Calculations and Simulations with Pads and Feedback Amplifiers

One of my favorite circuits uses a bipolar transistor with two forms of negative feedback. One is parallel feedback in the form of a resistance from the output collector back to the input base. The other feedback is emitter degeneration. We devoted an entire section (2.7) to this circuit in the discussion of amplifiers in Chapter 2 of **Experimental Methods in RF Design**.

We often follow these circuits with an attenuator or pad. Although this topology is one that has been around for a long time, the role of the pad is not universally understood. We thought it would be useful to examine the circuit with some readily available software tools. We will begin with a look at a pad and what it does to improve termination impedance. Then we investigate a feedback amplifier. The first amplifier will have a pure resistive termination, but this is soon replaced with a narrow bandwidth load. We then make things even worse with a terminated bandpass filter as a load. Finally, we insert a pad at the amplifier output and examine the results.

The software we use is all available at no cost on the internet. One of the programs is SPICE while the other is a Smith Chart program. If you are one of the many radio amateurs (or for that matter, engineers) who do not "embrace" the Smith Chart, just read past those sections.

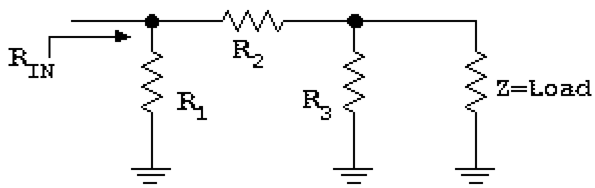
An editorial aside: I am generally not one to advocate excess use of circuit simulators. All too often such programs generate an attitude that "you don't have to build it if you can simulate it." This is just not true. It has always been my experience that a circuit that has been simulated to death, often for weeks on end, is always much better understood after that same circuit is exercised for just an afternoon on the bench. There is just nothing like experiments with measurements. On the other hand, simulators can be used to investigate circuits, refine component values, and actually expand our intuition about those circuits. But this happens best when great care is used in the simulation, sometimes including refinements such as parasitic reactances. I think that the best thing to recommend for the student or the radio amateur experimenter is to integrate his or her simulations with a liberal dose of experiments. These experiments should include, in the case at hand here, measurement of input return loss to amplifiers.

Pads

The pad, or resistive attenuator is a usually symmetrical circuit designed around a characteristic resistance. The resistive pad is intended to be inserted in a Z0 line where a typical value for Z0 would be 50 Ohms. If such a pad is terminated in 50 Ohms at the output, a well defined attenuation is established and the input impedance is also 50 Ohms. These conditions are sufficient to derive design equations for the circuit. The most common topology is a pi-network of three resistors, although a Tee is also common. Other circuits are sometimes used. An L-network will generate a well defined input resistance and attenuation, but does not present an identical output resistance.

A major feature of a pi pad becomes clear if we consider a termination, Z, and calculate the input resistance, R-in. The circuit and these calculations are summarized in the following:

Pad Analysis (w7zoi, 3Aug06)



Conductance of Z and R3 = $\frac{1}{R_3} + \frac{1}{Z} = \frac{(Z + R_3)}{R_3 \cdot Z}$

Equiv. resistance = $\frac{1}{(Z + R_3)} \cdot R_3 \cdot Z$

Now add in R2: $\frac{1}{(Z + R_3)} \cdot R_3 \cdot Z + R_2$

Conductance of this = $\frac{1}{\left[\frac{1}{(Z + R_3)} \cdot R_3 \cdot Z + R_2\right]} = \frac{(Z + R_3)}{(R_3 \cdot Z + R_2 \cdot Z + R_2 \cdot R_3)}$

Now add in conductance of R1:

