} = 3384$$

Let's go with $\underline{3k\Omega}$

For R_{B1}, we want it such that it biases the amplifier to have 5mA DC current.

$$V_E = 376 \cdot 4mA = 1.88\,V$$

$$V_B = V_E + 0.7$$

$$V_B = 2.58V = V_{RB2}$$

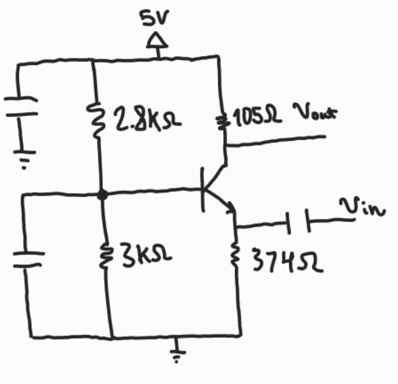
Voltage divider:

$$5 \cdot \frac{R_{B2}}{R_{B1} + R_{B2}} = 2.58V$$

$$\frac{5}{2.58} \cdot 3000 - 3000 = R_{B1} = 2813.95 \approx \underline{2.8k\Omega}$$

Now, let's put this all into the amplifier model.

Circuit:



Small sig Will derive later

derived from base current

$$R_i = \frac{1}{g_m} \| R_E \| r_\pi \longrightarrow r_\pi = (\beta + 1)\frac{V_T}{I_{EQ}}$$

typical

$$= 120 \cdot \frac{26mV}{5mA} = 624\,\Omega$$

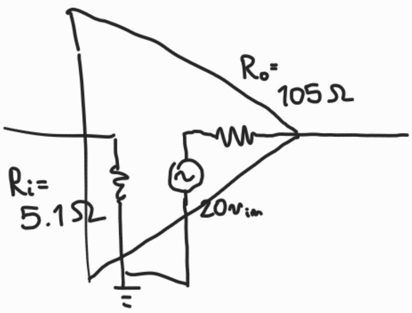
$$= \frac{1}{\frac{1}{5.2} + \frac{1}{374} + \frac{1}{624}}$$

$$= 5.08688\,\Omega$$

$$\approx 5.1\,\Omega$$

$$R_o = R_c \parallel r_o$$
$$\approx R_c = 105\,\Omega$$

Thus, we have



Thoughts:

- low input resistance is a result of "higher" I_C
- with ×20 gain, we only get output swing of 0.52 V.

  Why?

$\frac{1}{g_m} = \frac{26\text{mV}}{I_c}$

$A = 20 = g_m \cdot R_c$

$20 \cdot \frac{26\text{mV}}{I_c} = R_c$

$20 \cdot 26\text{mV} = V_c$

$0.52\text{V} = V_c$

Indep of I_c and R_c!

$V_{cc} - 0.52\text{V}$ is the middle point.

Mixer : LT5560

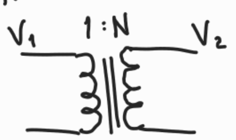
Available one with good documentation and desired input/output frequency range.

Input

The input is balanced, but our amplifier output is single-ended. Possible solutions are:

- Drive Mixer Input Single-Endedly
  - High loss

- Use a Balun (balancing unit)
  - This is a transformer, which can perform impedance transformation:



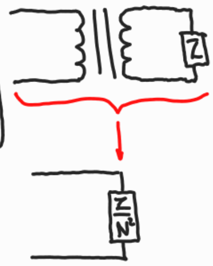
$V_2 = NV_1$

$V_1 I_1 = V_2 I_2$

$V_1 I_1 = N V_1 I_2$

$\frac{I_1}{N} = I_2$

Replace transformer with an impedance from the source end?



$Z = \frac{V_2}{I_2} = \frac{NV_1}{\left(\frac{I_1}{N}\right)} = N^2 \frac{V_1}{I_1}$

$\frac{V_1}{I_1} = \frac{Z}{N^2}$

- Use a 1:1 input transformer (balun), and match (simple)

From datasheet,

Table 2. Input Signal Port Differential Impedance

FREQUENCY (MHz)	INPUT IMPEDANCE (Ω)	REFLECTION COEFFICIENT (Z_0 = 50Ω)	
		MAG	ANGLE (DEG.)
70	28.5 + j0.8	0.274	177
140	28.5 + j1.6	0.274	174
240	28.6 + j2.7	0.275	171

