 March 8, 1995

A Look at Video Amplifier Peaking Networks

Jim Hagerman, *Member of the Technical Staff*, Hughes-JVC Technology Corporation

Abstract **- There are many options available to the video amplifier designer who wishes to maximize circuit performance. This article looks at the use of different peaking networks to minimize risetime for a given load resistor. Also introduced is a new simplified version of a bridged T-coil.**

I. INTRODUCTION

There are several topologies in use for video output amplifiers. Each has its own trade-offs for cost, complexity, and performance. This article will deal with the version shown in Figure 1. Although this specific design will be followed many of the concepts can be applied to other circuits and applications. Each section of the amplifier is analyzed and optimized for bandwidth.

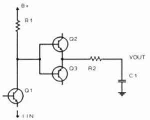

Figure 1. Basic amplifier topology

II. BASIC AMPLIFIER

The basic amplifier uses a cascode output transistor Q_1 to minimize the Miller capacitance effect on the input signal. The output current is fed into a load resistor to convert the signal to a voltage. This is then buffered by emitter followers for driving the capacitive CRT load C_1. R_2 is needed for arc protection. Many CRT drivers used today use this architecture. Biasing components for Q_2 and Q_3 are not shown since they should not affect dynamic performance.

Implied, but not specifically shown are the many parasitics which degrade performance at high frequencies. The output transistor Q_1 has collector-base capacitance, R_1 may be both inductive and capacitive, Q_2 and Q_3 have collector-base capacitances, and the wiring to the CRT has an inductive characteristic. Even the PCB traces will have stray inductance and capacitance. Additional arc protection components may add capacitance to some nodes. There are other parameters such as transistor f_t which can affect bandwidth but these will be assumed to have already been optimized.

III. LOAD RESISTOR

Recent advances in resistor technology have put a non-inductive resistor element in a TO-220 style package. This is very convenient because it allows for smaller packaging of the electronics (an undisputed trend). It also generates less EMI because of it's smaller antenna size. The main problem with these parts is that they can have a high element to substrate capacitance. This type of resistor can be modeled as shown in Figure 2. Although not entirely accurate because the capacitance is a distributed rather than a lumped element, this model will serve adequately for our purposes.

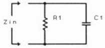

Figure 2. Equivalent circuit of load resistor.

The capacitance is undesireable because it will contribute to the rolloff of high frequencies. It can, however, be compensated for by the addition of a simple R-L network as shown in Figure 3.

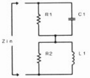

Figure 3. Load resistor compensation.

The impedance looking into this network Z_{in} is given by

$$Z_{in}=\frac{s^2(R_2R_1L_1C_1)+s(R_2L_1+R_1L_1)+R_2R_1}{s^2(R_1L_1C_1)+s(L_1+R_2R_1C_1)+R_2}.$$

If $R_2 = R_1 = R$, then we get

$$Z_{in}=R\left[\frac{s^2RL_1C_1+s2L_1+R}{s^2RL_1C_1+s(L_1+R^2C_1)+R}\right].$$

It can be seen by inspection that the numerator will equal the denominator if

$$2L_1=L_1+R^2C_1.$$

Therefore, if

$$L_1=R^2C_1$$

then the impedance reduces to

$$Z_{in}=R$$

at all frequencies. This load resistor compensation is key to making the following peaking networks function properly.

IV. STANDARD PEAKING NETWORKS

This section focuses on the amplifier from the cascode output transistor to the input of the emitter follower stage. Because of the large voltage swing required to drive a CRT a great amount of power must be dissipated in the load resistor. To reduce this power the resistor value must increase thereby decreasing bandwidth. Hence, this part of the amplifier has the greatest contribution to bandwidth reduction. Adding a peaking network is critical for preservation of bandwidth.

Figure 4 shows a simplified electrical model of our original amplifier without the load resistor compensation. Bandwidth rolloff is a simple R-C time constant where the capacitance is the total of the resistor body capacitance, the output transistor collector-base capacitance, the capacitance of the input to the push-pull emitter follower, and the reverse transfer capacitance due to the load seen through the buffer stage. An uncompensated amplifier will be relatively slow.

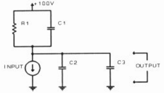

Figure 4. Uncompensated amplifier electrical equivalent.