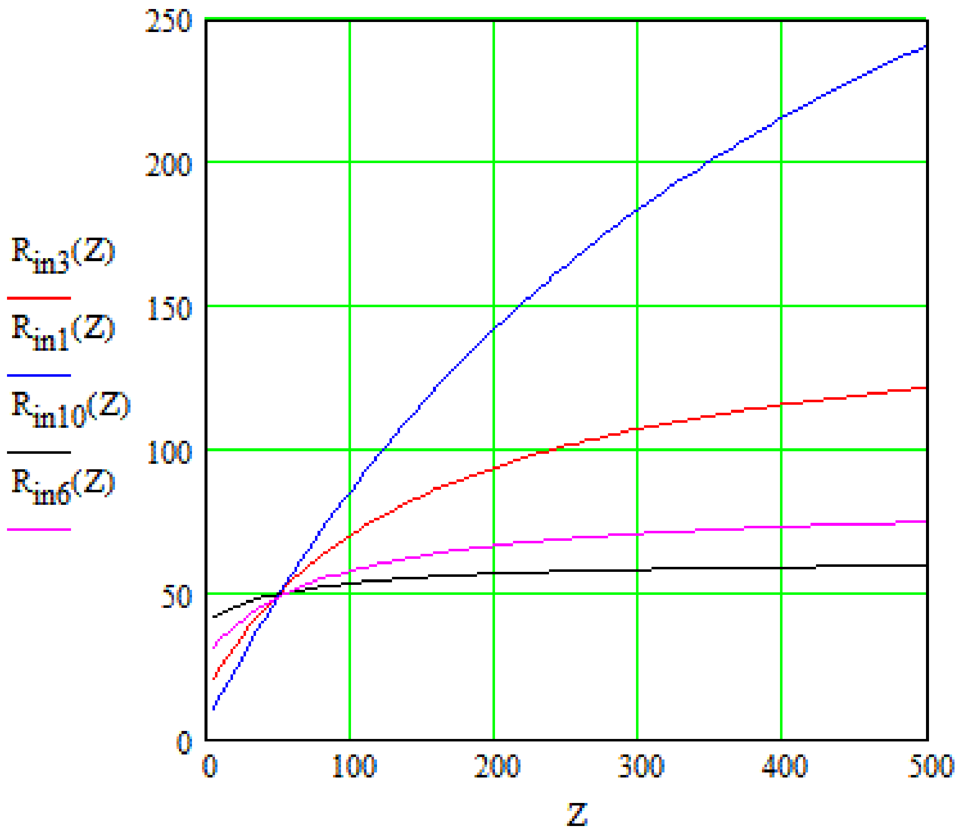
$$\frac{(Z + R_3)}{(R_3 \cdot Z + R_2 \cdot Z + R_2 \cdot R_3)} + \frac{1}{R_1}$$

And finally convert to a resistance

$$R_{in} = \frac{1}{\frac{(Z + R_3)}{(R_3 \cdot Z + R_2 \cdot Z + R_2 \cdot R_3)} + \frac{1}{R_1}} \qquad R_{in} = \frac{R_1 \cdot R_2 \cdot Z + R_1 \cdot R_3 \cdot Z + R_1 \cdot R_2 \cdot R_3}{(R_1 \cdot Z + R_1 \cdot R_3 + R_3 \cdot Z + R_2 \cdot Z + R_2 \cdot R_3)}$$

The result is converted to a function of the load, Z. We then consider four common cases for 50 Ohm Pads: 1 dB (R1=910, R2=5.6), 3 dB (R1=300, R2=18), 6 dB (R1=150, R2=36), and 10 dB (R1=100, R2=68). In all cases R3=R1. These resistor values are standard 5% values that are used when building practical circuits. The design equations are found in EMRFD, Ch. 7.

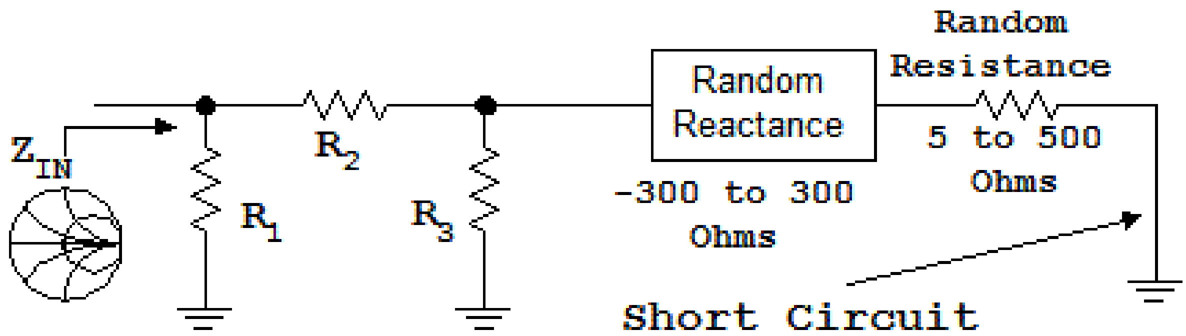
We allowed the output load resistance, Z, to vary from 5 to 500 Ohms and examined the input resistance of the four pads. The results are plotted here:



The terminated 1 dB pad has an input resistance that varies over a range nearly as large as the termination. In contrast, the 10 dB pad has an input R that is close to 50 Ohms, no matter what output termination is used on the output. Owing to symmetry, the usually pi pad will have an output resistance similarly related to the input termination, or source impedance.

