Brandon

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R.F. Note #97

J. Vincent
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Contents:

1.

Loose Coupling

In the past (I have heard) amplifiers were coupled onto a transmission line through a network which is much like a bandpass filter. This network allowed only the desired frequency signal to pass and, I believe, matched the amplifier to the characteristic impedance of the line. Now we couple onto the line in a non-resonant fashion through a single capacitor. This capacitor is set to maintain a calculated ratio of amplifier resonator voltage to line voltage. The line voltage is set equal to $V_L = \sqrt{Z_0 W}$ where $W = W_L + W_D$, W_L = line losses and W_D = Dee resonator losses. The amplifier resonator voltage is set to maintain a certain amount of circulating energy.

Although I have no desire to change the output coupling scheme, I will describe some pitfalls associated with it.

1. This scheme assumes all power launched onto the transmission line doesn't return.

2. The desired ratio of resonator voltage to line voltage is only valid for the fundamental frequency.

Number one isn't much of a problem if the harmonic content is low and if the impedance combination of transmitter, line, and Dee isn't resonant at F or any nF. If resonances exist, we can dampen the line with a parallel resistor, capacitor combination as was done for the K500. In order to do this though, adequate power must be available. This extra power may not be available on the K800.

Number two causes a bigger problem. The transmitter and driver stages operating in class AB naturally cause the whole spectrum of harmonics. Although all of these can be seen on a spectrum analyser, the most pronounced ones will be the odd harmonics. This is due to the nature of the shorted coaxial resonant cavities the tubes couple to. These harmonics will be more pronounced on the line then they were in the cavity due to the hi-pass nature of the capacitive output coupling used.

$$\frac{V\ line}{V\ amp} = \frac{Zo}{\sqrt{Zo^2 + (1/WC_C)^2}} \angle -\tan^{-1}\left(\frac{1}{WC_C Zo}\right)$$

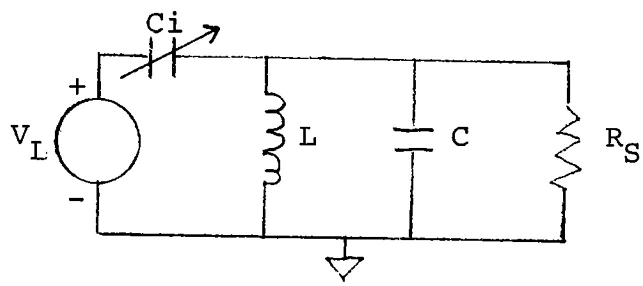
$$\lim_{W \to \infty} \left(\frac{V\ line}{V\ amp}\right) = 1 \angle 0^\circ$$

Because neither the Dee cavity nor transmitter output impendance is matched at any harmonic, the energy contained in the unwanted higher harmonics just rebounds around. This energy is dissipated by the losses in the line and whatever portion couples back into the cavities. How much couples back into the cavities and is dissipated there is not a trivial question. I will be taking transmitter output impedance and Dee cavity input impedance measurements through 5Fo for each Fo we measure. A computer data file will be generated to attempt to estimate the impedances for all frequencies so transmission line resonances may be predicted.

The high harmonic content on the line leads to messy waveforms; and mixer/phase detectors, mixer/peak detectors, etc., must be used to get regulation signals suitable for control of the fundamental. Future notes will speak of mixer/phase detectors and phase shifters, and will show responses and possibly improved designs. I can state that an unstable pole exists now that we attempt to conceal with a zero (not good procedure). The cause of this instability has not been pinpointed yet, but it will be.

Calculation of the necessary input capacitance and associated frequency shift to match the dee cavity to transmission line is a bit more complex. I will include my calculations for future reference.

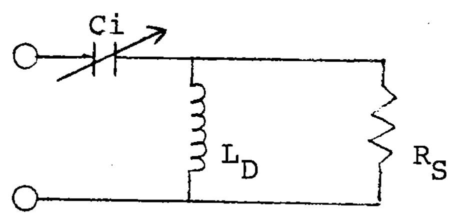
At any given Fo we can, through superfish and MV800, calculate the equivalent capacity, inductance, and resistance. We can then model the dee resonator as a parallel tuned circuit of lumped elements at Fo.



where: $$Q = \frac{W_o U_e}{Pd} = \frac{W_o U_m}{Pd}$$

$$R_S = \frac{V^2 \text{ Dee}}{Pd} \qquad C = \frac{Q}{W_O R_S} \qquad L = \frac{R}{W_O Q}$$

I first assume we tune off resonance slightly to the inductive side.



