Class A Radio Transmitter for Portable Low Power Use

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***Abstract*—We present a radio transmitter circuit that performs Amplitude Modulation (AM) of an audio input using only five discrete NPN Bipolar Junction Transistors (BJT) with all stages being Class A.**

***Index Terms*—Amplitude Modulation (AM), Bipolar Junction Transistor (BJT), Class A, Colpitts Oscillator.**

I. Introduction

In decades past, transistors were rather expensive. Therefore, it was desirable to minimize transistor count in radio circuits designed for mass manufacture. Beyond transistor count, we minimize costs by making our design not rely on exact component values, which allows us to cheaper components with large tolerances.

II. Principle of Operation

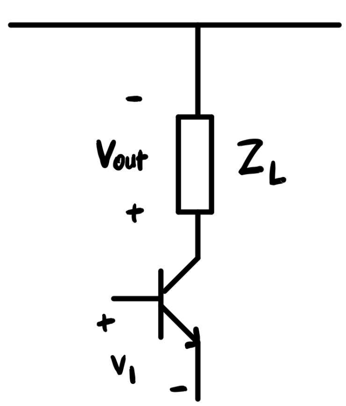

Fig. 1. An AM modulator based on non-linearity of BJT and filtering of unwanted frequency components.

A summing amplifier whose output varies non-linearly with the magnitude of the input can be used as a mixer, thus allowing it to be used for amplitude modulation.

We start with the transconductance relationship for a BJT in the Forward Active Region (F.A.R.):

$$i_c \approx I_s \exp\left(\frac{v_{BE}}{\phi_t}\right)$$

$$= I_s \exp\left(\frac{V_{BE} + v_1(t)}{\phi_t}\right) = I_C \exp\left(\frac{v_1(t)}{\phi_t}\right)$$

which we expand as a power series:

$$= I_C \left[1 + \frac{v_1}{\phi_t} + \frac{1}{2}\left(\frac{v_1}{\phi_t}\right)^2 + \frac{1}{6}\left(\frac{v_1}{\phi_t}\right)^3 + \ldots\right]$$

Given a modulating signal $v_s(t) = A\cos(2\pi f_s t)$ and carrier signal $v_c(t) = -\cos(2\pi f_c t)$, we have:

$$v_1 = v_s - v_c$$

$$v_1^2 = v_s^2 - 2v_s v_c + v_c^2$$

$$v_1^3 = v_s^3 - v_c^3 - 3v_s^2 v_c + 3v_s v_c^2$$

Based on the following trigonometric identities:

$$\cos^2 x = \frac{1}{2}\left[1 + \cos 2x\right]$$

$$\cos^3 x = \frac{1}{4}\left[3\cos x + \cos 3x\right]$$

$$\cos(a)\cos(b) = \frac{1}{2}\left[\cos(a+b) + \cos(a-b)\right]$$

We can determine that the 2nd and 3rd order terms in the power series contribute frequency components of:

- f_s^2 (corresponding to v_s^2)
- f_c^2 (corresponding to v_c^2)
- f_s^3 (corresponding to v_s^3)
- f_c^3 (corresponding to v_c^3)
- $2f_c \pm f_s$ (corresponding to $v_s v_c^2$)
- $f_c \pm f_s^2$ (corresponding to $v_s^2 v_c$)
- $f_c \pm f_s$ (corresponding to $v_s v_c$)

The last pair of frequencies represent the AM sidebands that we are interested in. However, notice that there is also a pair of extraneous sidebands at $f_c \pm f_s^2$ due to the cubic term in the power series. This will manifest as 2nd order harmonic distortion in the received signal. Therefore, this circuit is not an ideal AM modulator. If we analyzed the contributions of the higher order terms (by doing a binomial expansion of $(v_s - v_c)^n$ and looking at the $v_c v_s^{n-1}$ term), we would find additional higher order sidebands. However, the amplitudes of the higher order components are so small that we may consider them negligible.

As long as the load impedance Z_L can filter out the baseband and higher frequency terms while passing the carrier frequency and its sidebands, this circuit will function reasonably well as an AM modulator.

III. Implementation

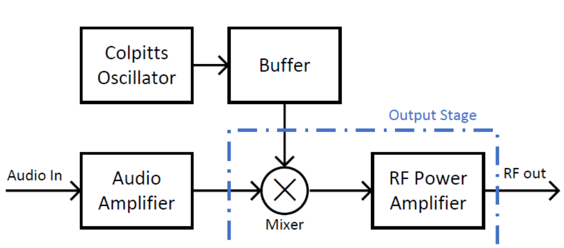

Fig. 2. Simplified block diagram for the circuit.

We implemented a Colpitts Oscillator to provide the carrier signal $v_c(t)$ and an audio amplifier to provide the modulating signal $v_s(t)$.