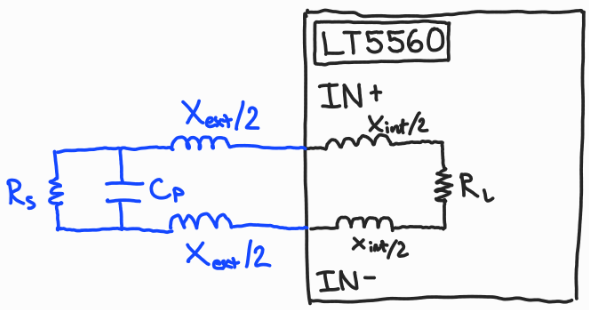
@ 70MHz $\mathrm{Im}(Z_{in}) = 0.8$

@ 140 MHz $\mathrm{Im}(Z_{in}) = 1.6$

Linear interpolation:

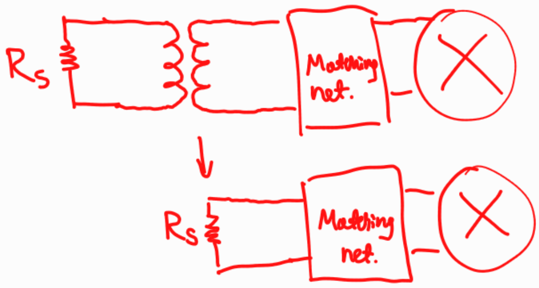
$\frac{1.6-0.8}{140-70} \cdot (98-70) + 0.8 = 1.12$

$\therefore Z = R_L + X_{int}$
$= 28.5 + j1.12$
@ 98 MHz



What model are these values from?
How do you match DIFFERENTIAL IMPEDANCES??

Firstly, because of the 1:1 turns ratio...



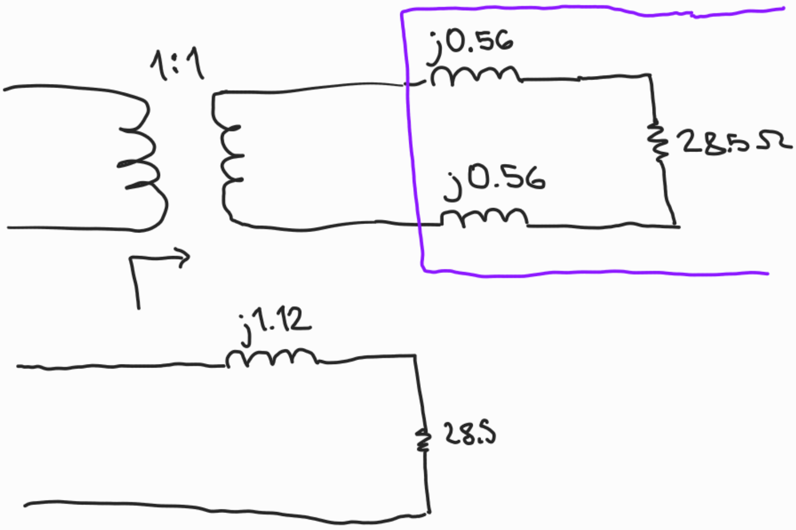
Secondly, putting our scenario in the single-ended model (where the matching equations are derived from)

Rearranging the circuit in the following way doesn't affect the voltage across / current through R_L at all:

OK so, we match the diff. impedances based off the single-ended circuit representation.

Note: the X_{int} (internal reactance) is used as part of the matching circuit.

Now, let's match the mixer input impedance up to 105 Ω

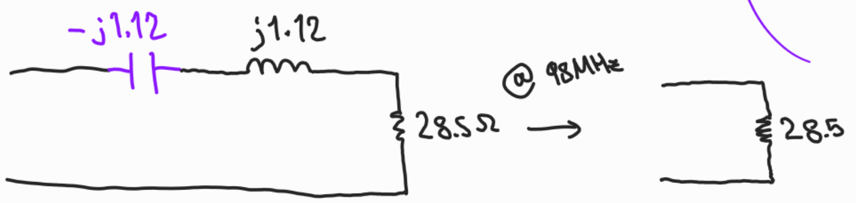


Q: Should we care about reflections/ losses due to impedance mismatch between transmission line and mixer input / non-50 Ohm matching circuitry?

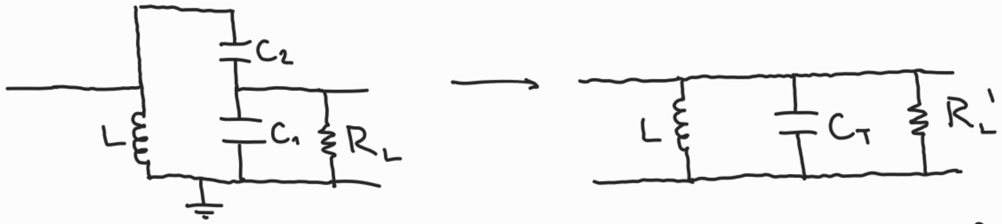
Let's resonate the inductance with a capacitance

$$C = \frac{1}{2\pi f \cdot 1.12} = 1.45\,nF$$

(inductance was: $L = \frac{1.12}{2\pi f} = 1.819\,pH$)



