Lab 2 Writeup

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1 Abstract

In this lab, two methods of achieving FM modulation were designed and tested for transmission clarity and performance. The first involved a discrete colpitts oscillator, and the second a two stage heterodyned system involving a voltage controlled oscillator (the LM566) with a center frequency at 300kHz and a multiplier (SA612) with a 24 MHz local oscillator to get the signal up to 24.3 MHz. In this write-up the two configurations will be compared.

In brief, the discrete system performed better, bringing better overall quality and signal strength with lower power usage. However, of the two implementations, the discrete colpitts is less frequency stable, and in cases where such stability is of concern, the LM566-SA612 combination should be used.

2 Design Theory

FM, or frequency modulation, involves taking a carrier signal at a frequency f_c and using the input signal (scaled by a factor m_f, the modulation index) to change the frequency around that carrier. The basic FM equation is shown below, and figure 1 shows a time domain representation of simple FM.

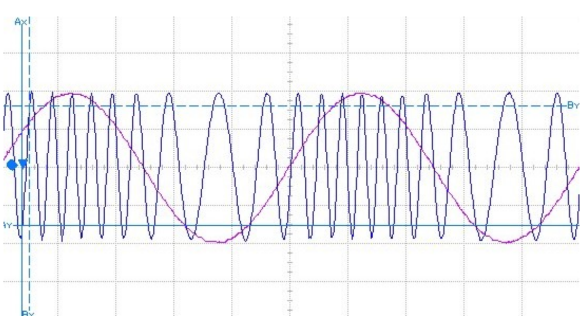

Figure 1: FM Modulation in Time Domain

$$v_0 = A\cos(2\pi f_c t + m_f \sin(2\pi f_m t)) \tag{1}$$

FM can be implemented either directly at radio frequencies or via heterodyning, which involves multiplying a signal of intermediate frequency up to RF. This lab discusses the design and implementation of both methods.

2.1 Colpitts Design

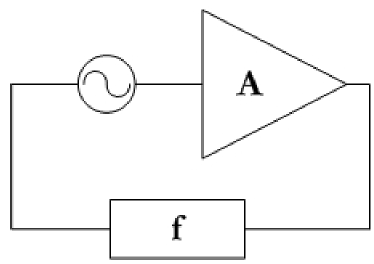

Figure 2: Block Diagram of Colpitts

A colpitts oscillator is simply the combination of a resonator with an active device, implemented using positive feedback to sustain the oscillation. A block diagram of the colpitts technique is shown in Figure 2, below. Transient AC signals inputted to the amplification block are multiplied by Af, the loop gain, and when $Af = 1$ with zero phase shift across the blocks, those sinusoids are sustained indefinitely. For $Af > 1$, their amplitude would grow exponentially, and for $Af < 1$, the sinusoids would simply die away. To select the oscillation frequency, we make the above f block into a LC tank with resonant frequency of $1/\sqrt{\omega LC}$, for at resonance, the phase shift of the tank should be zero. A simple representation of such a system is shown in Figure 3.

Note the variable capacitor–this can be used to change the frequency of the oscillation. This is done in the discrete design using a diode varactor, providing frequency modulation directly at RF. The SA612-based colpitts, on the other hand, relies on the VCO to provide modulation and uses a crystal resonator

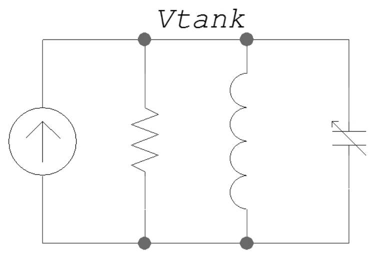

Figure 3: Basic Colpitts with LC tank

to provide a steady oscillation frequency for heterodyning the 300kHz VCO output to 24.3MHz.