A similar behavior occurs when complex terminations with series reactance are used. The analysis is not, however, as straight forward. Rather than go through these esoteric and probably less than illuminating calculations, we will take a more direct approach based on the Smith Chart. A Smith Chart program that we wrote several years ago for DOS, **MicroSmith,** is used. (MicroSmith is still available, now free of charge. It is part of the collection of DOS software distributed with the ARRL edition of **Introduction to RF Design**. The software is found on the ARRL web site by entering IRFD in the search box.) The analysis uses a little known, although not completely hidden feature that allows random component values to be programmed into the program. (The feature is accessed by pressing Alt-Z while viewing the Smith Chart.)

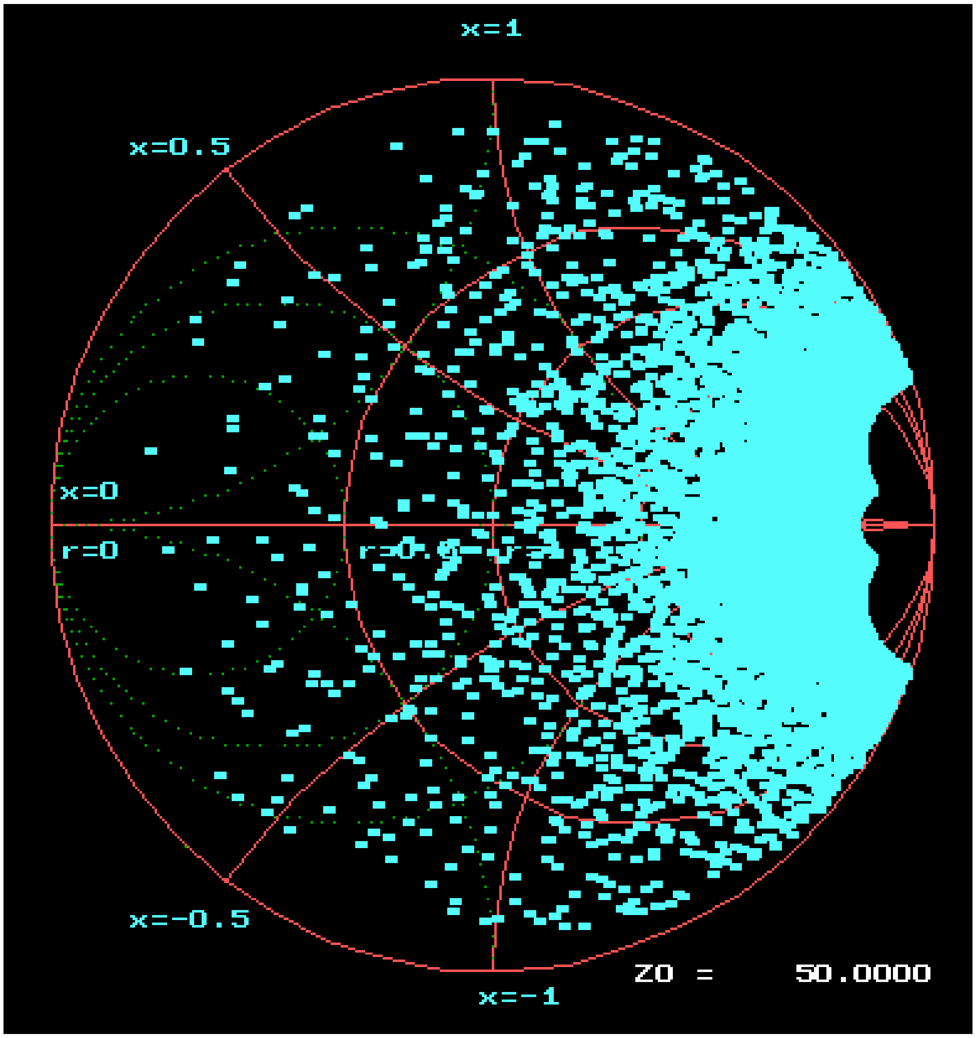
This circuit starts with a short circuit termination of 0+j0. This is then followed by a resistor that randomly varies from 5 to 500 Ohms. The next element is a random reactance that randomly assumes values from -300 to 300 Ohms. The next three elements are then the resistors that compromise the pi-pad. (The random variations are set up with Ctrl-F10 after activated with Alt-Z.) This is illustrated with a figure:



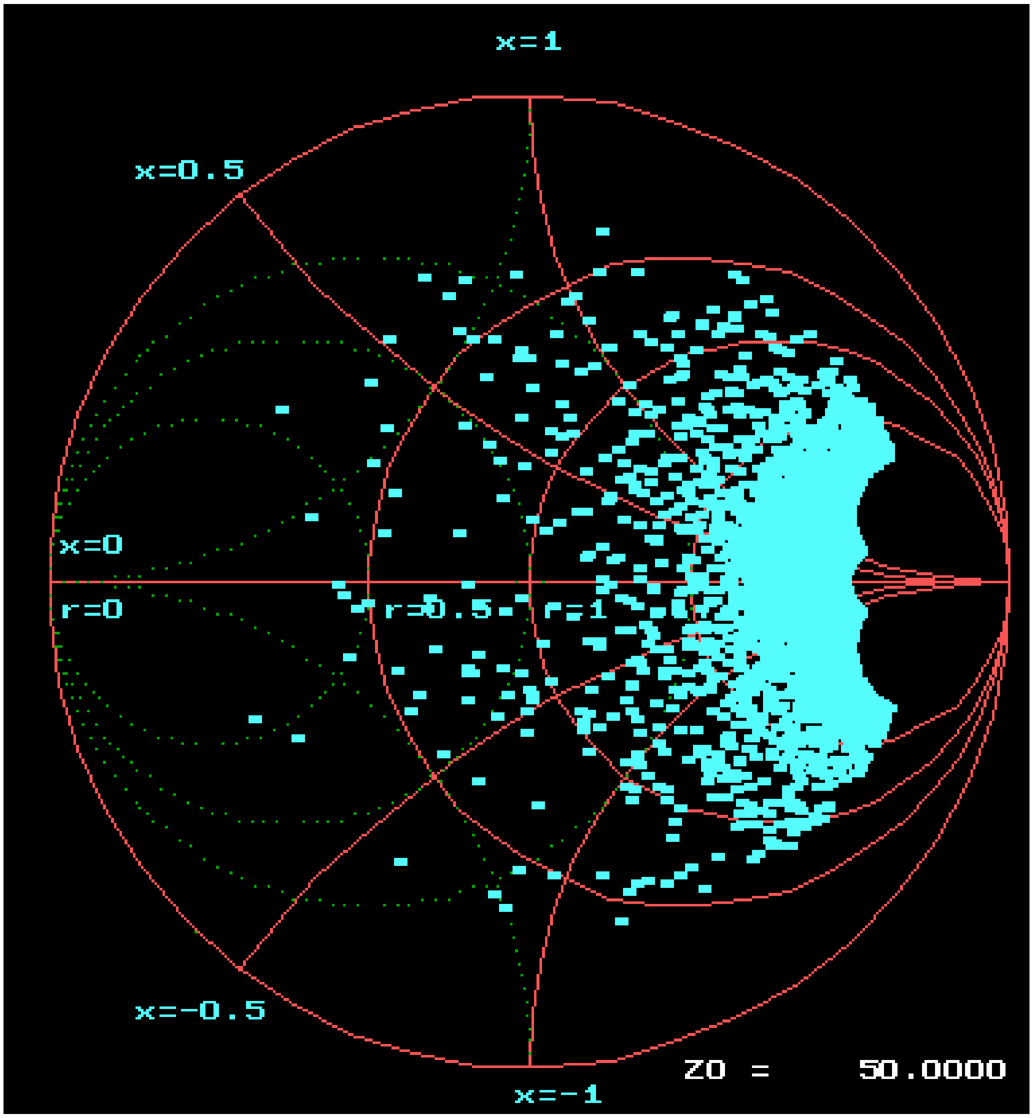
For example, assume MicroSmith generates a random resistance of 80 Ohms and a reactance of -100 Ohms. The resulting impedance is 80-j100. This Z is placed in parallel with R3 to generate a new complex impedance. R2 is added to this and the result is paralleled with R1 to produce a final input complex impedance, Z-in. The final result is plotted as a point (a small rectangle) on the Smith Chart. This process is repeated a large number of times to generate a family of points filling region, suggesting the possible impedances that are possible.