A. Colpitts Oscillator

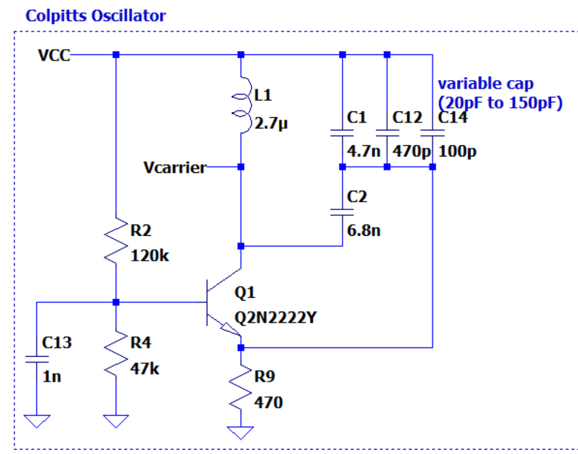

Fig. 3. Schematic for the Colpitts oscillator.

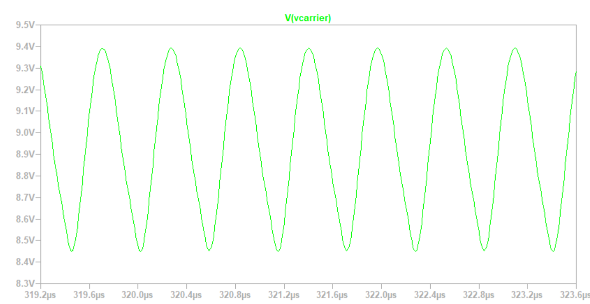

Fig. 4. Output of the Colpitts oscillator.

We implemented a Colpitts Oscillator in a Common Emitter with Degeneration topology. The capacitance to the collector is in parallel with a variable capacitor, so that the oscillation may be tuned manually. In order to minimize the phase noise of the oscillator, we want to maximize the Q factor of the RLC load at the collector of the transistor. This can be done by using a smaller inductance and larger capacitor, as well as a resistance as large as possible, because $Q = \frac{R}{\sqrt{L/C}}$.

Therefore, we did not include a resistor explicitly, but rather rely on the high input resistance of the following stage.

B. Carrier Buffer

In order to minimize loading on the Colpitts Oscillator, which would lower its Q factor, we implemented a buffer (Fig 5). It consists of a Darlington Pair in the Emitter Follower topology. We used a Darlington Pair rather than a single

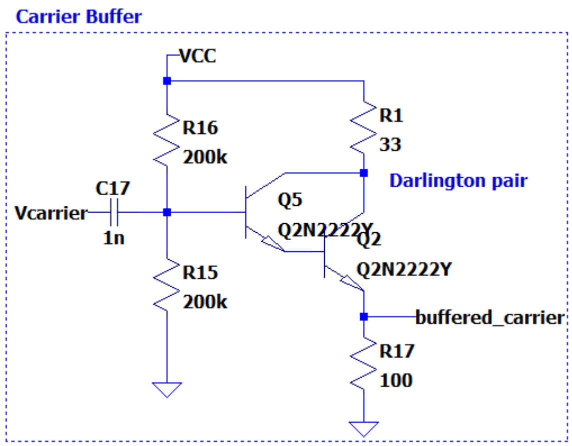

Fig. 5. Schematic for the carrier buffer.

transistor because it provides a current gain of β^2 rather than β. The input resistance looking into the emitter of the output stage is quite small, and this buffer allows us to isolate that low resistance from the carrier and reduce loading. To drive the low input resistance, we need large current gain from this stage, hence the choice of the Darlington pair. The resistor R_1 reduces quiescent power consumption in the transistors, and its value is chosen so as to avoid putting the transistors in saturation while accommodating the maximum signal swing.

C. Audio Amplifier

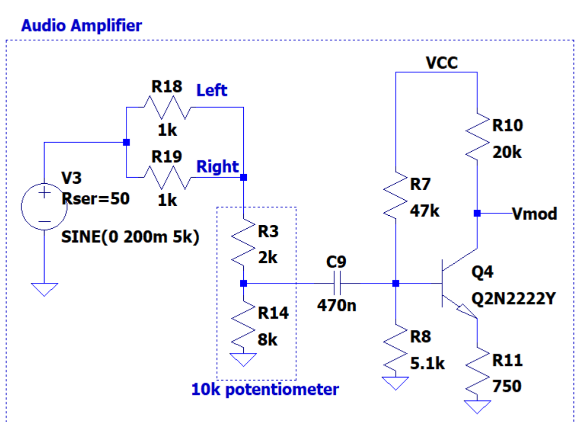

Fig. 6. Schematic for the audio amplifier.