Now, transform the 28.5 Ohms



$$C_T = \frac{C_1 C_2}{C_1 + C_2} \qquad R_L' = R_L\left(1 + \frac{C_1}{C_2}\right)^2$$

To choose a C_T, find the desired bandpass filter at the output

$$Q = \frac{98\,MHz}{20\,MHz} = 4.9 \qquad Q = \frac{R_p}{X_o} \rightarrow X_o = \frac{R_p}{Q} = \frac{52.5}{4.9} = \frac{75}{7}$$

$$X_o = \frac{1}{2\pi f C} \qquad X_o = 2\pi f L$$

$$C = \frac{1}{2\pi f X_o} \qquad L = \frac{X_o}{2\pi f}$$

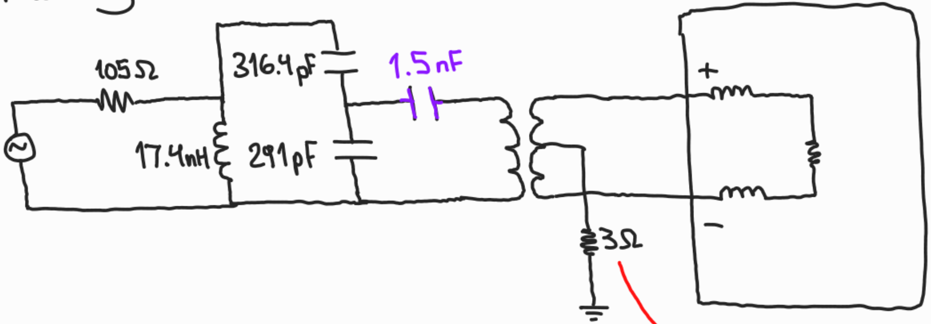
$$C = 151.576\,pF \qquad L = 17.4\,nH$$

gives us:

$$C_1 = 290.939\,pF$$

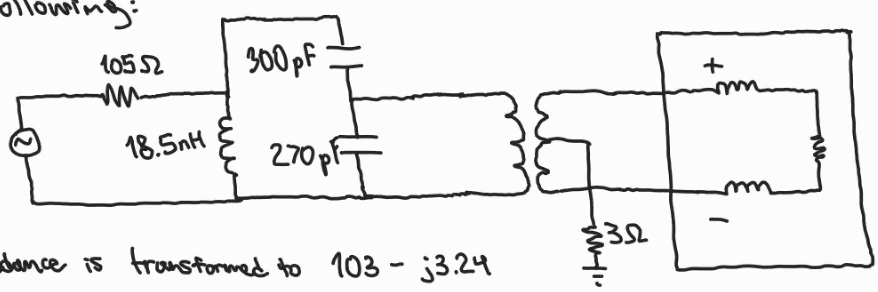
$$C_2 = 316.435\,pF$$

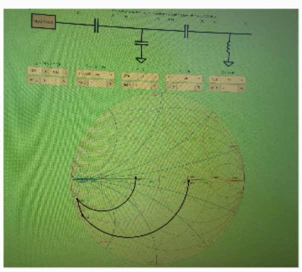
The ideal matching network is thus as follows:



current limiting (datasheet pg. 15)

However, with practical component values, using a Smith chart tool, I came up with the following:

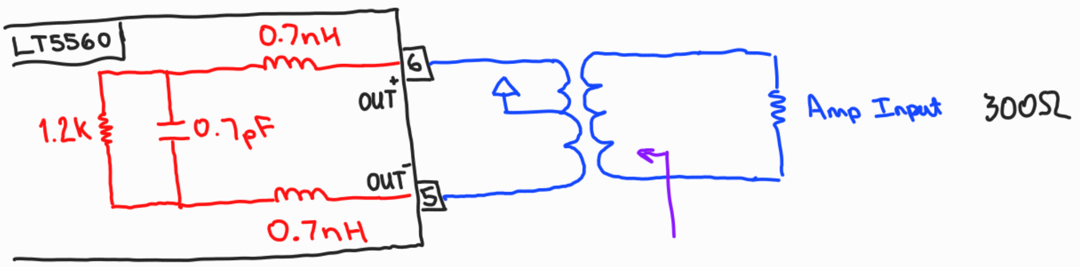


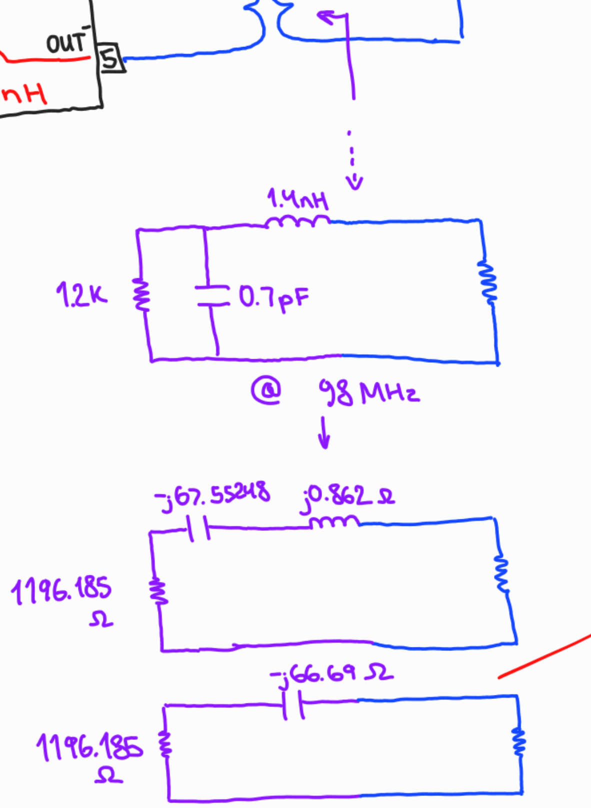


The input impedance is transformed to $103 - j3.24$

Output

- Need to find IF Amp input impedance
  ~300Ω (cascode amplifier in next section)
- Need to find Mixer output impedance





$66.69 = \frac{1}{2\pi f C}$

$\rightarrow C = 2.23 \times 10^{-10}$

$= 223\ pF$

the following transformation equations:

$$R_p = (Q^2 + 1)R_s \qquad \text{(Eq. 2-7)}$$

where

R_p = the equivalent parallel resistance,
R_s = the series resistance of the component,
$Q = Q_s$ which equals Q_p which equals the Q of the component.

and

$$X_p = \frac{R_p}{Q_p} \qquad \text{(Eq. 2-8)}$$

If the Q of the component is greater than 10, then,

$$R_p \approx Q^2 R_s \qquad \text{(Eq. 2-9)}$$

and

$$X_p \approx X_s \qquad \text{(Eq. 2-10)}$$

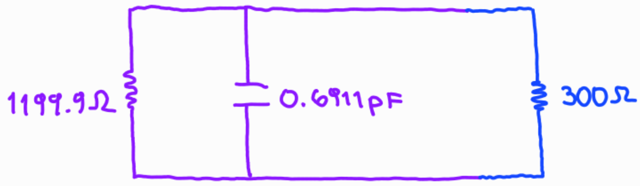
Now, because we want to match the mixer output impedance down with a shunt/parallel component, transform this output impedance to a parallel circuit:

$$Q_s = \frac{X_s}{R_s} = \frac{66.69}{1196.185} = 0.05575$$

$$R_p = (Q^2+1) \cdot R_s = 1199.9\ \Omega$$

$$X_p = \frac{R_p}{Q} = 21522.926\,\Omega \rightarrow 6.9109 \times 10^{-13}$$

0.6911 pF



– Match impedences

$R_p = 1199.9\,\Omega$

$R_s = 300\,\Omega$

$$Q = \sqrt{\frac{R_p}{R_s} - 1}$$

$$= 1.73195$$

Parallel capacitance:

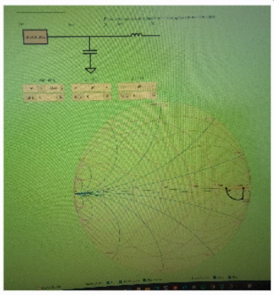
$$Q = \frac{R_p}{X_p} \rightarrow X_p = \frac{R_p}{Q} = 692.803\,\Omega \rightarrow C = \frac{1}{2\pi f X_p} = 21.4697\,pF$$

→ only $21.4697 - 0.6911 = 20.7786\,pF$ needs to be added due to internal capacitance.

Series inductance

$$Q = \frac{X_s}{R_s} \rightarrow X_s = Q \cdot R_s = 519.585\,\Omega \rightarrow L = \frac{X_s}{2\pi f} = 7.72846\,\mu H$$

Using practical values: $C = 20pF$, $L = 7.8\,\mu H$ we get $Z = 317 - 4.33j$



Filtering

Need $Q = \dfrac{10.7\,MHz}{400\,kHz}$ bandpass filter

$$= 26.75$$

channel width is 200KHz, but as done in the video series, a 400KHz -3dB bandwidth was used.
↳ because there are multiple filters like this in the receiver, the roll-off on out-of-channel frequencies will still be significant.