$$Zin = \frac{R_S (WL_D)^2}{R_S^2 + (WL_D)^2} + J \frac{R_S^2 Ci W^2 L_D - R_S^2 - W^2 L_D^2}{R_S^2 W Ci + W^3 L_D^2 Ci}$$

For a match, we require

$$Zin \text{ (real)} = Zo$$

$$Zin \text{ (imag.)} = 0$$

Imag: (1) $$W^2 R_S^2 L_D Ci - R_S^2 - W^2 L^2 = 0$$

Real: (2) $$(WL_D)^2 R_S = Z_0 R_S^2 + Z_0 (WL_D)^2$$

From (2) $$W^2 L_D^2 (R_S - Z_0) = Z_0 R_S^2$$

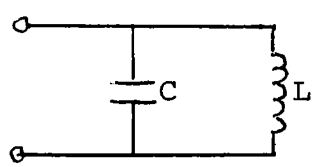
$$WL_D = R_S \sqrt{\frac{Z_0}{R_S - Z_0}}$$

From (1) $$R_S W Ci \sqrt{\frac{Z_0}{R_S - Z_0}} - 1 - \frac{Z_0}{R_S - Z_0} = 0$$

$$WC_i = \frac{1}{R_S} \sqrt{\frac{R_S - Z_0}{Z_0}} \left[1 + \frac{Z_0}{R_S - Z_0} \right]$$

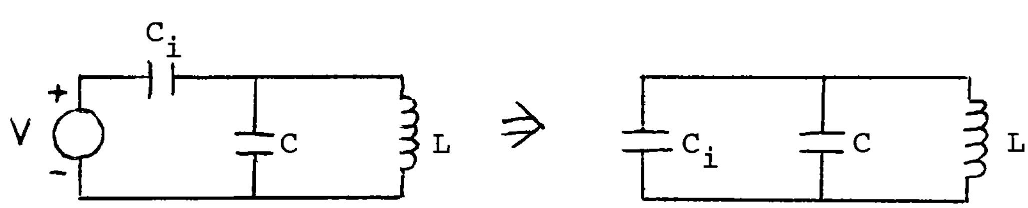
$$C_i \approx \frac{1}{W \sqrt{R_S Z_0}} , \quad R_S >> Z_0$$

Now to simplify the analysis I will drop R_S for the resonant frequency calculations.



$$W_0 = \frac{1}{\sqrt{LC}}$$

And cheating once again:



$$W^1_0 = \frac{1}{\sqrt{L (C+C_i)}} \qquad \Delta W = \frac{1}{\sqrt{LC}} - \frac{1}{\sqrt{L (C+C_i)}} \qquad (4)$$

From (3) we would like: $$\frac{WL}{1-W^2LC} = R_S\sqrt{\frac{Z_0}{R_S-Z_0}} \qquad (5)$$

and combining (4) & (5), $$\Delta F \approx \frac{1}{4\Pi R_S C}\sqrt{\frac{R_S-Z_0}{Z_0}} \approx \frac{1}{4\Pi C\sqrt{R_S Z_0}}, \quad R_S \gg Z_0$$

Furthermore, if one wished to get real picky, it can be shown that we must actually run 1 or 2 kHz off the new resonance to couple correctly.

As an example, taking some data from Jack's K500 programs:

$$@\ F = 20\ \text{MHz} \qquad C = 428\ \text{pf}$$
$$L = 148\ \mu\text{H}$$
$$R_S = 111.\ \text{K}$$

$$\Delta F = \frac{1}{4\Pi C\sqrt{R_S Z_0}} = 64.5\ \text{KHz}$$

$$F_0^1 = F_0 - \Delta F = 19.936\ \text{MHz}$$

$$C_i = \frac{1}{2\Pi F_0\sqrt{R_S Z_0}} = 2.77\ \text{pf}$$

Jack's program states $C_i = 2.83$ pf, which is not a bad approximation to the correct number above!

For 100 kV on the resonator

$$W = \frac{(100\text{kV})^2}{R_S} = 90.09\ \text{kW}$$

$$V_L = \sqrt{WZ_0} = 2.6\text{kV}$$

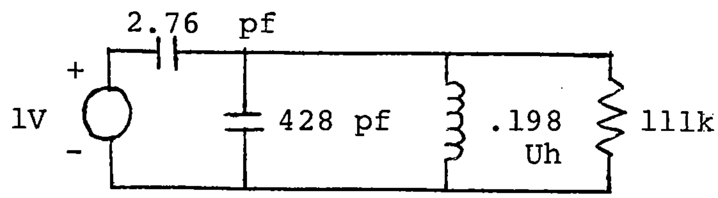
$$G = \text{resonator gain} = \frac{100\text{kV}}{2.6\text{kV}} = 38.46 = \frac{V\ \text{Dee}}{V_L}$$