This is a simple audio amplifier topology that allows a variable gain of the form $\frac{-g_m \times R_E}{1+g_m \times R_E}$.

This stage is responsible for the amplitude of the modulating signal, which becomes the modulation index. With the values as in Fig. 6, the gain for this stage is set to 2.5 V/V, with a g_m of 32 mA/V. The input resistance of the power amplifier, which is approximately 50kΩ, appears in parallel with R_{10}, which is a potentiometer. Varying the voltage division by the potentiometer allows us to change the amplitude of the signal that gets amplified. 4.7nF caps are used at the drain and base to filter out any RF that leaks into the audio circuit from the power amplifier.

D. Output Stage

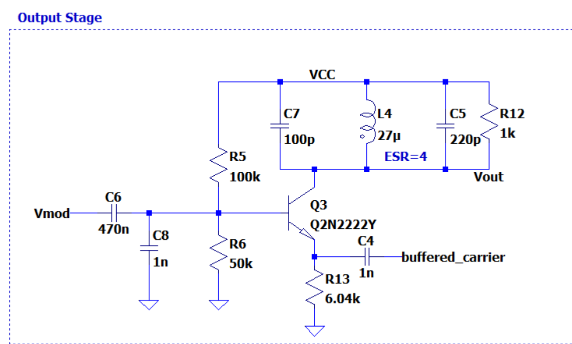

Fig. 7. Schematic for the output stage.

This Class A output stage simultaneously performs mixing and power amplification. The carrier wave is injected into the emitter of transistor Q_3, and the base of Q_3 is RF shorted to ground by C_8, a 4.7nF capacitor whose impedance is high for baseband signals and low for RF signals. The varying voltage at the base varies g_m of Q_3, due to the relationship between v_{BE} and I_E. This g_m modulation effectively modulates the amplitude of the carrier signal at the collector, yielding AM modulation. In the output stage, we want a smaller Q (and therefore larger bandwidth) compared to the oscillator so that it will pass not just the carrier, but also frequencies around it, which is necessary to transmit the sidebands of an AM signal. Given a modulating frequency of up to 5 kHz, we need a bandwidth of at least 10 kHz because of the 2 sidebands.

We have Q = 9.75 here for a bandwidth of 167 kHz, which is too large to pass the FCC emissions test. Perhaps we can improve this by shrinking L_5, so that the load resistor appears as a larger impedance at the primary. However, this would come at the cost of decreased output power. It is important to note that this is only a problem when the output stage is amplifying distorted and spurious signal - if we give a clean sinusoidal signal to the base of Q_3 then there are no undesired emissions.

E. Power Supply

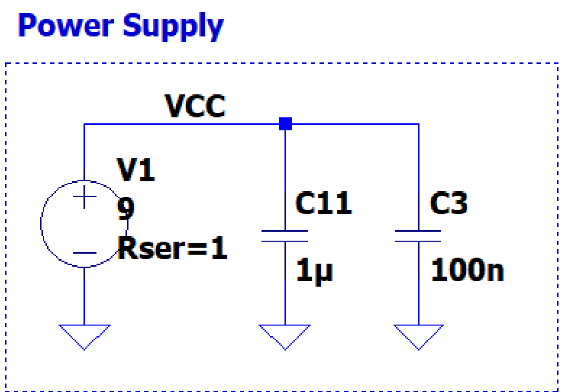

Fig. 8. Schematic for the power supply.

To account for the non-zero internal resistance of the 9V battery, we modeled our supply voltage to have an internal resistance of 1. Left unchecked, this internal resistance would lead to coupling between the audio and RF stages as their current draw changes the supply voltage. Therefore, we added decoupling caps, specifically a 1uF electrolytic and 100nF ceramic. Large electrolytic capacitors tend to have greater parasitics at high frequencies, so the 1uF capacitor provides bulk capacitance for low frequencies, while the 100nF capacitor provides lower impedance for high frequencies.

IV. Measurements

We used a function generator (Agilent 33500B Series) to produce a 5kHz audio test tone. We used a multimeter (Agilent 34450A) to make DC measurements, and an oscilloscope (Keysight DSO-X 4034A) to make AC measurements. The measurements were done with the antenna connected, rather than the 1 kΩ dummy load.

A. Power Dissipation

The DC power dissipation of the output stage was to be no more than 100 mW, in order to meet FCC requirements. The average current drawn by the output stage was 584 uA, meaning that our power stages power dissipation is 5.26 mW, well within requirements. Our carrier frequency was 1.685 MHz, which is slightly higher than the required 1.63MHz.