$$R_p = 300 \,||\, 300 = 150\,\Omega$$

$$Q = \frac{R_p}{X_o} \rightarrow X_o = 5.6075$$

$$X_o = 2\pi f L \qquad L = \frac{X_o}{2\pi f} = 83.41\,nH$$

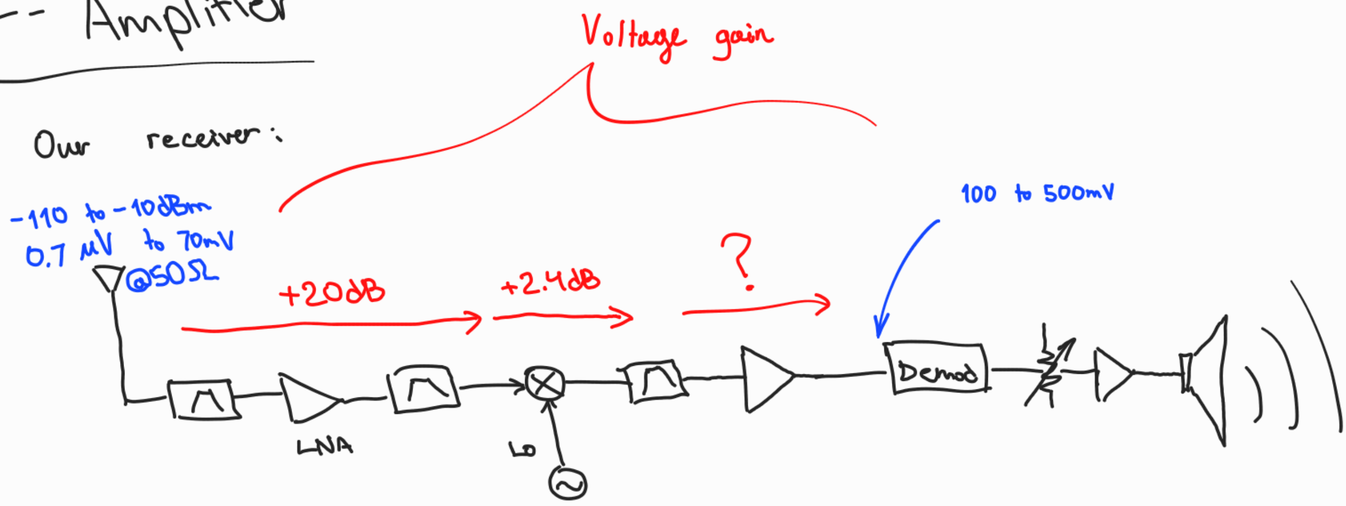
choose high-Q inductor

$$X_o = \frac{1}{2\pi f C} \qquad C = \frac{1}{2\pi f X_o} = 2.6526\,nF$$

Practical values are:
82 nH & 2700pF
$f_c = 10.696\,MHz$
BW = 392 KHz

IF- Amplifier

Our receiver:



As per the RF-receiver video series I followed, 0.7 µV signals should be amplified to ~ 100mV.

$$20 \log\left(\frac{0.1}{0.0000007}\right) = 103.098 \text{ dB}$$

Thus, the IF amplifier stage should have an amplification of

$$103 - 20 - 2.4 = 80.7 \text{ dB}$$

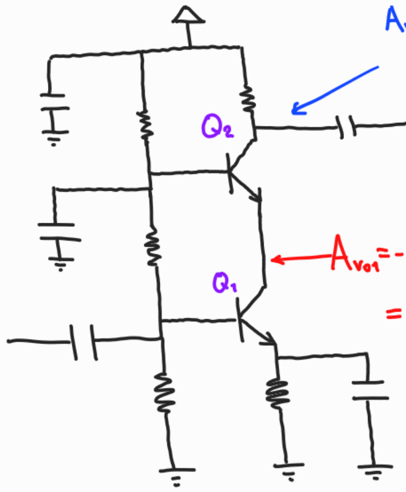
However, even in the video, only 40dB was used, which gave good results.

Now, let's design a cascode amplifier with ~40 dB of gain.

Why cascode?