2.2 Discrete Design

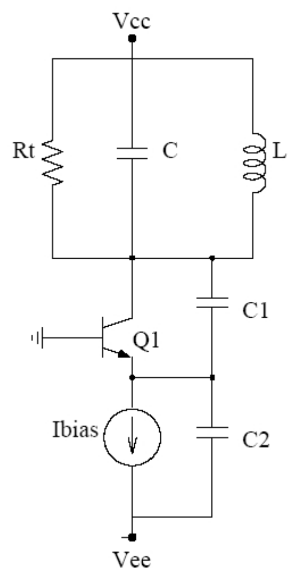

Figure 4: Implemented Colpitts Design

The discrete colpitts was designed basically as shown above. The active component used was a Toshiba 2SC3302 bipolar transistor in common-base configuration. A capacitor divider between collector, base, and emitter was implemented to set the feedback gain f. If we let $n = f = \frac{C_1}{C_1+C_2}$, then a simple expression for loop gain falls out.

$$\frac{g_m n}{G_{tank} + g_m n^2} = Af \tag{2}$$

Because of the non-linear reduction of g_m with large input amplitudes, if we set $Af > 1$, we can be assured that the loop gain will settle at 1 and the oscillator will be stable. We can also estimate the voltage of the oscillations, which appear at the collector according to equation (3).

$$V_{tank} = 2I_C(1-n)R_{tank} \tag{3}$$

In the actual implementation, the C above the collector is made up of a diode varactor in series with a blocking cap, and L is replaced by a tapped inductor configuration to avoid overly loading the Toshiba. This impedance transforms Rt upwards from the expected 50Ω impedance of the spectrum analyzer / antenna / or power amp block to come next. To set the resonance frequency of the tank, the varactor capacitance as well as C_1, C_2, and the C_π of the transistor need to be taken into account, for in small signal AC analysis, all appear at the collector node. To aid in manipulating oscillation frequency, a variable capacitor was placed in parallel with C_2.

With the inductor values chosen, $.1\mu H$ and $1\mu H$ we can estimate what the tank resistance will be assuming a 50Ω load by doing simple impedance transforms and keeping parasitic inductor resistances in mind. I calculated the tank resistance R_{tank} to be $3.6k\Omega$ with a Q of the overall tank to be 21.5. The transformation is shown below.

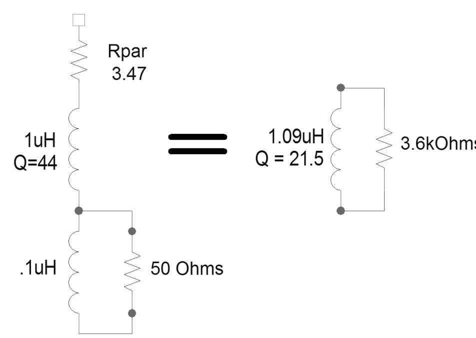

Figure 5: Tank Transformation

With R_{tank} characterized and choice of $\frac{1}{3}$ for n and $I_c = .5mA$, we can use (2) and (3) to find whether the oscillations will sustain and what the amplitude of those oscillations will be. Doing so, the loop gain turns out to be 3, more than enough to sustain oscillations, and the amplitude of V_{tank} to be $2.4V$. $\frac{1}{11}th$ of this will appearing at the 50Ω load, or $.22V$.

$$\omega = \frac{1}{\sqrt{L_{tank}(C_{varactor} + \frac{C_1(C_2+C_\pi)}{C_1+C_2+C_\pi})}} \tag{4}$$

To get a $24.3MHz$ oscillation frequency (4) must be satisfied for $\omega = 2\pi 24.3e6$. We also want n as defined above to be $\frac{1}{3}$. To get near these values, I used $C_1 = 22pF$ plus a variable cap, and $C_2 = 122pF$. C_π is assumed to be $2pF$ in these calculations. The $C_{varactor}$ is dependent on the input capacitance of the 2N3904 NPN transistor, which, when used as a diode, is a function of its V_{bias}, which is set via a potentiometer voltage divider. From the datasheet, I determined its capacitance ranges from $3 - 4.4pF$, effectively changing the frequency of oscillation by about 300kHz, which would be then be our effective transmission bandwidth.