Remember that the Smith Chart is merely a way of plotting impedances. All possible impedances with <u>positive</u> resistive parts, but all possible reactive parts (positive and negative) appear within the circle. The impedance at the center of the circle is the characteristic resistance of the chart, often 50 Ohms.

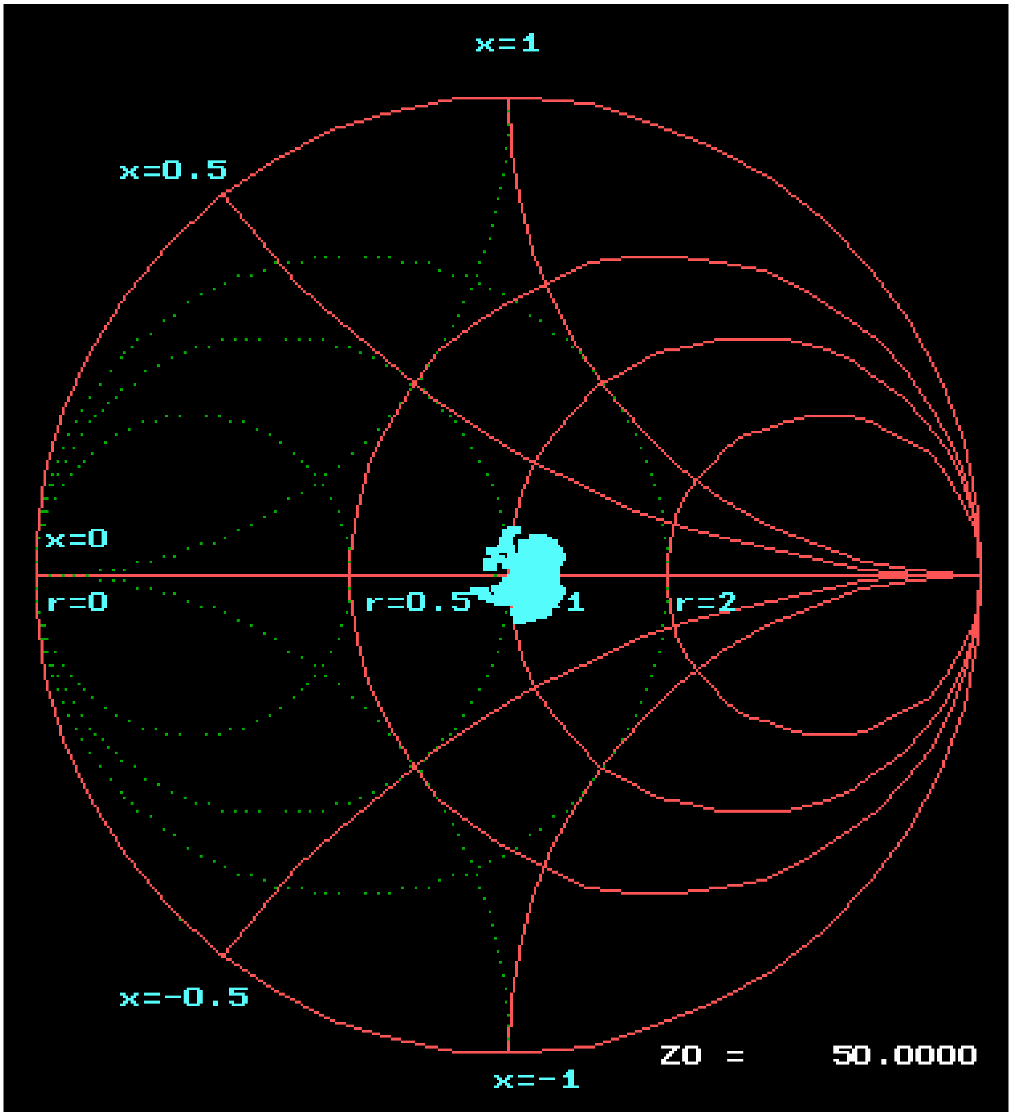
The results follow:



This plot shows the impedances created at the circuit input with the output terminations described. **There is no pad present for this simulation.** Each calculation produces a small rectangle that is plotted within the Smith Chart circle. Regions near the left edge are precluded because the resistance does not drop below 5 Ohms. The blank region at the right extreme results because there are no resistances above 500 Ohms. "Dots" above the chart center line correspond to a positive, or inductive input impedance while the negative reactance capacitive impedances are below the center line. Several thousand calculations were used for this simulation.