- No Miller-Effect ⟶ higher bandwidth ?
- High gain

$A_{vo2} = g_m \cdot R_c$

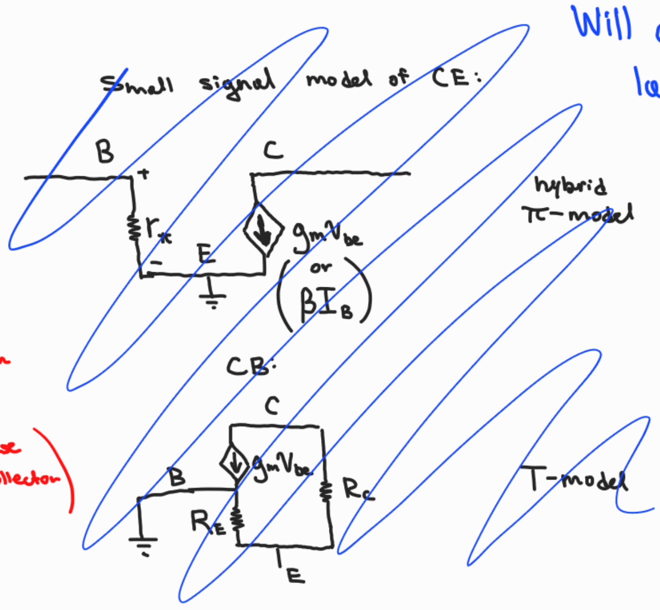


$A_{vo1} = -g_m R_c = -g_m \cdot \frac{1}{g_m}$

$= -1$

(A_{vo1}: gain from base of Q_1 to collector of Q_1)

Will derive later



For 50dB of gain,

$$40 = 20\log\left(\frac{\left(\frac{V_o}{2}\right)}{V_{in}}\right)$$

$$40 = 20\log\left(\frac{A_{vo}}{2}\right)$$

$$2 = \log(A_{vo}) - \log(2)$$

$$10^{(2+\log 2)} = A_{vo}$$

$$200 = A_{vo}$$

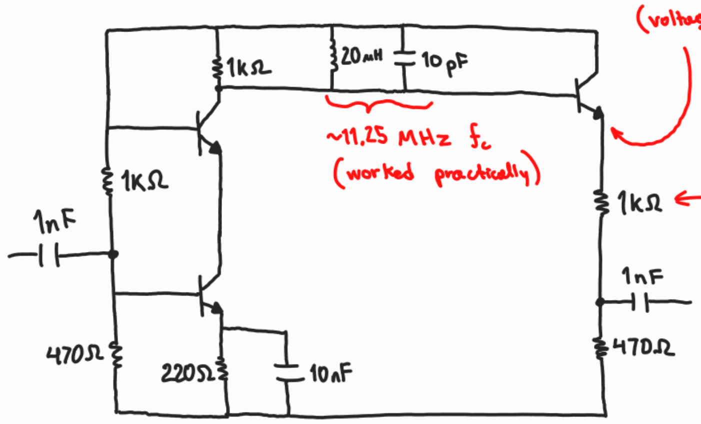
$$g_m = \frac{I_c}{26mV}$$

$$A_{vo} = g_m \cdot R_c$$

$$R_c = 1040\Omega$$

Note: Because I'm not sure how to bias a cascode amplifier, I will be copying an online design which has an $R_c = 1k\Omega$ (what I need)

common collector (voltage follower)

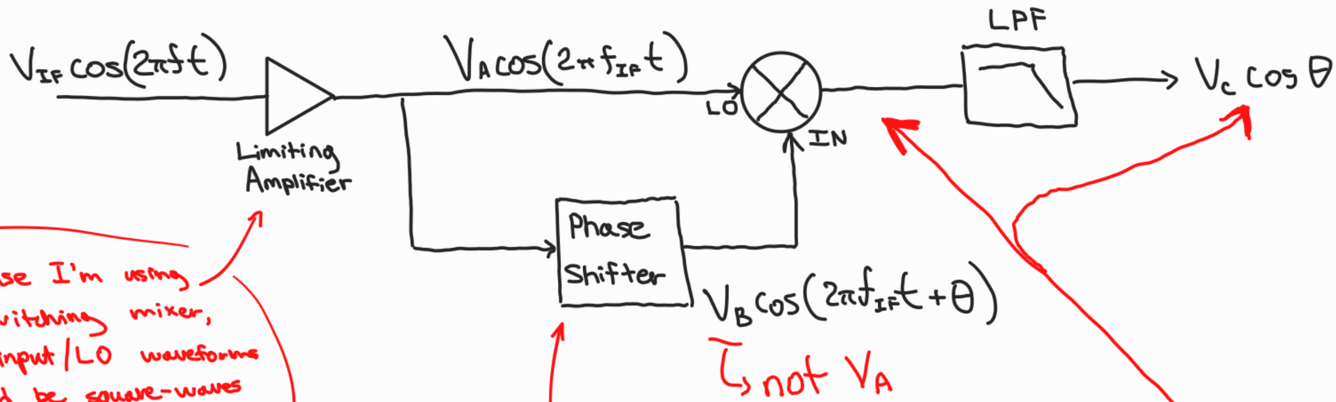


~11.25 MHz f_c (worked practically)

might not be desired as it is for voltage division.

DE-MOD (frequency discriminator)

"Classic FM Quadrature Subsystem"



Because I'm using a switching mixer, the input / LO waveforms should be square-waves for the best switching. I believe

+ The amplification should be determined by the max mixer input... Then an attenuator can be used to ensure the signal doesn't destroy a speaker/amp.

the purpose of this stage is to provide a phase shift (θ) which is linear with respect to frequency in the 10.7 MHz ± 75 kHz range.
The mixer will then map that phase shift to a signal value at the output, de-modulating the FM data!