2.3 LM566, SA612

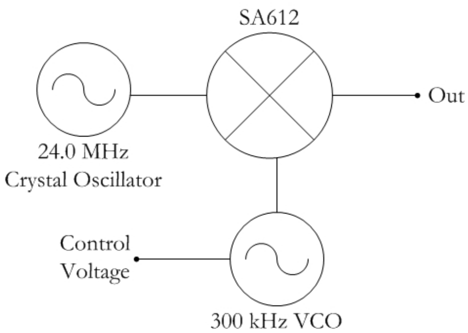

Figure 6: Block Diagram of Colpitts

The structure of the VCO-multiplier transmitter block is shown in Figure 4. The LM566 is a voltage controlled oscillator that can be used to linearly change frequency with input voltage over a specific range, defined by the LM566 k parameter, about $200\frac{kHz}{V}$ from the datasheet for $f_0 = 300kHz$. The VCO is biased to a center frequency via a voltage divider, and then AC signals around that bias voltage change the frequency of oscillation. The design specifics were 100 kHz bandwidth around a center frequency of 300 kHz. To set the center frequency, the LM566 datasheet describes the following equation.

$$f_o = \frac{2.4(V^+ - V_{bias})}{R_o C_o V^+} \tag{5}$$

To set a center frequency of $300kHz$ with the specified V_{bias} of $7.5V$, I chose a $10k\Omega$ potentiometer for R_o and $C_0 = 330pF$ to give me flexibility to get an exact match. With the above k value, I expected to need $+/-.25V$ to achieve $100kHz$ bandwidth.

As shown in Figure 6, this signal is mixed with a $24MHz$ local crystal oscillator on the SA612 multiplier. The design of the LO comes from application note AN1983 and it involves a simple colpitts oscillator with $n = \frac{1}{2}$ using two $33pF$ capacitors that act similar to the C_1 and C_2 in the discrete case. When the LO is multiplied with the VCO out, this should heterodyne the FM modulated signals to 24.3 and $23.7MHz$ with some figure of gain over the VCO output due to the mixer's conversion gain (g_{conv}). The $24MHz$ LO signal will also appear at the multiplier output, though suppressed some.

3 Simulation and Final Results

In this section, the characterization of the two implementations is highlighted and compared.

3.1 Discrete

The discrete colpitts was built and found to output at $-3dBm$ while consuming $63mW$. The results of pre-lab calculations compared with HSpice simulations and final lab runs are shown in Table 1 and the frequency vs. varactor-diode bias plot is shown in Figure 7.

Result	R_{bias}	V_{tank}	V_{load}
Prelab	$6.8k\Omega$	$2.4V$	$0.22V$
HSpice1	$6.8k\Omega$	$2.42V$	$0.242V$
HSpice2	$1k\Omega$	$6.82V$	$0.594V$
Lab1	$6.8k\Omega$	$2.15V$	$0.09V$
Lab2	$1k\Omega$	$9V$	$0.8V$

Table 1: Discrete Tank and Load Voltages

As Table 1 shows, the HSpice calculations agreed nicely with the pre-lab calculations, while the in-lab numbers varied somewhat, with the final configuration, with 1k bias resistor greatly outperforming the HSpice numbers. In all cases, as expected, a smaller bias resister leads to higher I_c and thus greater voltage amplitudes at the tank, as (3) suggests.

In Figure 7, k is shown to be rather linear and about $180kHz/V$, though, as the curve is steeper closer to the center voltage (set to be $7.5V$), 100kHz bandwidth could be achieved with $+/-250mV$ voltage increments. The max bandwidth of the implementation was found to be $120kHz$, rather than the expected $300kHz$. This is due to attenuation by a $10k\Omega$ resistor in series with the input audio signal. We wish to operate the receiver with $100kHz$ bandwidth, so this attenuation is not a problem–rather it is a design benefit.