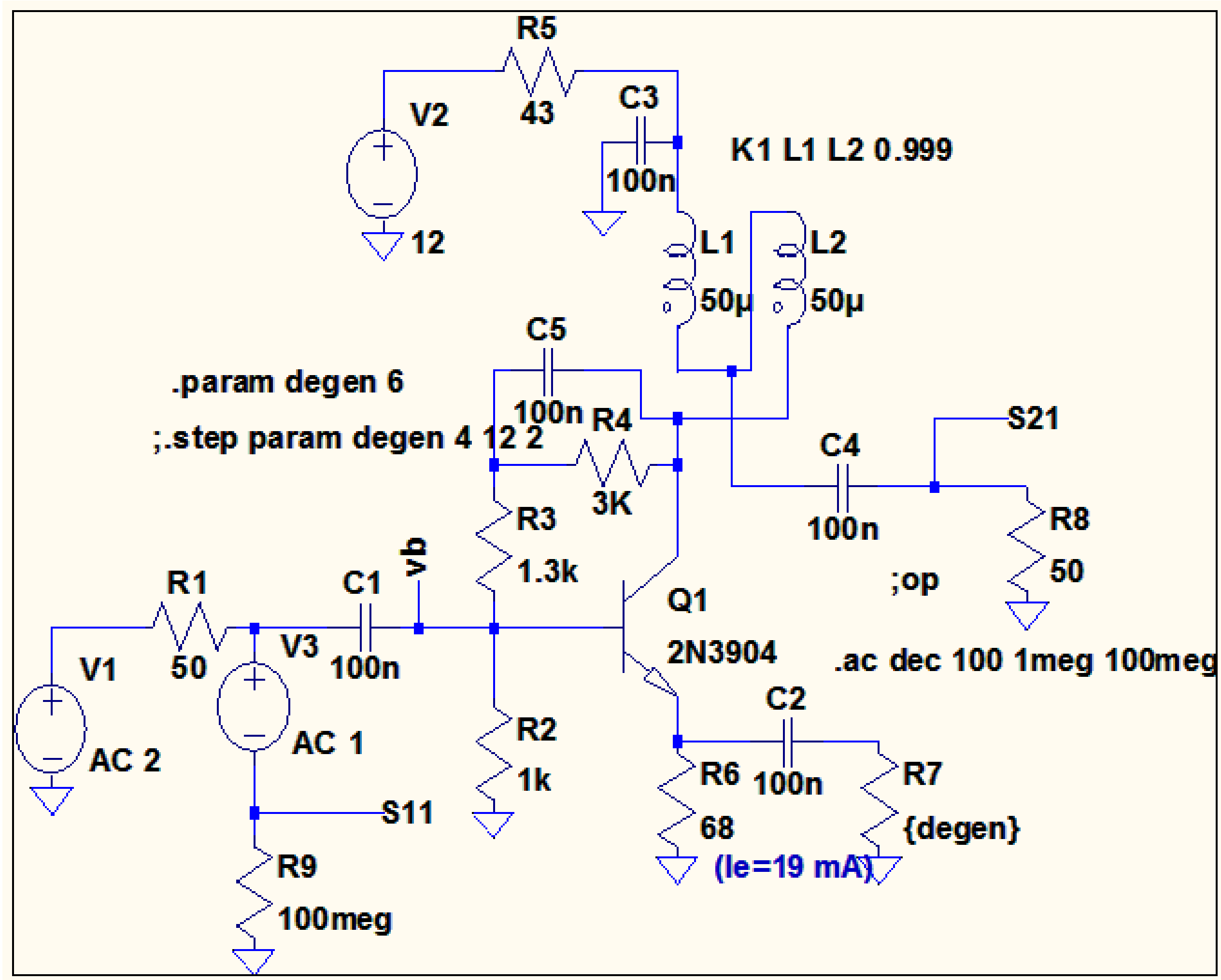
A1 dB pad is now used in the circuit. The effect is that no impedances close to the edge of the chart are allowed. This simulation used 2000 calculations.



