

EE3102 Design Proposal

AM Radio

Design Team #11

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System Specifications

- Single frequency reception of 1.4 MHz
- 1/8 Watt audio output to an 8-ohm load with 5 microvolt RF signal input at 50% amplitude modulation (provides 115 dB of gain)
- Low cost (less than \$60.)
- Minimal tuning
- Selective in the reception of only 1 AM BCB station (10 KHz channel width)

System Block Diagram and Description

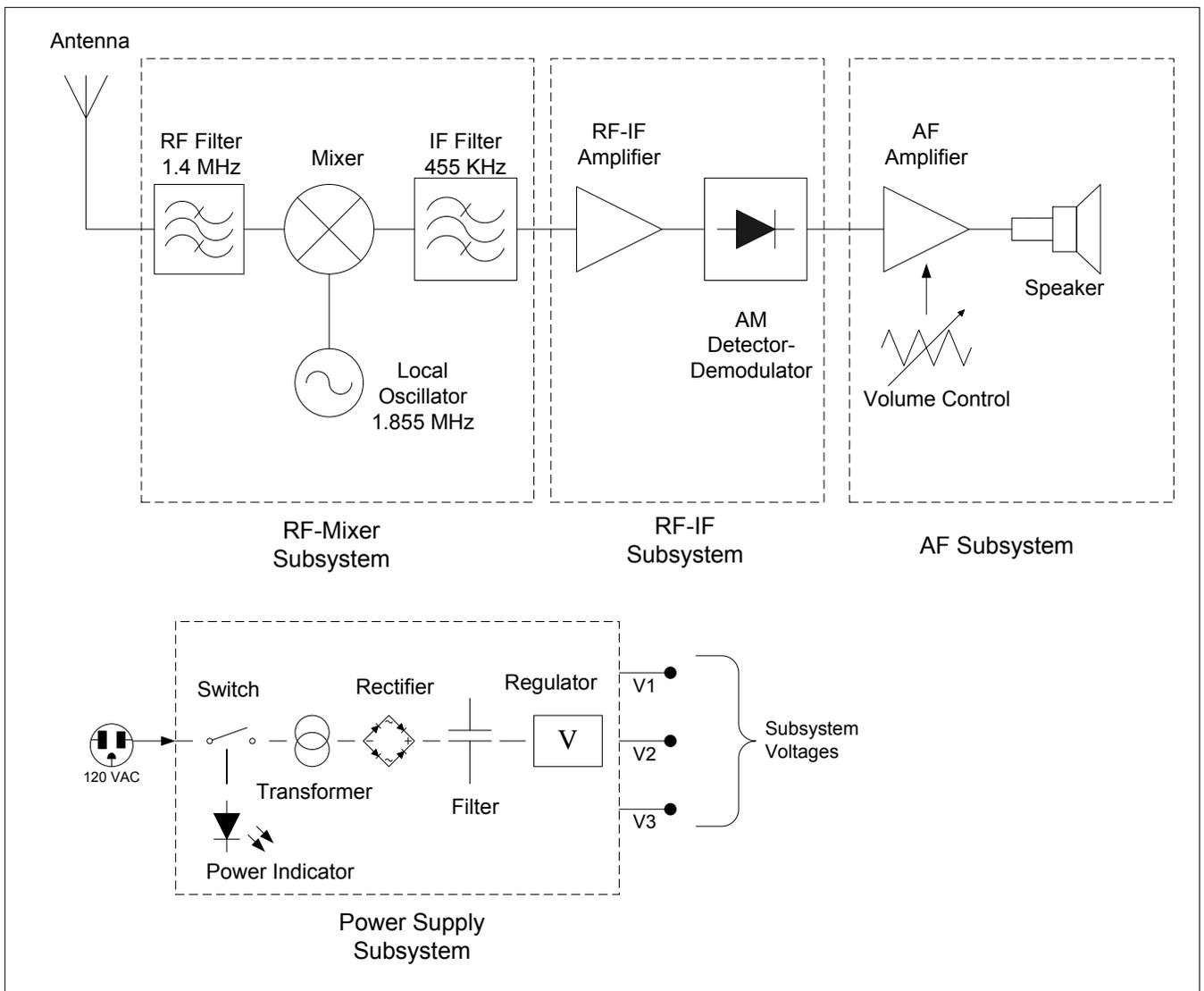


Figure 1: AM Radio Block Diagram

Block Diagram Description

It was decided early on to use a super heterodyne receiver topology for the following reasons

- More stable RF-IF amplifier design
- Higher shape factor of main bandwidth filter element
- Capable of tuning more than one frequency
- Readily available parts

Figure 1 shows the block diagram of this super heterodyne receiver. There are four subsystems to this receiver called the RF-Mixer, RF-IF Amplifier, AF Amplifier, and Power Supply. A description of each subsystem is as follows.

The RF-Mixer subsystem has a tuned circuit on the input tuned to 1.4 MHz. This circuit provides not only selectivity to the receiver, but also reduces the Local Oscillator signal from radiating out the antenna. The 1.4 MHz signal is then mixed with a 1.855 MHz Local Oscillator signal to produce the original signals plus the sum and difference frequencies. The difference frequency is chosen, 455 KHz, by the ceramic filter element, which has a 6dB bandwidth of +/- 12 KHz, and a stop bandwidth of +/- 24 KHz.

The AM modulated 455 KHz signal is then amplified by the RF-IF Subsystem by 60 dB. A standard diode detector circuit demodulates this amplified 455 KHz IF (Intermediate Frequency.) this demodulation process produces the original audio frequency modulation signal.

The AF Subsystem takes the detected audio and amplifies it to a level that is audible to a user through a speaker. It also has a volume control to allow the user to adjust the audio to the proper level for comfortable listening.

Finally the Power Supply Subsystem provides regulated voltages required by the other subsystems for proper operation. The subsystem uses the 120 VAC circuit mains voltage as a source for its power. An indicator lamp is also use along with a switch to turn on and off the power delivered by the mains and indicate when the power is on.

Breakdown of Group 11 Tasks

The following subsystems will be divided into the following group 11 team members for design and development:

Team Member	Subsystem
Samuel Annor	RF-IF

Erik Newman	Power Supply
Eliot Estrine	AF
Brian Teravskis	RF-Mixer

RF-Mixer Subsystem

Subsystem Specification

- Provide at least 13 dB of conversion gain of the input RF signal
- Provide 20 KHz of signal selectivity
- Minimize radiated LO signal
- LO tunes to 1.855 MHz
- IF frequency of 455 KHz
- Optional – provide tuning of a segment of the AM broadcast band

Subsystem Description

To meet these specifications, and to minimize the cost and development time it was decided to use the SA612 Integrated Circuit by Philips Semiconductor. This IC uses a Gilbert-Cell Doubly Balanced Mixer, which provides a conversion gain of 14 dB at 100 MHz, and increases as the frequency decreases. The SA612 also has an internally pre-biased transistor that can be used as a Local Oscillator.

Figure 2 shows the proposed circuit using the SA612 Gilbert Cell Mixer Integrated Circuit in the RF-Mixer Subsystem. The Gilbert cell is a cross-coupled differential amplifier where the gain is controlled by varying the emitter bias current. If a modulating signal is applied to the emitter bias control the result is a highly linear control of the amplitude of the RF input voltage. The Gilbert cell configuration used in the SA612 provides a Double Balanced Modulator. This kind of modulator suppresses the original frequency components of the mixing process so that only the sum and difference frequency is present on the output. The suppressions reduces the harmonic content of the output and eases the design of any post mixing filtering.

The oscillator section of the SA612 is nothing more than a pre-biased transistor providing base and emitter connectivity to the external circuitry. It is important to note that the external circuitry must not alter the bias to this transistor, and often capacitors are required to isolate the internal DC bias from the external circuitry. There are several oscillator topologies that can be used with this internal transistor; in this design a Colpitts Oscillator circuit is employed due to both the stability of the circuit as well as the limited availability of tapped inductors. An external oscillator may be used with the SA612, but a limit of 200-300 millivolts must be maintained to prevent overdriving the part.

The SA612 is a good mixer, but has its limitations. The ability of the mixer to handle a wide dynamic range of RF inputs signals are limited, and therefore usually not desirable

as a first mixer of a super heterodyne circuit. Since only the reception of one frequency is required for this circuit, the SA612 will perform adequately.