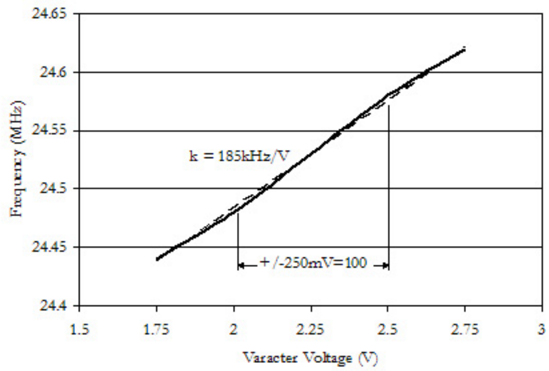

Figure 7: Colpitts Results

When connected to an $11'$ antenna, the colpitts performed well albeit somewhat jittery. I found I was able to (and had to) tune the colpitts frequency rather often, though it was easily done with the variable cap. In fact, the variable cap gave me the flexibility to achieve carrier frequencies between 26.93 and $23.83MHz$. The spectrum analyzer at my bench registered $-25dBm$ on it's $11'$ antenna and the sound quality was very high if not perfect. When I connected the receiving antenna to an analyzer 20 feet away, the power had dropped to $-45dBM$ but the signal quality was still high.

3.2 Chip-based

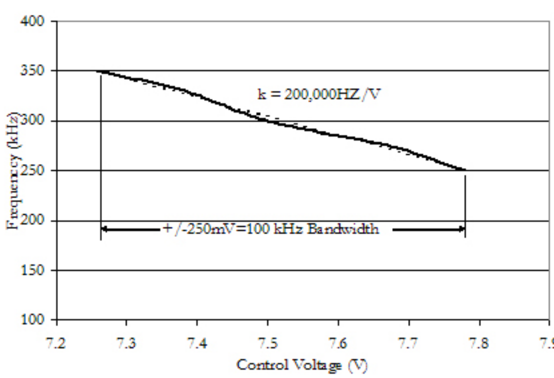

Figure 8: VCO Results

The SA612-LM566 combination performed stably with output power of $-18dBm$ and consumption of 90mW. Most of the output power losses associated with this design probably fall in the lack of reactive components in the VCO–the output square wave is max at 5Vp-p, compared to the 18Vp-p output wave of the inductor/capacitor combination in the discrete tank. The k, as the datasheet lists, turned out to be near exactly $200kHz/V$, and $+/-250mV$ provided the desired $100kHz$ bandwidth. The crystal in the SA612 LO was not entirely clean, with harmonics at every multiple of the fundamental, of power $-45dBm$ @ $48MHz$, $-50dBm$ @ $72MHz$ and so on.

Perhaps because of these harmonics or other third order multiplication artifacts, the output of the SA612-LM566 combination was of lower quality than the discrete colpitts. Power wise, at my bench I received the signal on my analyzer at $-38dBm$ and about 20 feet away at $-50dBm$. Though at lower power and with crackly sound quality, the chip-based design only needed to be tuned once and was quite stable.

4 Conclusion

Both the chip-based and discrete implementations of FM modulation at $24.3MHz$ have strengths and weaknesses. The discrete design offers more output strength with less power consumption, though with the downside of needing constant tuning to achieve the proper carrier frequency. The SA612 and LM566 combination provides stability for $30mW$ more power, though quality and output power seemed to suffer in tandem with the exchange. Certainly, at higher frequencies the SA612-LM566 combination may be necessary to achieve the needed stability. However, for our frequency range, the more compact and presumably lower-cost discrete option performed well, and thus I have chosen to use it in the later stages of our SPAM project.