Each peaking network described in the following sections will be optimized for step response with an overshoot of 5% or less. We will use somewhat realistic but arbitrary component values of

$$R_1=330\Omega$$
$$C_1=5pF$$
$$C_2=5pF$$
$$C_3=5pF.$$

4.1 Load Resistor Compensation

Using the load resistor compensation developed in the previous section the equivalent circuit becomes as shown in Figure 5. Applying the formula developed in the previous section we get

$$L_1=0.54\mu H.$$

The new time constant is based on a capacitance of only C_2 and C_3.

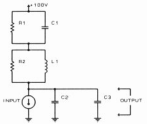

Figure 5. Equivalent circuit with load resistor compensation.

4.2 Shunt Peaking

The simplest method of peaking is either the series peak or the shunt peak. The results are roughly equivalent so only one will be considered. Shunt peaking is done by adding an inductor L_2 as shown in Figure 6.

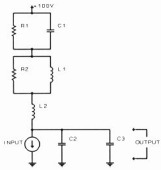

Figure 6. Equivalent circuit for shunt peaking.

For an overshoot of about 5% it can be shown that

$$L_2 = \frac{R_1^2(C_2+C_3)}{2.3}$$

which gives

$$L_2 = 0.47\mu H.$$

4.3 Series-Shunt Peaking

A more complex peaking network is that of the series-shunt peak as shown in Figure 7. This takes advantage of the fact that the capacitances C_2 and C_3 are physically separated. This circuit performs peaking on each of these capacitances but gives best response, however, when $C_2 = C_3$.

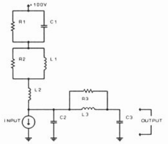

Figure 7. Equivalent circuit for series-shunt peaking.

The equations for calculating component values are given by

$$L_2 = \frac{R_1^2 C_2}{2}$$
$$L_3 = R_1^2 C_3$$
$$R_3 = 2R_1$$

which give

$$L_2 = 0.27\mu H$$
$$L_3 = 0.54\mu H$$
$$R_3 = 660\Omega.$$

4.4 Comparison of Standard Peaking Networks

Each of the above circuits were simulated in SPICE. Figure 8 is a plot of the results. The initial amplifier has a very slow R-C time constant as expected with the compensated R-C a bit faster. The shunt and series-shunt peaking networks are both significantly faster, each with a well behaved rising edge and settling response. Relative to the compensated R-C risetime, shunt peaking is 1.8 times faster and series-shunt peaking is 2.1 times faster. These are significant improvements.

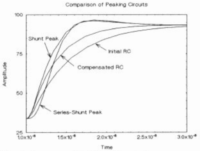

Figure 8. Step response of standard peaking networks.

V. T-COIL

Many years ago Tektronix developed a unique peaking network which has some outstanding qualities. This network is called the bridged T-coil and is shown in Figure 9. The inductors L_1 and L_2 are loosely coupled magnetically to form a transformer with polarity as shown. This network drives a load consisting of a series resistance R_S, inductance L_S, and a capacitance C_S. The output voltage is taken across C_S. This will fit into our application very nicely ($L_S = R_S = 0$).

The bridged T-coil has the property of having a perfectly resistive input impedance Z_{in} at all frequencies. This effectively isolates the output of one amplifier from the input of the next.

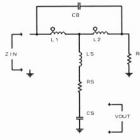

Figure 9. Bridged T-coil.

The solution to this network is rather difficult to derive and was kept a trade secret at Tektronix for many years. Thankfully they have shared the design equations with the rest of the world[1]. They are

$$L_1=\frac{C_L}{4}\left[1+\frac{1}{4\zeta^2}\right](R_L+R_S)^2-R_LR_SC_L-L_S$$
$$L_2=\frac{C_L}{4}\left[1+\frac{1}{4\zeta^2}\right](R_L+R_S)^2-L_S$$
$$M=\frac{C_L}{4}\left[R_L^2-R_S^2-\frac{1}{4\zeta^2}(R_L+R_S)^2\right]+L_S$$
$$C_B=\frac{C_L}{16\zeta^2}\left[\frac{R_L+R_S}{R_L}\right]^2$$
$$k=\frac{M}{\sqrt{L_1L_2}}.$$

where ζ is the damping coefficient and k is the coupling coefficient. For an overshoot of about 5% ζ is 0.69. Other values of overshoot can be obtained where the relationship is given by