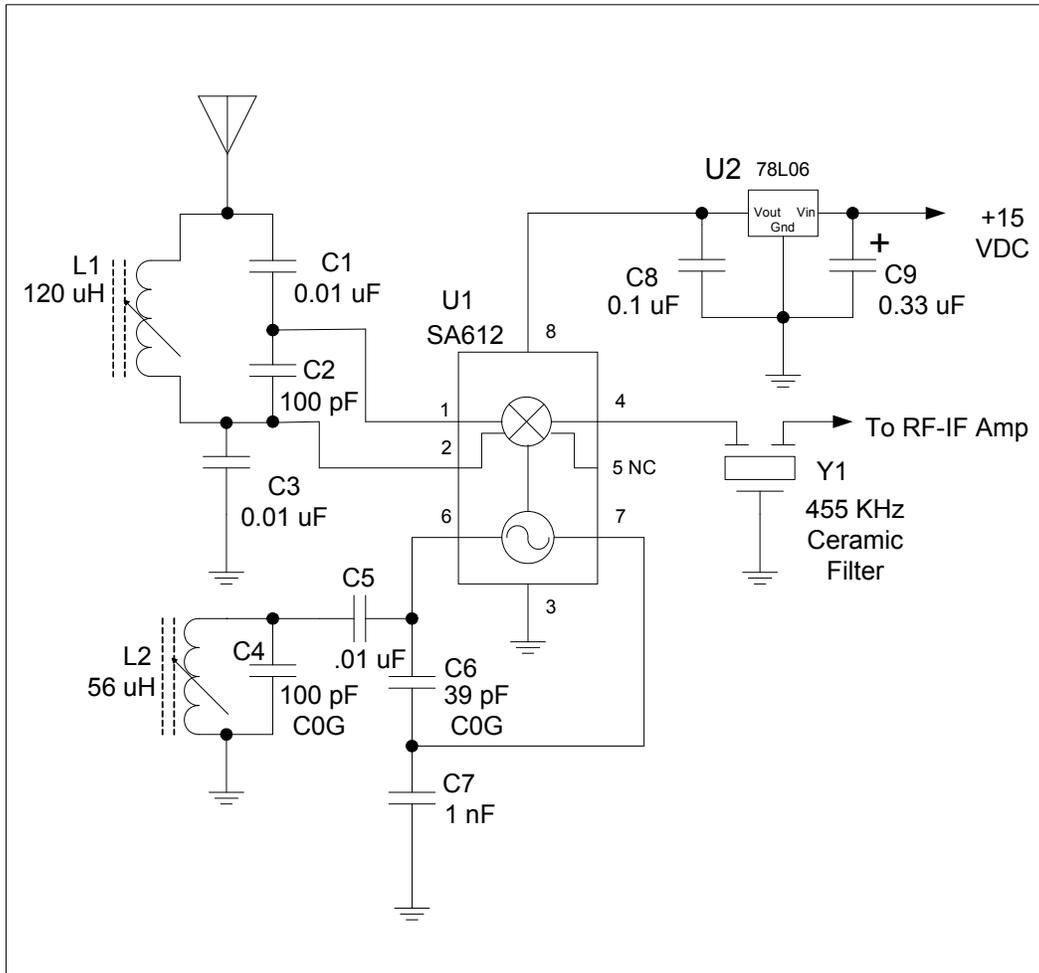


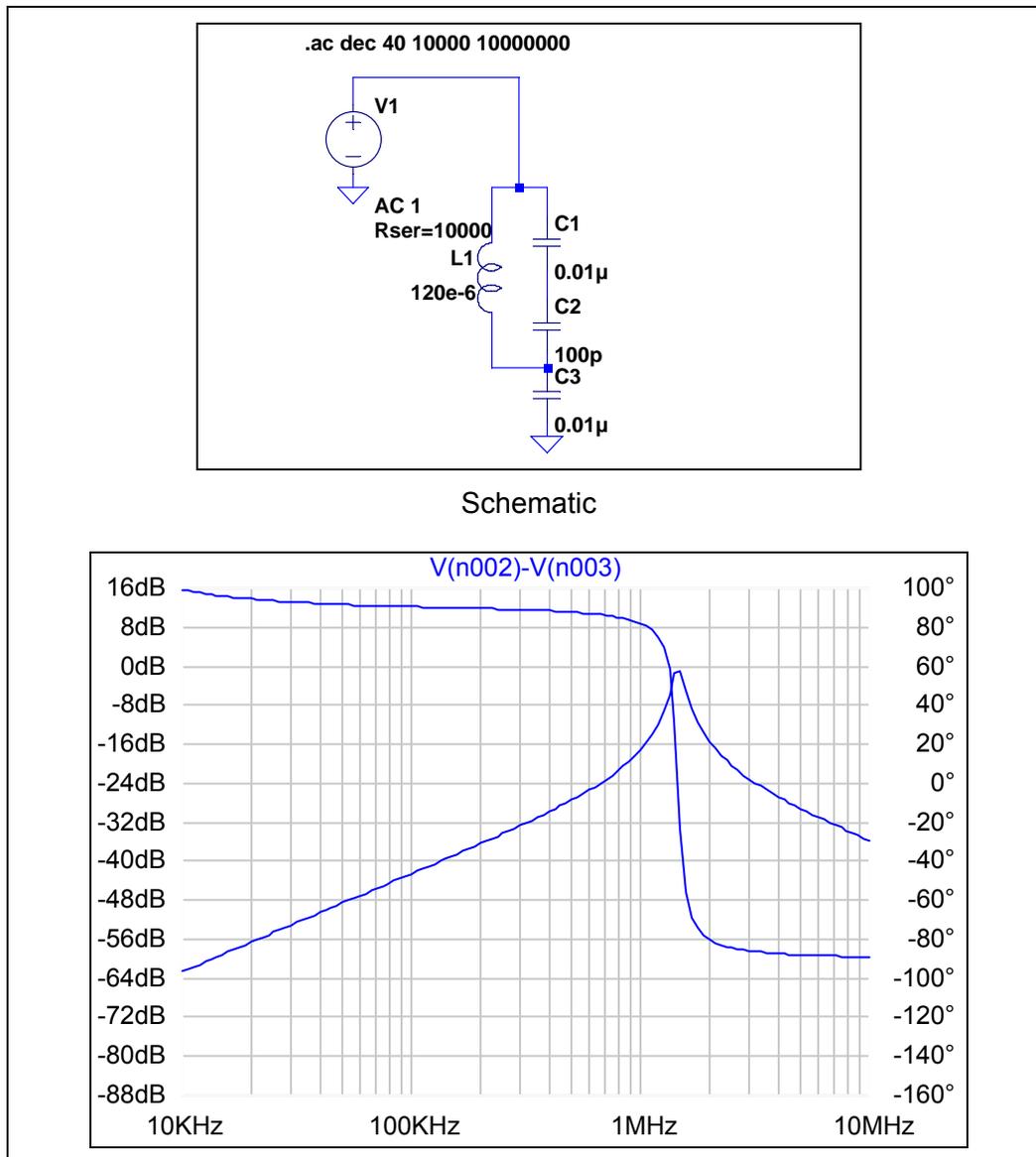
Figure 2: RF-Mixer Subsystem

The output of the Gilbert cell DBM is fed to a ceramic filter. The ceramic filter uses piezoelectric ceramic elements as electromechanical transducers. A ceramic element is energized by an external signal that sets up vibrations within the structure. The ceramic element is shaped in a fashion that causes it to resonate at a predetermined frequency. The Q of this configuration is very high making the frequency selection fairly narrow. In the ceramic filter several of these ceramic piezoelectric elements are connected together to form a shaped bandpass filter response. The part selected for this design utilizes six elements in ladder formation. The bandpass response has a 6dB roll off of +/- 12 KHz. This response of this filter meets the selectivity requirements of the receiver design.

Simulations

Front End Tuned Circuit

The SA612 has pre-biased circuitry that must be isolated from external references. The data sheet recommended a front-end tuned circuit that was adopted for this design and referenced in Figure 3. This tuned circuit has a center frequency of 1.4 MHz, and has approximately -30 dB attenuation at 455 KHz and -13 dB attenuation at 1.855 MHz. This attenuation should minimize the LO and IF frequencies, as well as any other higher harmonics, from radiating out of the antenna. The antenna has high impedance at 1.4 MHz, and therefore will not load down the tuned circuit when connected directly. Figure 4 shows the LTSpice schematic and simulation of this tuned circuit.



Frequency Response

Figure 4: Front-end Tuned Circuit Simulation

Gilbert Cell Mixer

No Spice model of the SA612A could be found, and therefore the equivalent circuit was used to estimate the amount of conversion gain the mixer would produce at 455 KHz. The SA612A product specification sheet has an equivalent circuit in figure 4, but there are internal bias voltages and certain resistors values that are not shown. A best guess was made using voltage sources, and a current source for the current mirror (represented by a resistor in the equivalent circuit.) A VHF BJT transistor was used (2N5089) in the simulation as the SA612A operates up to 500 MHz input frequency. Figure 3 shows the circuit used in the LTSpice simulator.