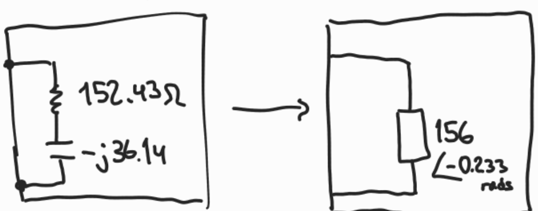
What limiting amplifier should be used?

The LO/Input maximum power (differential) is +10 dBm (10mW)

LO input impedance:

@ 150 MHz, this is the equivalent input impedance:

↑ from datasheet



At higher frequencies the impedance is smaller, thus it's safe to assume impedance ↑ with frequency ↓...

From this, we get a lower bound on the max voltage the LO input can tolerate.

$$\frac{V^2}{|Z|} = P \rightarrow V = \sqrt{P\cdot|Z|}$$

$$= \sqrt{0.01W \cdot 156}$$

$$= 1.249\,V$$

Same process for...

IF input impedance:

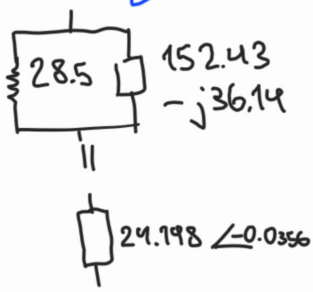
The input impedence seems to approach $28.5\,\Omega$.

$$V = \sqrt{P \cdot Z} = \sqrt{0.01\text{W} \cdot 28.5}$$
$$= 0.534\text{ V}$$

Now I need to find a limitting amp

who's V_{out} is s.t. $0.53 > V_{out} \cdot \dfrac{R_{out}}{24.198} = V_{in_{mixer}}$

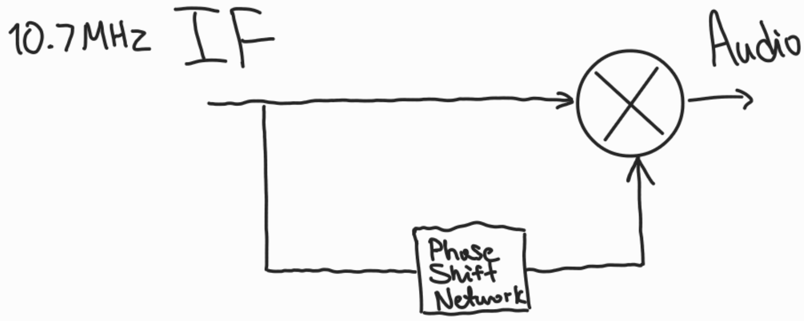
load impedance seen by amplifier



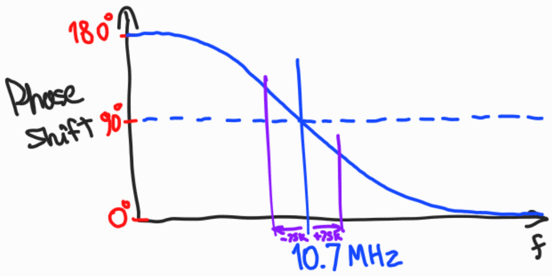
Although the video series mentions using a limitting amp, there isn't one in the final build.

Phase Shift Network

- frequency modulation is done by deviating from the channel's center frequency by ± 75 kHz
  - the frequency deviation should linearly map to a phase-shift value



A bandpass filter can be utilized to produce such a frequency response:



Here's such a filter

- Center frequency = 10.7 MHz
- Parallel resistance = Mixer IN input impedance = 28.5 Ω
- Filter LC values:
  - L = 12 nH
  - C = 0.018 μF
  - which are resulting from a desired Q = ~26.75 (for 400 kHz filter bandwidth)

(video series neglects amplifier output impedance & mixer LO input impedance for some reason... that is additional R_p! My mixer's LO input impedance is much smaller than that in the video series so I should be fine... he was fine too.)