$$P.O.=e^{\left(\frac{-\zeta\pi}{\sqrt{1-\zeta^2}}\right)}\times 100\%.$$

5.1 T-coil Peaking

Applying the bridged T-coil to our circuit we get the circuit shown in Figure 10. This network isolates the load C_3 from the cascode transistor current source and appears as just a resistive load of value R_1. Therefore, the voltage step response across C_2 will be an R-C of just R_1 and C_2. This is the input voltage to the T-coil. The output voltage across C_3 will probably not overshoot because the input voltage risetime is still slow and undershot. Of course, if the input voltage was a perfect step then the output voltage across C_3 would have the 5% overshoot (damping of 0.69).

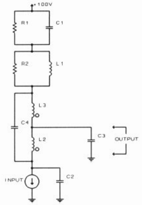

Figure 10. Equivalent circuit using T-coil.

Using the equations given above the component values are calculated as

L_2=0.21μH
L_3=0.21μH
C_4=0.65pF
k=0.31.

To overcome the remaining capacitive component C_2, we can add inductive peaking as shown in Figure 11 by the addition of L_4. This will add some peaking to the input of the T-coil and should give a faster overall response.

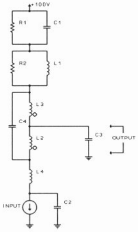

Figure 11. Equivalent circuit using T-coil with peaking.

The value of this inductor can be calculated same as a shunt peak by

$$L_4 = \frac{R_1^2 C_2}{2.3}$$

which gives

L_4=0.24μH.

This is only approximate and will probably require trial and error to obtain the desired response.

5.2 Comparison of T-Coil & Standard Peaking

The T-coil circuits were simulated using SPICE. Figure 12 is a plot of the output waveforms. They are compared with series-shunt peaking and the compensated R-C.

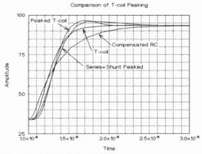

Figure 12. SPICE simulation of T-coil.

The T-coil alone does indeed remain undershot as expected but is corrected with the addition of inductive peaking. Compared to the compensated R-C, the peaked T-coil risetime is 2.5 times faster with a greatly reduced settling time.

5.3 *Simplified T-coil*

The T-coil network as defined in the previous section can be difficult to build and tune. The two inductors are magnetically coupled forming an auto-transformer. As previously mentioned this coupling can be rather loose so winding such a high frequency transformer with any repeatability is no easy task. Constructing a manufacturable T-coil network would be much easier if there was no magnetic coupling. Fortunately, this is possible.

We can derive a simplified T-coil by setting the equation for mutual inductance M equal to zero. All of the parameters are fixed except for the damping coefficient ζ. Therefore, there exists some ζ where M is zero. Solving the equation for ζ shows that it is a function of R_S, R_L, L_S, and C_L.

$$\zeta=\frac{(R_L+R_S)}{2}\cdot\sqrt{\frac{1}{R_L^2-R_S^2+\frac{4L_S}{C_L}}}$$

If Rs was variable then we could set ζ to a desired value (for a given overshoot). Adding a resistance R_P in series with R_S gives us that control. See Figure 13.

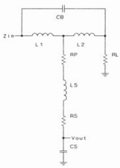

Figure 13. Author's simplified T-coil.

Adding R_P and solving we get

$$R_P=\left(\frac{\sqrt{R_L^2-(1+4\zeta^2)\cdot\left[R_L^2\cdot(1-4\zeta^2)-\left(\frac{16\zeta^2L_S}{C_L}\right)\right]}-R_L}{(1+4\zeta^2)}\right)-R_S$$

The equations for L_1, L_2, and C_B are the same as before but with R_S replaced by $(R_P + R_S)$. Values for the given application are calculated as

$$R_P=103$$
$$L_1=0.19\mu H$$
$$L_2=0.36\mu H$$
$$C_B=1.1pF.$$

The performance of the simplified T-coil network is identical to the original. However, under some circumstances the resistor R_P may calculate as negative or imaginary, in which case the simplified form is unrealizeable.

VI. PUSH-PULL BUFFER PEAKING

The buffer section must now be considered. In Figure 1 both Q_2 and Q_3 form a push-pull emitter follower buffer which has a low output impedance to drive the CRT load capacitance. The series resistor R_2 is for arc protection and damping. This section of the amplifier is modeled in Figure 14. The equivalent circuit includes a series inductance L_1 which simulates the effect of the connecting wire to the CRT electron gun.