The results of the simulation are shown in the FFT of the output waveform at the collector of Q3. The output frequency is marked and measured at -88.5 dB. With an input voltage of 5 microvolts, or -106 dB, the conversion gain is 17.5 dB.

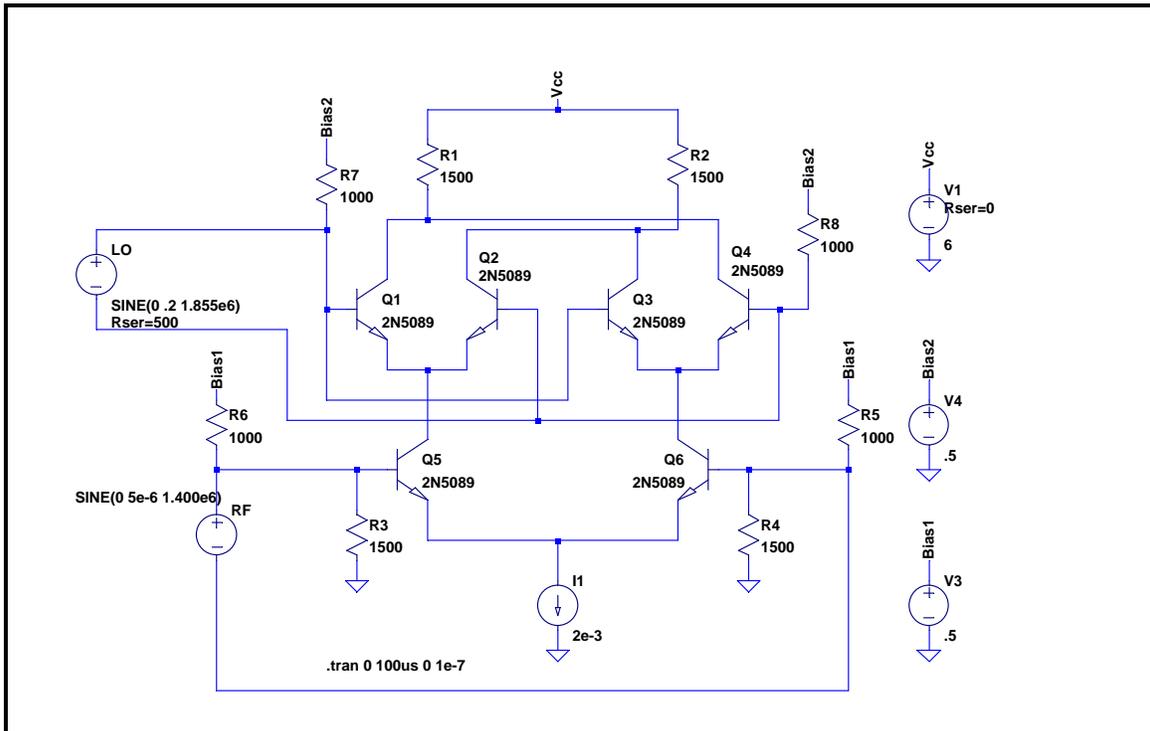


Figure 3: Gilbert Cell Mixer Simulation Schematic

There are many assumptions in this simulation that impact the accuracy as the bias voltages and current greatly impact the conversion gain, and it is not know what the

manufacturer of the part uses. It is safe to say that a conversion gain of 14 dB or better (specified in the data sheet) can be realized in the circuit, and therefore meets the design requirements of the subsystem.

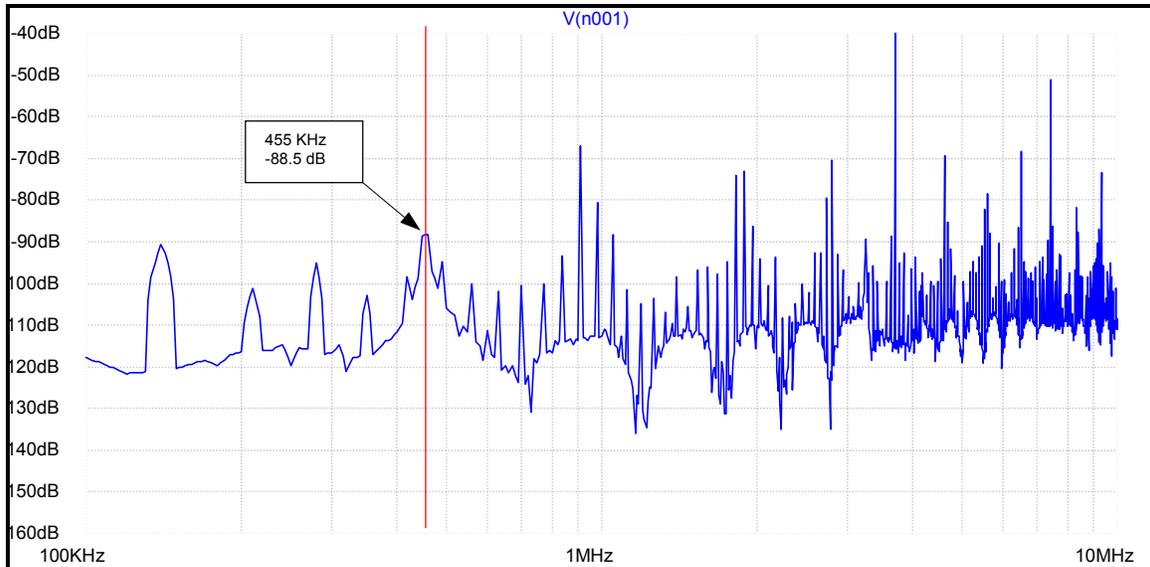


Figure 4: Gilbert Cell Mixer Output FFT

Colpitts Oscillator

As in the Gilbert cell simulation above, the SA612A does not give all of the details concerning the oscillator BJT parameters. Again, the SA612A oscillator section is capable of sustaining oscillations up to 200 MHz, and therefore the simulation uses the same BJT as was used in the Gilbert cell simulation (2N5089.) Figure 5 shows the circuit utilized in the simulation, and eventually in the design.

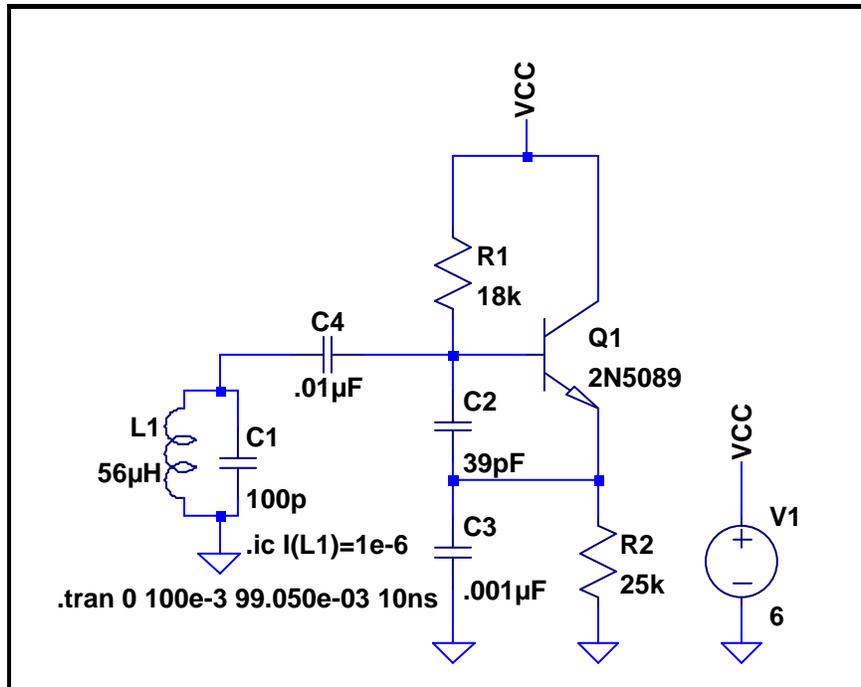
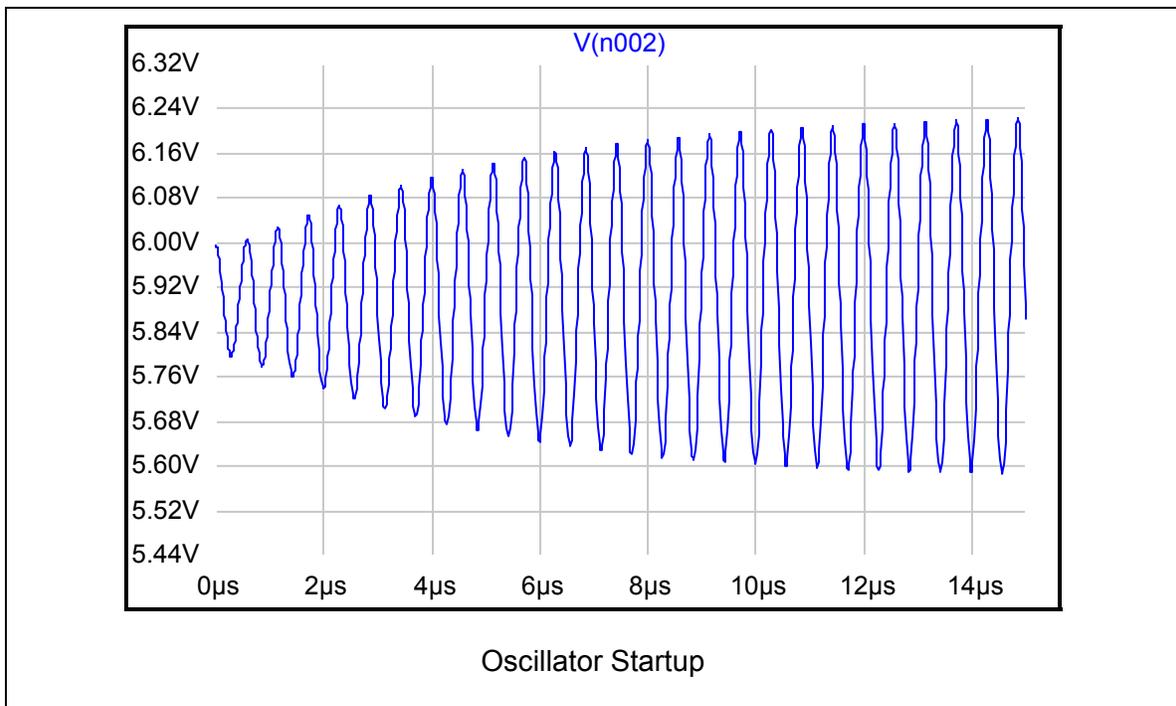