(Values were obtained using simulations)

$Q = \frac{10.7M}{400k} = 26.75$ $R_p = 28.5$

$Q = \frac{R_p}{X_0}$

$X_0 = \frac{R_p}{Q} = 1.065\Omega$

$X_0 = 2\pi f L$

$\frac{X_0}{2\pi f} = L$

$15.847 nH = L$

$X_0 = \frac{1}{2\pi f C}$

$C = \frac{1}{2\pi f X_0}$

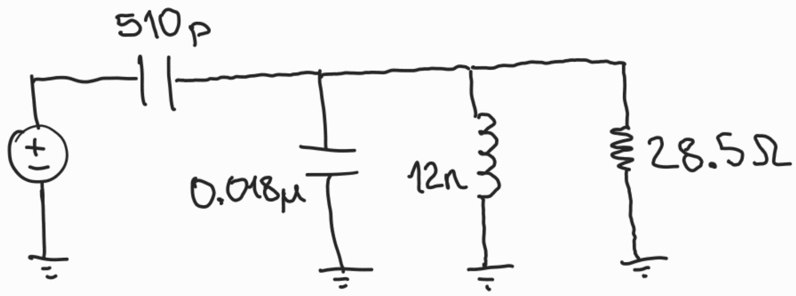
$C = 13.961 nH$

$C = \frac{1}{2\pi f \cdot 28.5}$

$= 521.9 pF$

value don't work

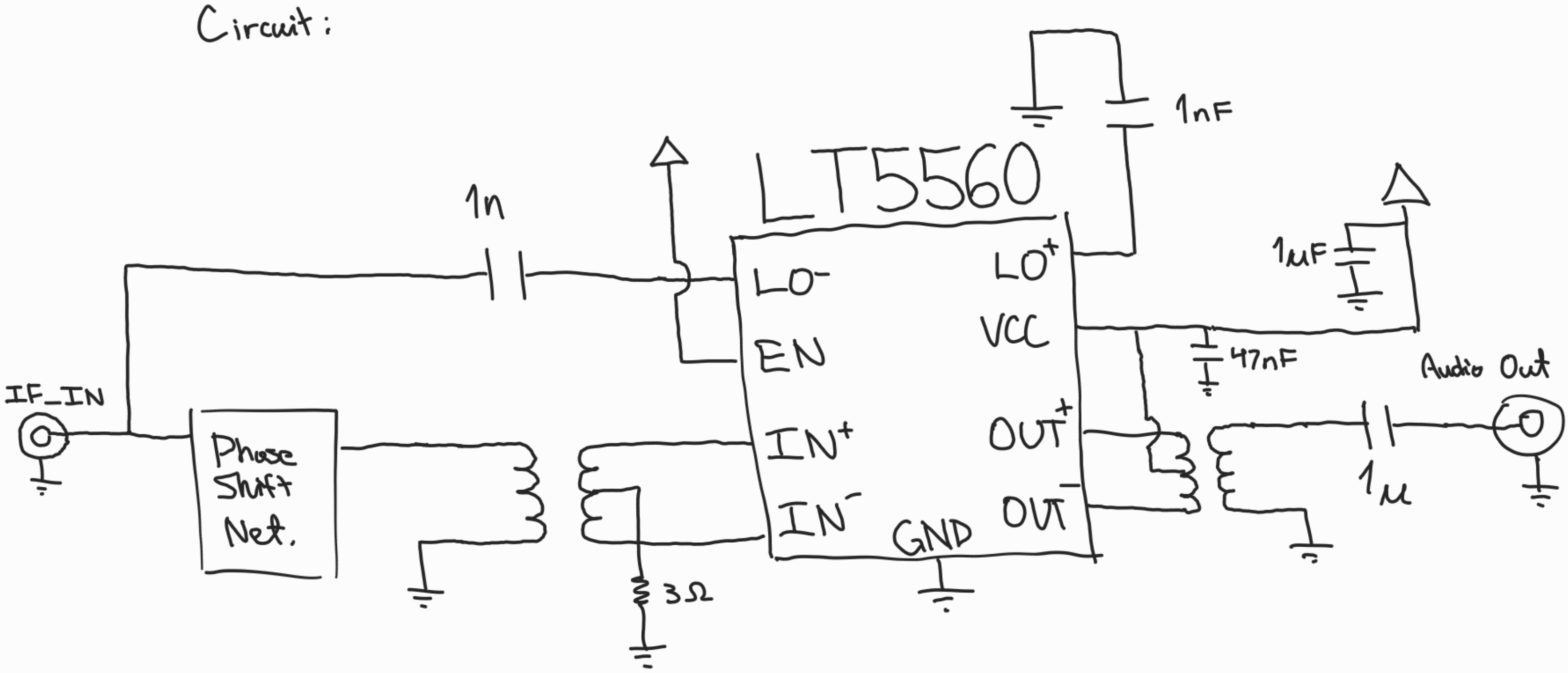
- Series capacitance = 510 pF



Choose high-Q components.

The filter inductor and/or cap should be tuneable due to the unpredictable effects tolerances & parasitics can cause.

Circuit:



* I am not adding Z-matching circuitry because the video series went without it. This causes some loss, but hopefully the amplifiers compensate for it (need to test!)

Audio Output Stage

Connect DE-MOD mixer output to powered, amplified speakers.

No stereo.

↳ How can stereo be implemented?