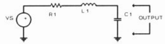

Figure 14. Equivalent circuit of push-pull section.

This circuit can be transformed to it's Norton equivalent shown in Figure 15, and takes the form of a series peaking network.

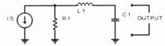

Figure 15. Norton equivalent of push-pull section.

This circuit can be tuned using the parasitic inductance L_1 in the wire and the damping (arc) resistor R_1. Realistic values for these components are

$R_1 = 100\Omega$
$L_1 = 0.05\mu H$
$C_1 = 10pF.$

Hopefully this network has a much smaller risetime than the previous section so as not to degrade the overall performance too much. Using the above values, a SPICE simulation was done resulting in a risetime of 1.7ns with 4.8% overshoot, and 3.7ns settling time. These arbitrary component values just happened to work out quite nicely. They can be modified to improve performance but care must be taken not to reduce R_1 too much or the amplifier may be damaged by arcs. It must be noted that arc protection requires additional components such as clamping diodes, spark gaps and surge protectors. These however are modeled by lumping their capacitances into the equivalent circuit. If possible, it is best to try and get the components of Figure 14 to fit the relationship

$$L_1 = R_1^2 C_1.$$

Indeed in some situations performance can be enhanced by increasing the length (inductance) of the wire to the CRT.

The final configuration of our CRT amplifier using a simplified peaked T-coil is shown in Figure 16.

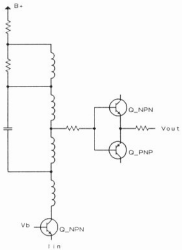

Figure 16. Final circuit configuration.

VII. OVERALL RESULTS

To simulate the overall response of the amplifier the SPICE outputs from the peaking networks were used as SPICE inputs to the push-pull section. This was done for each peaking network. Figure 17 shows the two best responses, the peaked T-coil and the series-shunt peaking. Both had very clean transitions with close to our desired 5% overshoot. The peaked T-coil settling time was much less.

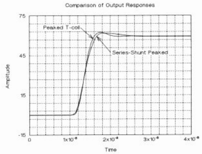

Figure 17. Overall step response for best two networks.

The results of the compensated R-C, shunt peak, series-shunt peak, and the peaked T-coil are compared in Table 1 in order of performance. Settling time is defined as the time from the 50% crossing to within 1% of final value. Bandwidth is -3dB and defined by

$$BW = \frac{0.35}{risetime}.$$

Each network can be further tuned closer to 5% overshoot but the relative risetime reduction will be insignificant. The results speak for themselves. A decent peaking network can greatly increase the bandwidth of an amplifier with no increase in power.

Peaking Network	Risetime	Overshoot	Settling Time	Bandwidth
Compensated R-C	7.2ns	0%	12.7ns	49MHz
Shunt	4.3ns	3.8%	10.4ns	81MHz
Series-shunt	3.9ns	3.3%	9.1ns	90MHz
Peaked T-coil	3.3ns	4.7%	6.1ns	106MHz

Table I. Overall performance results.

The overall response of the peaked T-coil plus buffer section is not quite 2.5 times better than the compensated R-C because of the effects of the buffer section.

VIII. CONCLUSION

Several peaking networks for video amplifiers were analyzed and compared. A new simplified form of the bridged T-coil was given. The design equations were given for each case. All networks were then simulated in SPICE to compare performance. Results showed that the peaked T-coil gave the best step response being significantly better than the equivalent R-C.

IX. REFERENCES

1. "Analog Circuit Design, Art, Science, and Personalities," Jim Williams, John Addis, Ch. 14, Butterworth-Heinemann, 1991.

2. "Handbook of Analog Circuit Design," Dennis Feucht, Academic Press, 1990.

X. ABOUT THE AUTHOR

Jim Hagerman is a senior engineer at Hughes-JVC Technology where he has worked for the past 4 years designing video projector circuits. Prior to this he was at Tempo Research and Digital Equipment Corp. He earned a BSEE from the University of Minnesota in 1982. Most of his spare time is spent enjoying his 0.5 year old daughter and playing soccer. E-mail: 72230.1704@compuserve.com.