Figure 5: Colpitts Oscillator Simulation Schematic

Figure 6 shows the startup, sustained, and FFT transient simulation of the oscillator used for this design.



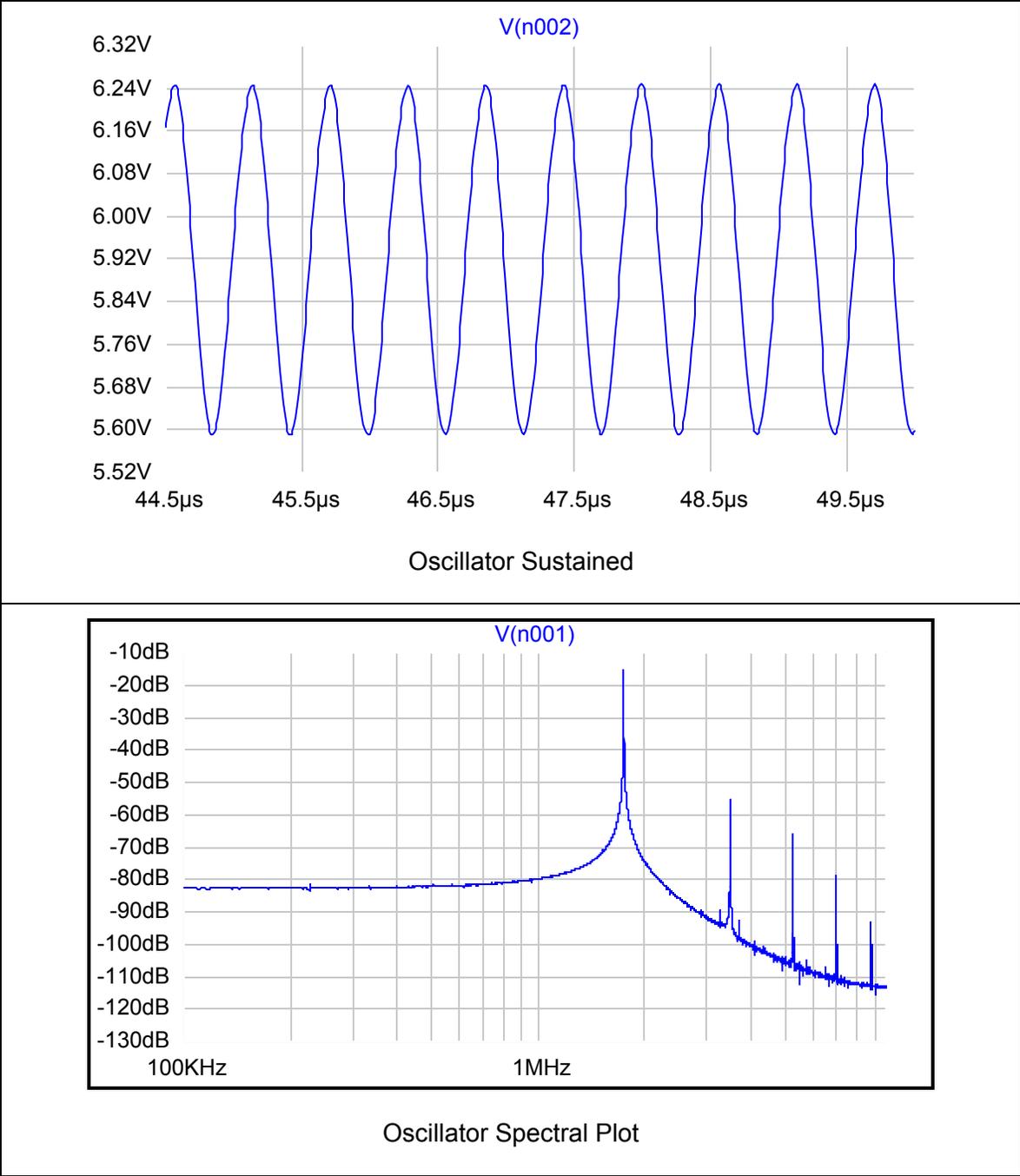


Figure 6: Colpitts Oscillator Simulation Results

Prototype Test Results

There were only two of the four tests that were done before this document. Table 1 lists the test that were required, what was expected as a result, and the measured value.

Measurement	Expected Value	Measured Result
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Mixer conversion gain	14 dB @ 455 KHz	5dB @ 455 KHz
Mixer oscillator frequency	1.855 MHz	1.470 to 1.987 MHz
Mixer oscillator amplitude	200 mVp-p	220 mVp-p
IF selectivity – 3dB roll off	20 KHz	13 KHz
LO front end leakage	-30 dB from LO level	Not measured

Table 1: RF-Mixer Specification Measurements

The mixer was probably driven into gain compression to make the conversion gain measurement, as a fairly high voltage input was required to get a reading on the output. Once the RF-IF amplifier gain is determined the conversion gain will be measured again at the output of the RF-IF amplifier.

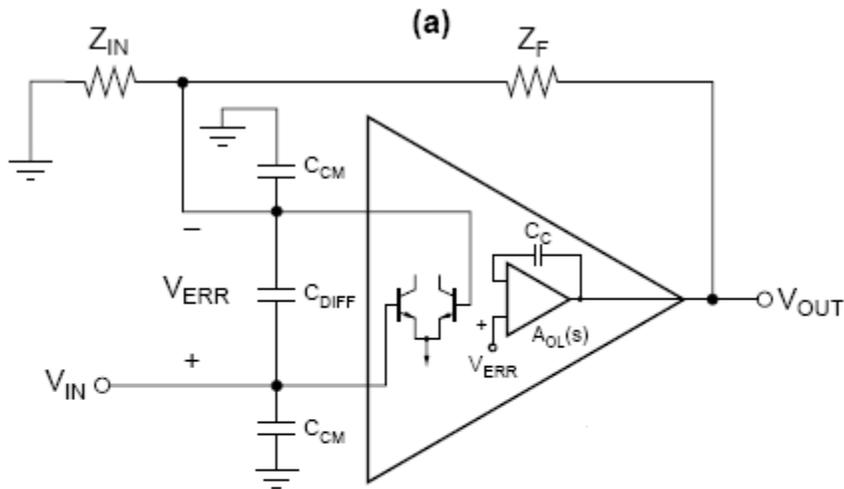
The oscillator coil adjustment will span the AM broadcast band from 1.125 MHz to 1.542 MHz. Just before completing the measurements it was tuned to 1.4 MHz for a final test in the lab at a later date.

The LO leakage requires a spectrum analyzer connected to the antenna. The plan is to use the spectrum analyzer in the lab to perform the measurement, but there was not enough time to complete this before this document was due. It will also be measured at a later date.