B. Spurious Emissions

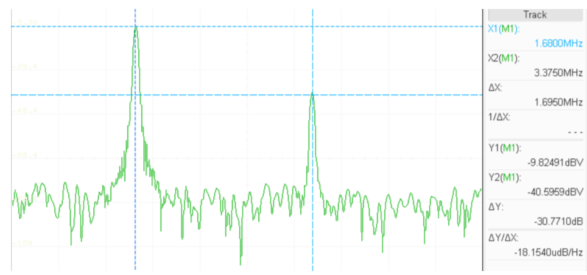

Fig. 9. FFT display at the collector of Q_3.

The FFT of our transmitted signal above shows clearly the second harmonics presence at the output, at a lower power than the fundamental frequency. The second harmonic is 30.8 dB weaker than the carrier wave, meeting the specification of spurious emissions being 20dB weaker than the modulated carrier signal.

C. Audio Passband

The audio circuit only needs to pass frequencies between 20Hz and 20kHz for messages to be fully recognizable. The final audio circuit had a lower cutoff frequency of 23Hz and an upper cutoff frequency of 56kHz.

D. Distortion & Modulation Index

1) Modulation Index: The highest peak-to-peak envelope modulation of the carrier wave that we could achieve without distortion was 2.1 V_{pp}. Any higher envelope would lead to nonlinear modulation. Our unmodulated carrier wave had an amplitude of 4.23 V, putting our maximum achievable modulation index at 49.6%.

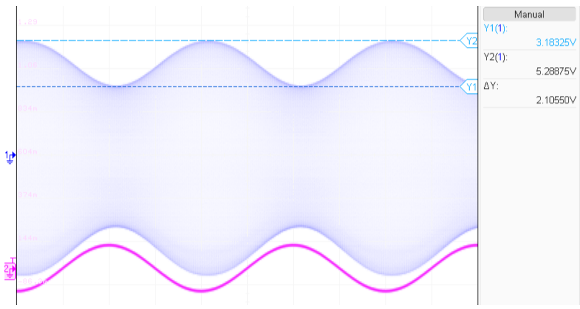

Fig. 10. Output of Q_3 with modulating signal overlaid.

2) Distortion: We were not able to achieve a modulation depth of m=60% without overmodulating and therefore introducing considerable distortion. However, we were able to come quite close with m=49.6% (Fig 10). The envelope of the output closely follows the sinusoidal shape of the input (Fig 10). The 2nd order sideband is 32dB below the fundamental sideband, and all higher order harmonics are below the noise floor. Therefore the harmonic distortion at maximum modulation is 2.5%.

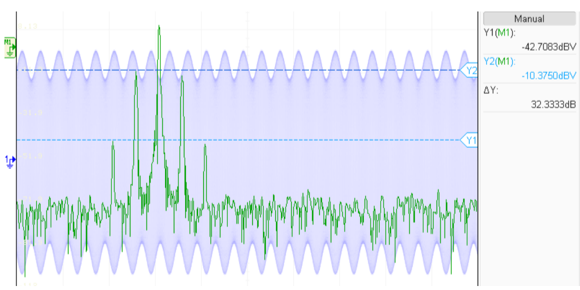

Fig. 11. FFT of signal at collector of Q_3 with modulating signal overlaid.

V. Conclusion

During the demo session, we were able to transmit a music signal with sufficient fidelity to satisfy Professor Tsividis. Our tuning capacitor at our oscillator allowed for broadcasting at a frequency at exactly the frequency received by the demodulator, resulting in a relatively high fidelity signal. Further development of this topology could be done by raising the Q factor of the output stage, decreasing the current draw of the overall circuit (currently 40mA), and perhaps omitting the audio amplifier stage in exchange for an additional gain stage at the output for greater output power with 5 transistors.

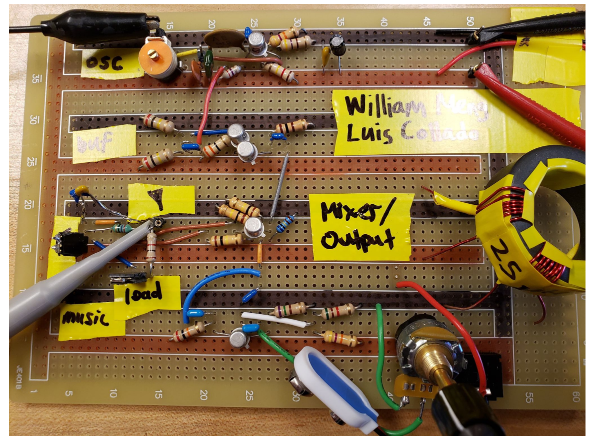

Fig. 12. Photo of complete circuit.

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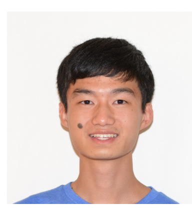

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