Appendix I has a spice output which shows for this case:

$G = 38.37$

$F_0^1 = 19.9312$ MHz $\quad Z_{in} = 75.36$

For the circuit:



To summarize:

C_i = input coupling capacitance value

C = equivalent Dee capacitance

L = equivalent Dee inductance

R_S = equivalent shunt resistance due to losses

F_0^1 = the adjusted operating frequency

F_0 = the non-coupled operating frequency

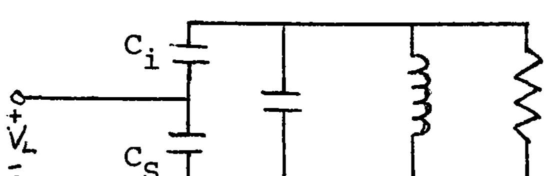
P_D = power dissipated in the dee cavity

$$F_0^1 = F_0^0 - \Delta F \qquad R_S = \frac{V^2\text{Dee}}{P_D} \qquad Q = \frac{W_0 U_e}{P_D} = \frac{W_0 U_m}{P_D}$$

$$C = \frac{Q}{W_0 R_S} \qquad L = \frac{R}{W_0 Q} \qquad C_i \cong \frac{1}{W\sqrt{Z_0 R_S}}$$

$$\Delta F \cong \frac{1}{4\Pi C\sqrt{R_S Z_0}}$$

There is another possibility of coupling into the Dee:

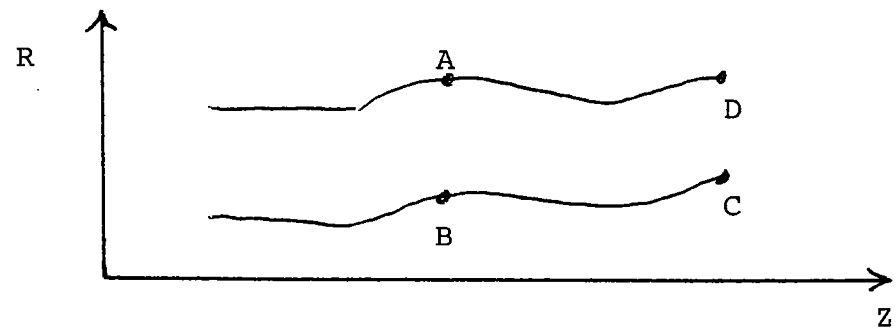


This method is designed to be broadband, hence, Ci would not <u>have</u> to be moved ever, theoretically. I haven't done the analysis yet to check feasibility, but I will.

2. SF08 Calculations

SF08 is a program currently being written for superfish. It will calculate C, L, R_S, P_D, Q, and various voltages and currents. The backbone of this program will be SF07 which is a program that calculates the vectors E_r, E_z, H_ϕ along a specified path. Superfish is the program which solves the TM_Z modes of any circularly symetric resonator. SF07 uses data generated from Superfish, and SF08 uses SF07 as a subroutine.

Given the following cross-section of a resonator, circularly symetric about Z:



at points A and B or between them for the TM case:

$$I_{A\,rMS} = 2 \, H_\phi A \, \Pi r_A \qquad I_{B\,rMS} = 2 \, H_\phi B \, \Pi r_B$$

and $$V = \int_A^B E_r \, dr$$

We will use a simple integrating routine and probably break this into 100 steps from A to B. This should be plenty accurate enough.

Superfish normalizes everything to a arbitrary voltage. With respect to our needs, a constant K is defined to normalize a voltage at say D→C. All other values are then scaled accordingly.

So each $V^1 = kV$, Each $I^1 = kI$

Superfish also calculates stored energy and power dissipated. Again we must normalize this.

$$Ue = \frac{\varepsilon r}{2} \int_V E^2 \, dV \qquad \text{if} \quad E^1 = kE$$

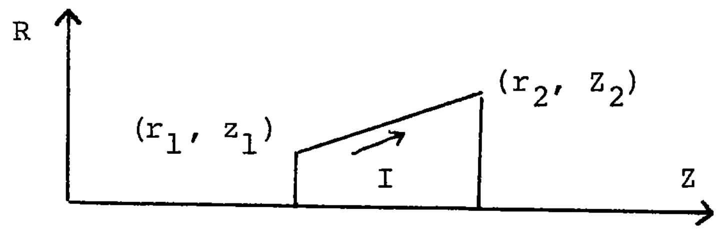
Then $$U^1e = \frac{\varepsilon}{2} \int_V KE^2 \, dV = k^2 \frac{\varepsilon}{2} \int_V E^2 \, dV$$

$$Ue^1 = k^2 Ue \qquad \text{At resonance } Ue = Um$$

$$Um = \frac{\mu}{2} \int_V H^2 \, dV$$

Because Superfish has problems around regions with Insulators for power calculations, and I don't particularly like the output format, SF08 will have its own power routine.