Bill of Materials

Part Number	Supplier	Qty	Each	Total	RefDes	Description
81-CFWLA455KCFA-B0	Mouser	1	\$3.35	\$3.35	CF1	Filter, Ceramic, 455 KHz
TK2418-ND	Digi-Key	1	\$4.20	\$4.20	L1	COIL 120UH TYPE 7P
TK2414-ND	Digi-Key	1	\$4.31	\$4.31	L2	COIL 56UH TYPE 7PH
568-2089-5-ND	Digi-Key	1	\$1.75	\$1.75	U1	IC MIXER/OSC DOUBLE-BAL 8-DIP
497-1010-ND	Digi-Key	1	\$0.72	\$0.72	U2	78L06 IC POSITIVE REGULATOR 6V TO-92
P4849-ND	Digi-Key	2	\$0.11	\$0.22	C2, C4	CAP 100PF 200V 5% CER RADIAL C0G
495-1007-1-ND	Digi-Key	1	\$0.16	\$0.16	C6	CAP 39PF 100V 5% CER MONO C0G
P4951-ND	Digi-Key	3	\$0.13	\$0.39	C1, C3, C5	.010UF 50V 10% MONOLIT CERM CAP X7R
P4923-ND	Digi-Key	1	\$0.12	\$0.12	C8	.1UF 50V 10% MONOLIT CERM CAP X7R
P4937-ND	Digi-Key	1	\$0.12	\$0.12	C7	CAP 1000PF 50V CERAMIC MONO 5% C0G
P928-ND	Digi-Key	1	\$0.18	\$0.18	C9	KG RAD ALUM ELEC CAP 0.33MFD 50V
RF-Mixer Subsection Total				\$15.52		

RF-IF Amplification and Demodulation Subsystem



$$A_{OL}(s) = \frac{G_{DC}}{1 + r_O C_C s}, \quad s = j\omega$$

$$V_{OUT} = A_{OL}(s) \cdot V_{ERR}$$

$$\frac{V_{IN} - V_{ERR}}{Z_{IN}} = \frac{V_{OUT} - (V_{IN} - V_{ERR})}{Z_F}$$

Figure 7a: Voltage Feedback Amplifier Configuration in a non-inverting Close-Loop configuration.

Even though the current feedback amplifier is easier to design than the voltage feedback amplifier since all it takes is for it to have correct feedback impedance is selected, the voltage feedback amplifier is used here.

The voltage feedback amplifier is the most abundant amplifier on the market due to their used in several high speed as well as precision applications. Since the preferred frame of reference for most analog designers is the voltage feedback amplifier, the stability analysis begins with the topology shown in Figure 7a. This simplified block diagram illustrates many of the key characteristics needed in a frequency analysis of the voltage feedback amplifier. Starting with the input segment of the amplifier, the inputs to the voltage feedback amplifier are evenly matched. In most cases, the input bias currents are very close to being the same magnitude and the difference between the two input bias currents is usually small. Additionally, the input impedances of the amplifier are

close to equal and relatively high. A small error voltage at the input of the amplifier is gained by the open-loop gain, $A_{OL}(s)$, which is usually fairly high.

The open-loop gain is frequency dependent and starts rolling off at a relatively low frequency. The resulting open-loop voltage output is the product of the open-loop gain times the input voltage error.

As expected, the calculation in Figure 1b proves that the closed-loop DC gain is equal to $1 + R_F/R_{IN}$. At low frequencies the open-loop gain of the amplifier is sufficiently high to allow for ignoring the gain error. As frequency increases, $A_{OL}(s)$ begins to decrease and finally becomes the dominant controlling factor in the gain of the circuit. The calculation of the intersection of the open-loop gain, $A_{OL}(s)$ and the noise gain (also called $1/\beta$), $G_N = (1 + Z_F/Z_{IN})$, gives a close approximation to the bandwidth of the closed-loop amplifier circuit. Gain peaking, which is caused by the phase response of the amplifier and the feedback circuit, can increase the bandwidth of the closed-loop system at the expense of decreased stability. The criterion for stability is determined where the closed-loop noise gain intersects the open-loop gain of the amplifier. Generally, if the phase margin, at that frequency, is smaller than a theoretical 45 degrees, the circuit can oscillate given part-to-part variation. A conservative phase margin is more like 60 degrees for flat frequency response.

(b)

$$\frac{V_{OUT}(s)}{V_{IN}(s)} = \frac{(1 + Z_F/Z_{IN})}{1 + \underbrace{(1 + Z_F/Z_{IN})/A_{OL}(s)}_{G_N}}$$

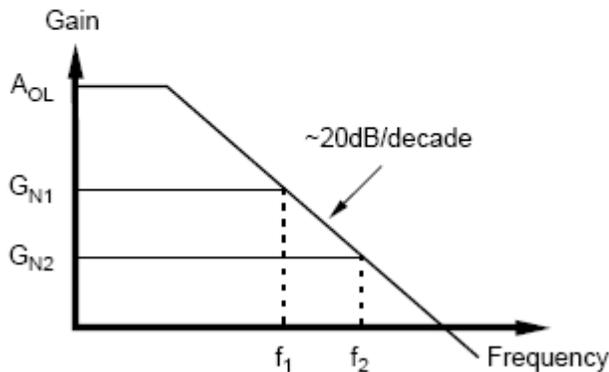


Fig 7b: Transfer function of the voltage feedback amplifier in a non-inverting configuration.

The open-loop gain plot of the amplifier assumes a single pole system. As the closed-loop gain increases, the closed-loop bandwidth decreases.

Matched inputs may or may not be a benefit when using voltage feedback amplifiers. In addition, offset voltages and offset currents are relatively low compared to the current feedback amplifier topology. A possible disadvantage of the voltage feedback amplifier is the very close relationship between the bandwidth and closed-loop gain however in the AM radio RF amplification such disadvantages are minimized.

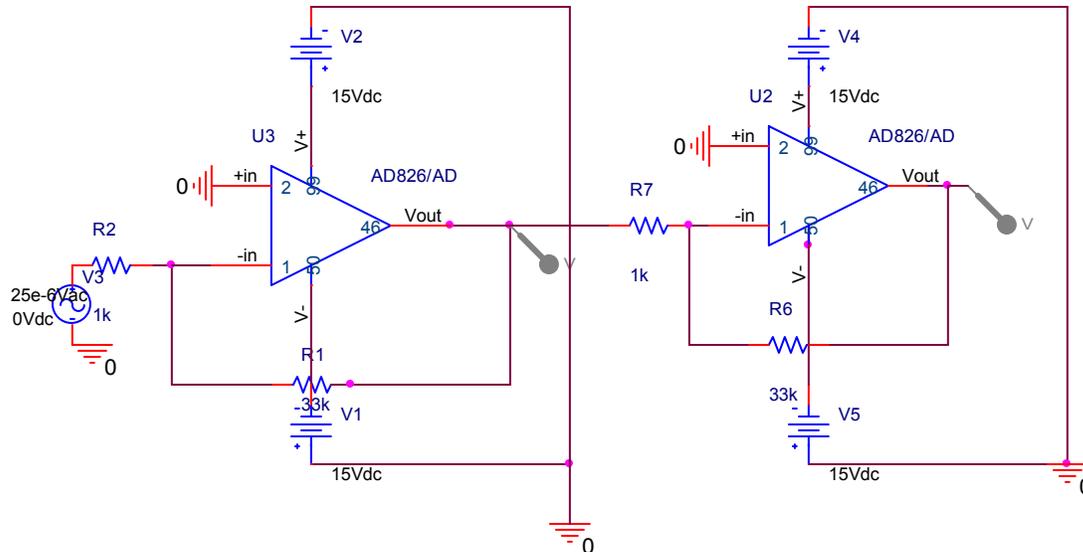


Fig 2a The 2 cascaded inverting voltage amplifier.

Apart from choosing the voltage amplifier for this application, the type of operational amplifier (Op-Amp) that was chosen Analog Devices' AD826 op-amp because it has the GBWP or the Unity Gain Bandwidth of 50MHz.

This will give a close-loop gain close of about 30dB (thus choosing $R_1=33k$ and $R_2=1k$

and since $A_v = -\frac{R_1}{R_2} = -\frac{33k}{1k} = -33 = 30.37dB$) from each stage with a carrier

frequency of 455 kHz. From the Bode plot below one can see that the gain is very close to constant gain and with about a gain of 60db

Subsystem Simulations

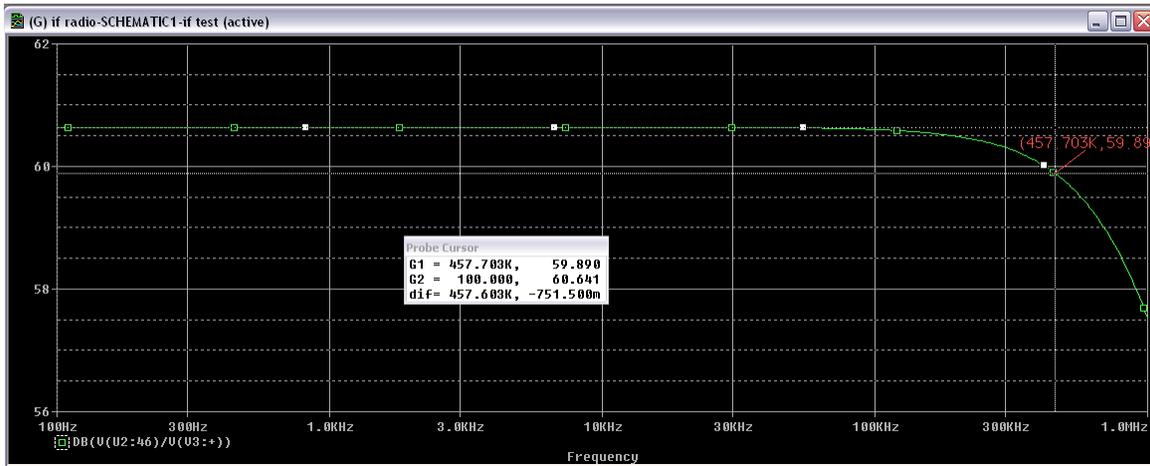


Fig 2b The Frequency Response of the output of the cascaded amplifier.

The next was to design the detector circuit which is shown in figure 3a.

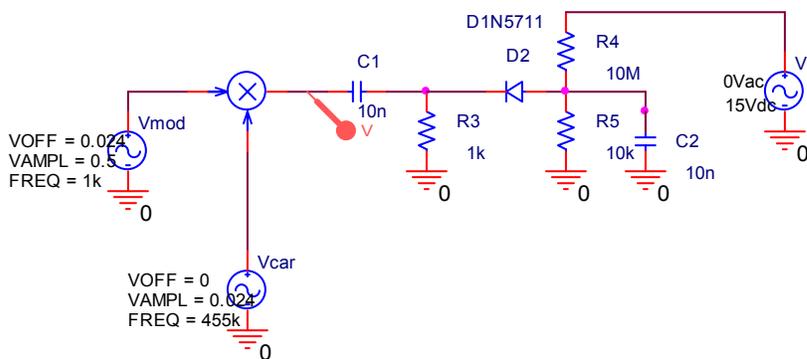


Fig 3a Detector circuit.

The detector consists of a Schottky diode with a RC network in front of it. However we could have had a precise rectifier circuits like the one that is based on the AD8037 which is a clamping operational amplifier. This circuit does not use any diodes to provide rectification. Problem is it is expensive (\$8.00), requires a two rail power supply, and surface mount is the only thing available at Digi-Key. Very undesirable..

So due to the above reason, the Schottky diode is only 41 cents or so and will do a great job since with a pull up resistor, of 10MΩ, and a DC supply, of 15V, will ensure that the 1N5711 is always turn on.

From fig 3a, a simple modulator circuit is used to generate the AM consisting of a Multiplier and the modulating signal and the carrier signal as done in PSpice. A sinusoidal source at a frequency of 455 kHz is used to represent the carrier signal and a

second source at a frequency of 1 kHz is used to represent the modulating signal. Notice that the peak amplitude for the carrier is set to 24mV, the estimated amplitude of output signal of the RF amplifier. A typical broadcast

AM signal includes the carrier as well as the sidebands in the transmission. To get such a double sideband transmitted carrier signal (DSB-TC) we must bias or offset the modulating signal by a value equal to the carrier's peak voltage.

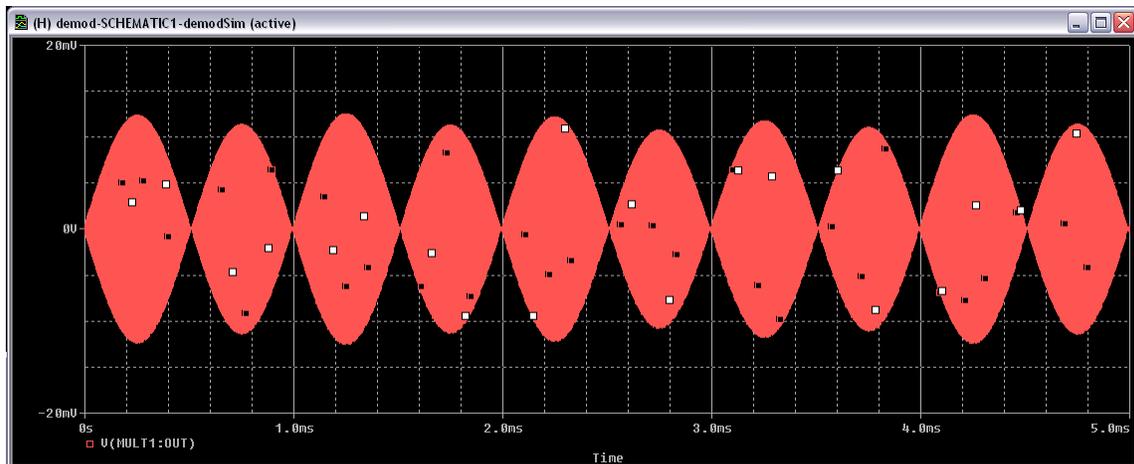


Fig 3b, Generated AM signal

AF Subsystem

Subsystem Specification

- Audible at 6 feet

Subsystem Description

The audio amplification subsystem produces the audio output of the entire system. The input signal is in the range of several millivolts, this value is far too low to drive a speaker and produce an audible output, and must be amplified. For the output of the radio to be audible, approximately $1/8^{\text{th}}$ of a watt must be delivered to the speaker. To accomplish this, the overall gain of the audio amplifier subsystem must be around 45 to 60 dB. In addition, the circuit must be able to provide a significant amount of current to the load without altering its performance. Finally, this subsystem will be responsible for the volume control requirements of the radio.

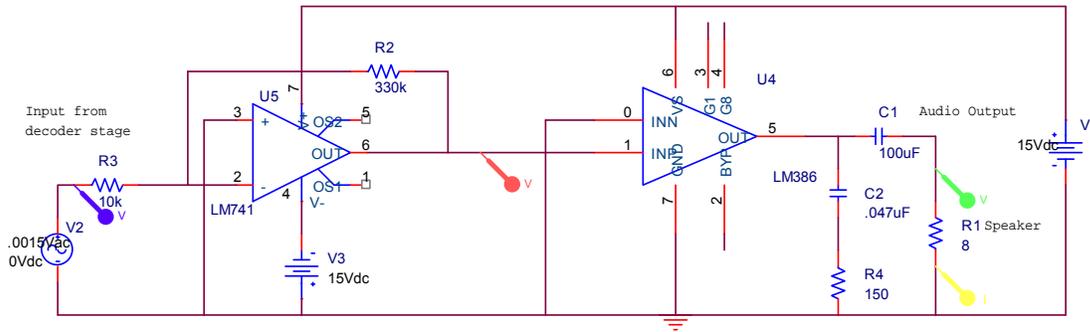


Figure 1: The two stage circuit used to produce the audio output. The gain of the circuit shown is 330.

The output load of this subsystem is an 8 ohm speaker. In addition for the output to be audible, approximately $1/8^{\text{th}}$ of a watt must be delivered to the speaker. For an 8 ohm speaker, the current required is:

$$I = \sqrt{P/R} = \sqrt{(1/8 \text{ watts}) / (8 \text{ ohms})} = 125 \text{ mA}$$

An LM741 can provide a maximum of 40 mA and therefore is inadequate for this application. The output stage also uses an RC network to create a band pass filter and filter out any signals outside the normal audible range which may have been caused by noise or resonance effects. The upper cutoff frequency was chosen to be 30 kHz while the lower cutoff frequency was chosen to be 200 Hz. The speaker itself has a frequency range of 450 Hz to 13 kHz, so it will provide additional band pass characteristics to the subsystem.

The primary device used in this subsystem is the LM386 audio power amplifier. This device has several advantages over the standard operational amplifiers used in lab and elsewhere in this project. The first, and primary reason for choosing this amplifier, is that it is able to produce a large output current. In addition, this device only requires a single power supply. The model chosen was the LM386N-3 because its voltage requirement closely matched those of other components used in the project. The LM386 uses internal feedback resistors to produce a gain of 20 without any external resistors. However, the gain can be varied up to 200 by using external components. Downsides of this device include lower maximum gain, and a lower gain bandwidth product than other amplifiers. However, because this stage is only required to amplify audio frequencies, the lower gain bandwidth did not affect the design.

The equivalent schematic of the LM386 is shown below. The device can be represented by a differential pair followed by a class AB power amplifier circuit. Voltage gain is produced by the differential pair, a common input circuit configuration. Power gain is produced by a class AB amplifier. The class AB amplifier is biased using the current source and diodes. In addition, the LM386 uses internal resistors, to provide feedback for the two stage amplifier. In the circuit below, R13 and R12 provide feedback to produce the overall gain of the device. By placing external components such as resistors or capacitors in parallel with R12, the amount of feedback can be altered, and the overall gain of the device can be controlled.