To calculate the power lost on an arbitrary conductor in small steps assuming we know the current there, we need a non-standard geometry that will default to any case. Any case can be thought of as an anulas, cylinder, or combination of the two. I therefore select a cone as my geometry which can default to either, or any combination.



$$Pd = Rs \int_s \left|H_{tan}\right|^2 dS \qquad ds = 2\pi \left|r\right| \sqrt{1+ \left(\frac{dr}{dz}\right)^2}\ dz$$

$$Rs = \sqrt{\frac{W\mu}{2\sigma}} \qquad \left|H_{tan}\right| = \frac{I}{2\pi r}$$

$$r = \frac{r_2 - r_1}{Z_2 - Z_1}(Z - Z_1) + r_1 \qquad dr = \frac{r_2 - r_1}{Z_2 - Z_1}\, dz$$

$$\sqrt{1+ \left(\frac{dr}{dz}\right)^2} = \frac{\sqrt{(Z_2 - Z_1)^2 + (r_2 - r_1)^2}}{(Z_2 - Z_1)}$$

$$2\pi \left|r\right| \cdot \left(\frac{I}{2\pi r}\right)^2 = \frac{I^2}{2\pi r}$$

$$\Rightarrow \quad \left|H_{tan}\right|^2 \cdot 2\pi \left|r\right| \cdot \sqrt{1+ \left(\frac{dr}{dz}\right)^2}\ dz$$

$$= \frac{I^2}{2\pi} \frac{\sqrt{(Z_2-Z_1)^2 + (r_2-r_1)^2}\ dz}{\left[Z(r_2-r_1) - Z_1(r_2-r_1) + r_1(Z_2-Z_1)\right]}$$

$$Pd = \frac{R_s I^2}{2\pi} \frac{\sqrt{(Z_2-Z_1)^2 + (r_2-r_1)^2}}{(r_2-r_1)}$$

$$\cdot \int_{Z_1}^{Z_2} \frac{dz}{Z(r_2-r_1) - 2(r_2-r_1) + r_1(Z_2-Z1)}$$

Let $u = Z(r_2-r_1)-Z_1\ (r_2-r_1)+r_1(Z_2-Z_1)$

$$du = (r_2-r_1)dz$$

$$u1 = r_1(Z2-Z1)$$

$$u2 = r_2(Z2-Z1)$$

$$Pd = \frac{R_s I^2}{2\Pi} \sqrt{\frac{(Z_2-Z_1)^2+(r_2-r_1)^2}{(r_2-r_1)}} \int_{u1}^{u2} \frac{du}{u}$$

$$Pd = \frac{R_s I^2}{2\Pi} \sqrt{\frac{(Z_2-Z_1)^2 + (r_2-r_1)^2}{(r_2-r_1)}} \ln \left(\frac{r_2}{r_1}\right)$$

$$Pd = \frac{RsI^2}{2\Pi r_1} \sqrt{(Z_2-Z_1)^2 + (r_2-r_1)^2} \frac{\ln (^{r2}/_{r1})}{(^{r2}/_{r1}) - 1}$$

Check:

$r2 = r1 \Rightarrow$ cylinder

$$Pd = \frac{R_s I^2}{2\Pi r_1} \cdot (Z_2-Z_1) \cdot \frac{0}{0}$$

$$\lim_{r2/r1 \to 1} \frac{\ln (^{r_2}/_{r_1})}{(^{r2}/_{r1})-1} = \left.\frac{^{r1}/_{r2}}{1}\right|_{r2/r1 = 1} = 1$$

$$Pd = \frac{R_s I^2}{2\Pi r_1} (Z_2-Z_1) \quad \text{correct}$$

$Z2 = Z1 \Rightarrow$ anulus

$$Pd = \frac{P_s I^2}{2\Pi r_1} \cdot (r_2-r_1) \cdot \frac{\ln (^{r2}/_{r1})}{(^{r2}/_{r1})^{-1}} = \frac{R_s I^2}{2\Pi} \cdot \ln (^{r2}/_{r1}) \Rightarrow \text{correct}$$

A simple subroutine can now calculate powers lost stepping along an arbitrary symetic geometry provided steps are kept reasonable. We just need to keep r's and z's in correct perspective.

We can now use earlier formulations and all the information desired can be calculated and displayed in a reasonable format.

JV:as