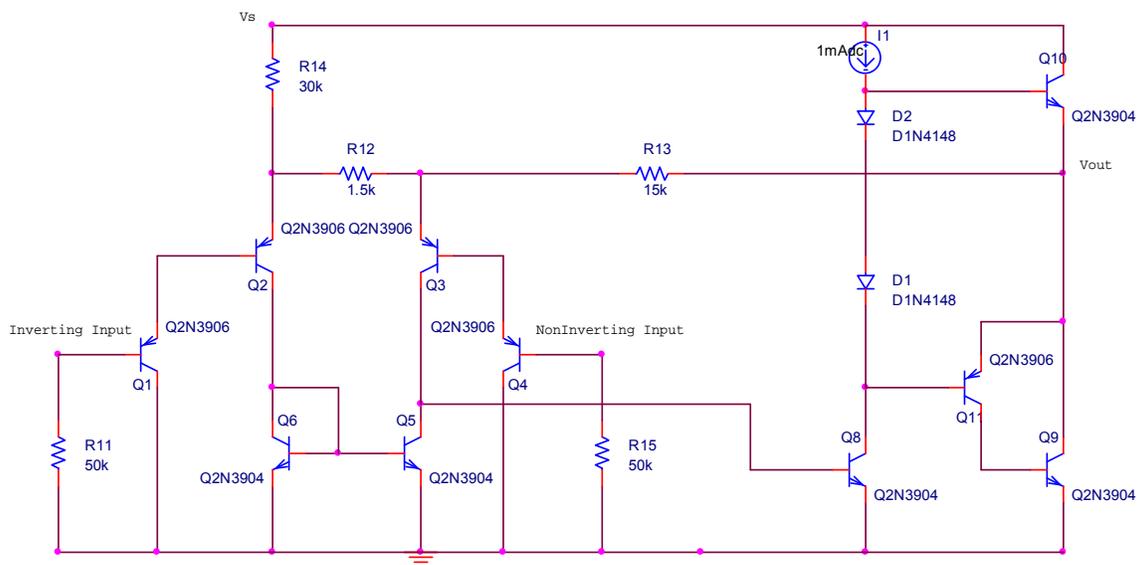


Figure 2: The internal structure of a LM386 amplifier.

To compensate for the lower maximum gain, a second amplification stage was used to increase the overall gain of the subsystem. This amplification stage was composed of a LM741 operational amplifier in an inverting configuration. The gain of this stage is controlled by the resistor R4 and R5 in the schematic below, with the gain equal to $-R4 / R5$. For the resistor values shown below, the gain of the LM741 stage is equal to $-330k/10k$ or -33 . When combined with the gain of 20 provided by the LM386, the total gain of the subsystem is -660 . Since the subsystem is amplifying an audio signal, inversion of the signal does not affect the quality of the signal.

A potentiometer is used between the LM741 and the LM386 to provide volume control. By varying the potentiometer, a user can control the voltage division that occurs between the two stages. If no voltage division occurs, the output will be at maximum value, however, if the potentiometer is altered, the output volume can be linearly reduced to nothing.

The functionality of the subsystem was tested using PSpice. The circuit shown above was simulated, and produced the output shown below. The high pass cutoff frequency was found to be 197 Hz, while the low pass cutoff frequency was found to be 29.85 kHz. These numbers are consistent with what was predicted for the circuit. By limiting the bandwidth of the device, high and low frequency noise can be filtered out to produce a clearer sound.

Subsystem Simulations

To simulate the LM386 audio amplifier, a new PSpice device and model had to be created. The model is based on a Spice model created by Dave Dilatush, and made publicly available.

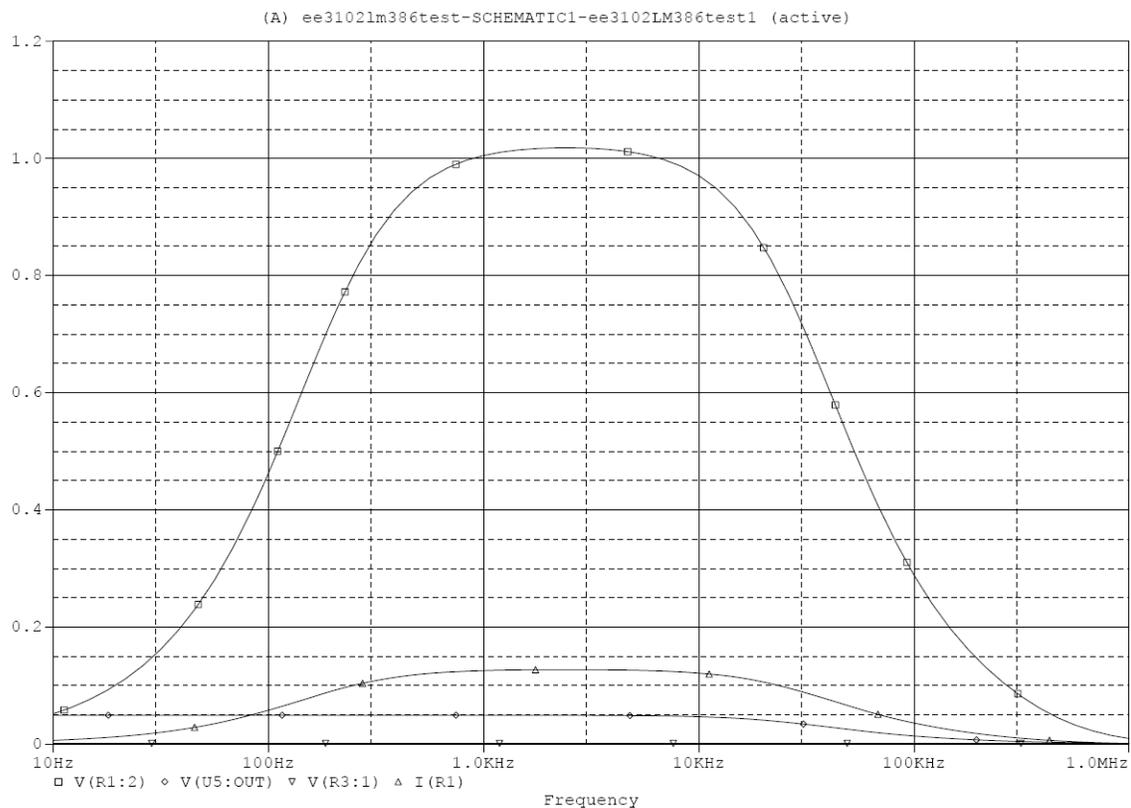


Figure 3: PSpice simulation of the Audio Amplifier subsystem

The mid-band gain of the subsystem was 678, or 56.4dB. The maximum voltage output was 1.018 volts for a 1.5mV input. This value was very close to the predicted mid-band gain of 660. This gain is necessary to produce the 1/8th of a watt power output needed for the output to be audible. The simulation also showed that the subsystem drew 64mA of current from the positive voltage supply. This value is within the limits of the power supply. These results show that the audio amplifier subsystem is predicted to meet the design requirements.

Prototype test results were very promising. The LM386 operated as expected, and the gain was found to be almost exactly 20 when no external components were used, and increasable to 200 is a 10 uF capacitor was placed across the gain pins, shorting the feedback resistor. However, the LM386 drew more current than expected, especially when configured for a larger gain. To counteract this, it was decided to use a LM741 gain stage and allow the LM386 to operate at its minimum gain level. This combination was effective because the total gain of the subsystem could be varied as easily as adjusting the gain of an inverting amplifier. In addition, this two amplifier configuration reduced the current draw of the subsystem while maintaining the required gain.

Subsystem Bill of Materials

Part Number	Supplier	Quantity	Cost	Total Cost	RefDes	Description
LM386N-4-ND		1	0.94	0.94		
LM741CNNS-ND		1	0.84	0.84		
604-1055-ND		2	0.39	0.78		
Storeroom		1	0.1	0.1		
Storeroom		3	0.09	0.27		
102-1186-ND		1	3.5	3.5		
Storeroom		1	0.9	0.9		
Audio Amp Sub Section Total				7.33		

Power Supply Subsystem

Subsystem Specifications

- Power available is standard 110V rms 60Hz
- Power on indicator light

- Power On/Off switch
- Power drawn from source can not exceed 10 watts
- Voltage needed is +15V and -15V

Subsystem Description

The power supply is the most flexible of the subsystems. Due to the power supplies flexibility and the other subsystems limited options, the power supply was designed to meet the requirements of the other subsystems.

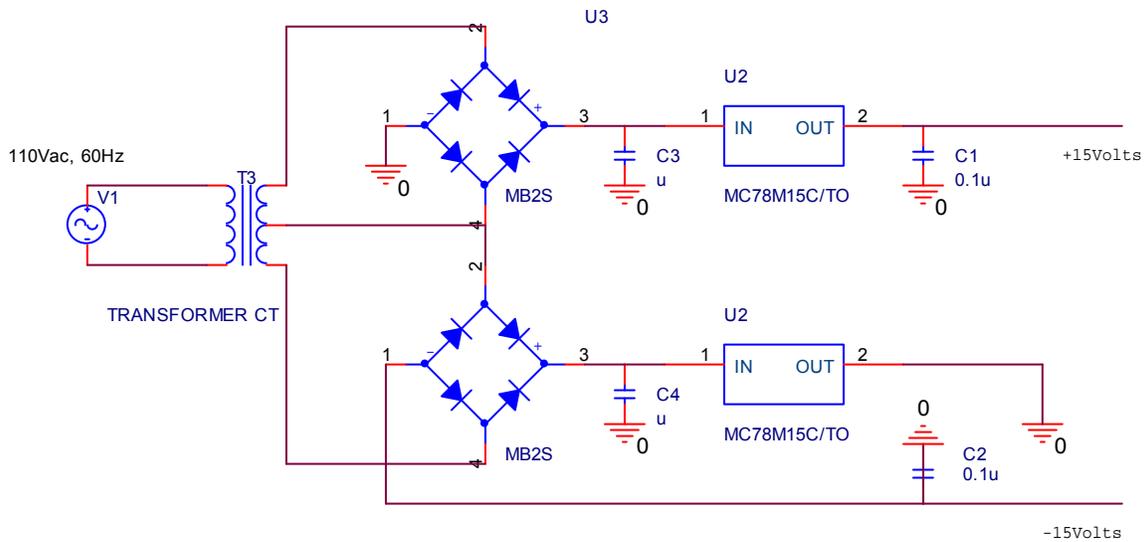
The power supply uses a peak rectifier circuit with a smoothing capacitor. The peak rectifier consists of a diode bridge which operates as follows: During the positive half-cycles v_s will be positive and travel through the load. During the negative half-cycles $-v_s$ will be positive and travel through the load. When both the positive and negative half-cycles travel through the load in the same direction it allows for V_{out} to be always positive. The RF-IF Amplifier required the ± 15 volts so we designed the transformer bridge rectifier circuit to output two positive 20 Vac lines. Then after regulation we invert one of the lines to get the desired negative 15 volts.

The voltage regulator was chosen for its output voltage and its current capability. The amount of current needed was used mostly by the speaker which is about 250mA. Choosing the MC78M15AC gave a 500mA positive voltage regulation with a 15 volt output voltage.

One limitation of using the diode bridge is a voltage drop of 1 volt per leg due to the current passing through two diodes for each half cycle. Also the voltage regulator needed V_{out} to be 2 volts higher than V_{in} . Thus we needed a transformer outputting greater than 18 volts.

A filter capacitor is used for smoothing the output of the rectifier voltage. In order to pick an appropriate capacitor value four things are needed: V_p (peak voltage), V_r (ripple voltage), f (frequency, and R (resistance). The needed capacitor is picked using this equation:

$C = V_p / (V_r * f * R)$. The audio amp uses approximately 250mA of current which gives a resistance of about 60 ohms. With the other subsystems taking very little current and all of the subsystems being in parallel we use the 60 ohms for the resistance value.



Subsystem Bill of Materials

Part Number	Supplier	Qty	Cost	Total Cost	RefDes	Description
595-1139-ND	Digikey	1	10.69	10.69		XFRMR PWR 115V 40VCT 500MA
	Digikey	2				Capacitor
IC-LM7815T	Storeroom	2	0.39	0.78		500mA 15V Regulator
CT-35-.1	Storeroom	2	.21	.42	C1,C2	Resin Dipped 0.1uF 35V

Summary/Conclusions

The circuit appears to function as it stands, and also receives KLBB as designed. There are a few issues that need to be addressed in the following sections to refine the design, and improve the quality of the received signals.

Unresolved Problems

1. Audio amplifier tends to become unstable when the potentiometer is above $\frac{1}{2}$ its range.
2. The audio amplifier tends to distort the signal badly at larger input levels
3. Possible thermal issues with the LM386 amplifier
4. There is a lot of noise on the signal coming from the RF-IF amplifier
5. The AM detector distorts the signal at low levels
6. The RF-Mixer input tuned circuit is not peaking as it should

7. When factoring in all of the external parts and hardware we are getting close to the \$60 limit

Possible Solutions to Problems

1. This is possibly caused by limitations on the input voltage for the LM386 audio amplifier. In order to prevent this, the gain provided by the LM386 and LM741 amplifier stages will have to be altered so as to reduce the input to the LM386 while maintaining the same overall subsystem gain. It may also be possible to add a limiter circuit (consisting of four resistors and two diodes) to reduce the gain of the LM741 stage when a stronger signal is applied.
2. For the audio amplifier subsystem, future work will focus on improving the quality of sound produced by the speaker. When driven using a sinusoidal input test signal, the audio amplifier distorts the top half of the sine wave. This is possibly because the LM741 stage shares its positive voltage supply with LM386, while its negative supply voltage is much less loaded. If this is the cause, then the amplifier stages will need to be better isolated from the power supply.
3. In addition, the power consumption of the subsystem will be addressed. The LM386 currently produces noticeable heat, even when drawing less than 100mA. Over short periods of time, this does not present a problem. However, if operating for an extended period of time in an enclosed container, the amount of heat produced by the LM386 may become significant. One methods of addressing this is to add a heat sink to the LM386. Since this component is responsible for the vast majority of the current draw of the entire system, it is the only device that heats up significantly. However, a heat sink would have to fit within the physical constraints of the system. The audio amplifier subsystem uses several large components such as capacitors, a potentiometer, and a speaker which take up significant space surrounding the LM386. In addition, a small amount of air circulation would have to be added, possibly with several small holes or a vent.
4. The specification for the AD826 RF-IF amplifier is very specific about the layout of the PC board around the part. It stressed the importance of keeping the inputs and outputs as separated as possible, and keeping the feedback resistors close to the device. While the prototype board kept feedback resistors close to the pins

of the part, the small IC mounting boards were just cut from a small PC board with a Dremel tool. These cuts resulted in the two amplifier sections having their pins close together, and this might be the reason for the noise observed. The specification also stressed having both electrolytic and ceramic capacitors directly connected to the power supplies to keep the voltage stable for low and high frequency noise. Further power supply decoupling measure might be required.

5. It appears that the diode is being pulled to some rail limit, and higher input signals pull it away from that limit. Either the resistor across the capacitor is responsible for this pulling, or the RF-IF amplifier offset voltage is pulling the voltage level from ground. More simulation is required to understand the problem fully. One possible solution is to use a capacitor to block the DC offset voltage on the RF-IF amplifier from affecting this circuit. As stated earlier, higher input signals reduce this problem, and the KLBB signal in the lab appeared high enough to work for this application, and therefore the problem might be overstated for this design.
6. A lower impedance of the mixer input might swamp the input tuned circuit, and therefore lowering the Q of the circuit. The specification sheet states that the input resistance is 1.5 K-ohms. The coil is supposed to have a Q of 80, and that would result in a resistance (X_L/R) of 1.055 K-ohms. The 1.5 K-ohms would be in parallel with this value and so would broaden the bandpass to an effective Q of 1.7, or a bandwidth of 820 KHz. See next numbered item for possible solution.
7. Some components could be cut to reduce cost. In particular the tuned input circuit could be removed, as it is not working correctly. This would cut \$5.00 from the circuit cost. Some of the components were priced from the Digi-Key catalog at the low quantity values (higher dollar amount,) and the stockroom's prices are lower as they probably purchase in higher quantities or have an arrangement with Digi-key for a lower price break. Redoing the pricing of all of the items may also reduce the cost

Work remaining schedule

Table 2 shows the remaining schedule to complete the project.

Task	Completion Date
Debug prototype issues	November 3, 2006
Complete PC board layout and order	October 31, 2006
Complete order of parts	October 30, 2006
Populate PCB	November 6, 2006
Assemble and test	November 20, 2006
Test box fabrication completion	November 27, 2006
Final assembly	December 4, 2006
Demo and report	December 11, 2006

Table #: Completion Schedule

It is our feeling that we are fairly far along in the process with a working prototype, even with the instability issues. The PCB layout is a little behind as there was a communication issue between the two individual who were going to work on it. This has been ironed out, and the schematic and preliminary part placement has been completed in two days. A show stopper would be new instabilities due to the PCB layout that are difficult to determine the impact of. We feel very confident that we will be able to complete a working design by the projected due date as outlined above and have a working unit to demonstrate.