The third example, a simulation with only 1000 calculations, uses a 10 dB pad. Impedances are now confined to a small region around the 50 Ohm center of the chart.

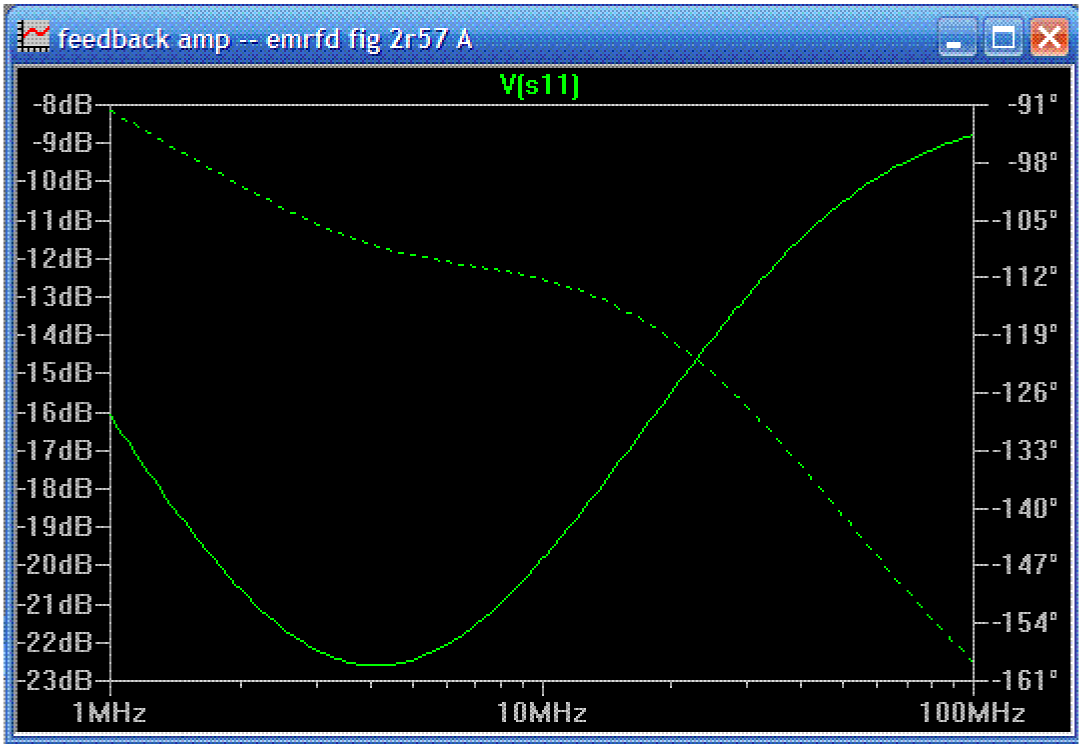
A Feedback Amplifier

We now consider a feedback amplifier simulation using SW Cad, a free version of SPICE distributed by Linear Technology, a Bay Area Semiconductor house. We are grateful to LT for this wonderful tool that they maintain and provide, free of charge. The schematic for the amplifier considered here is from Fig 2.57 of EMRFD with the simulation circuit shown:



The amplifier is driven by a 50 Ohm generator with an open circuit voltage of 2. Note that if this 50 Ohm generator is loaded with 50 Ohms, the voltage that results at the load is 1. A floating (non grounded) 1 volt voltage source is attached to the generator output. If the generator was terminated in an exact 50 Ohms, the floating generator voltage would subtract from the signal at the output of the 50 Ohm source to produce zero volts at the point marked "S11." The voltage at this point is the voltage reflection coefficient. (R9 is a dummy resistor that keeps SPICE happy -- the simulator program does not "like" an open circuited voltage generator.) This circuit for the examination of the amplifier input scattering parameter, S11, which is the input voltage reflection coefficient, was discussed in a short 1993 QEX paper, Reflections on the Reflection Coefficient, and is paper #33 on the EMRFD CD. The degeneration resistance, R7, is set at 6 Ohms for all of the simulations presented.

The AC voltages are set to sweep from 1 to 100 MHz. A result of this analysis is shown below:



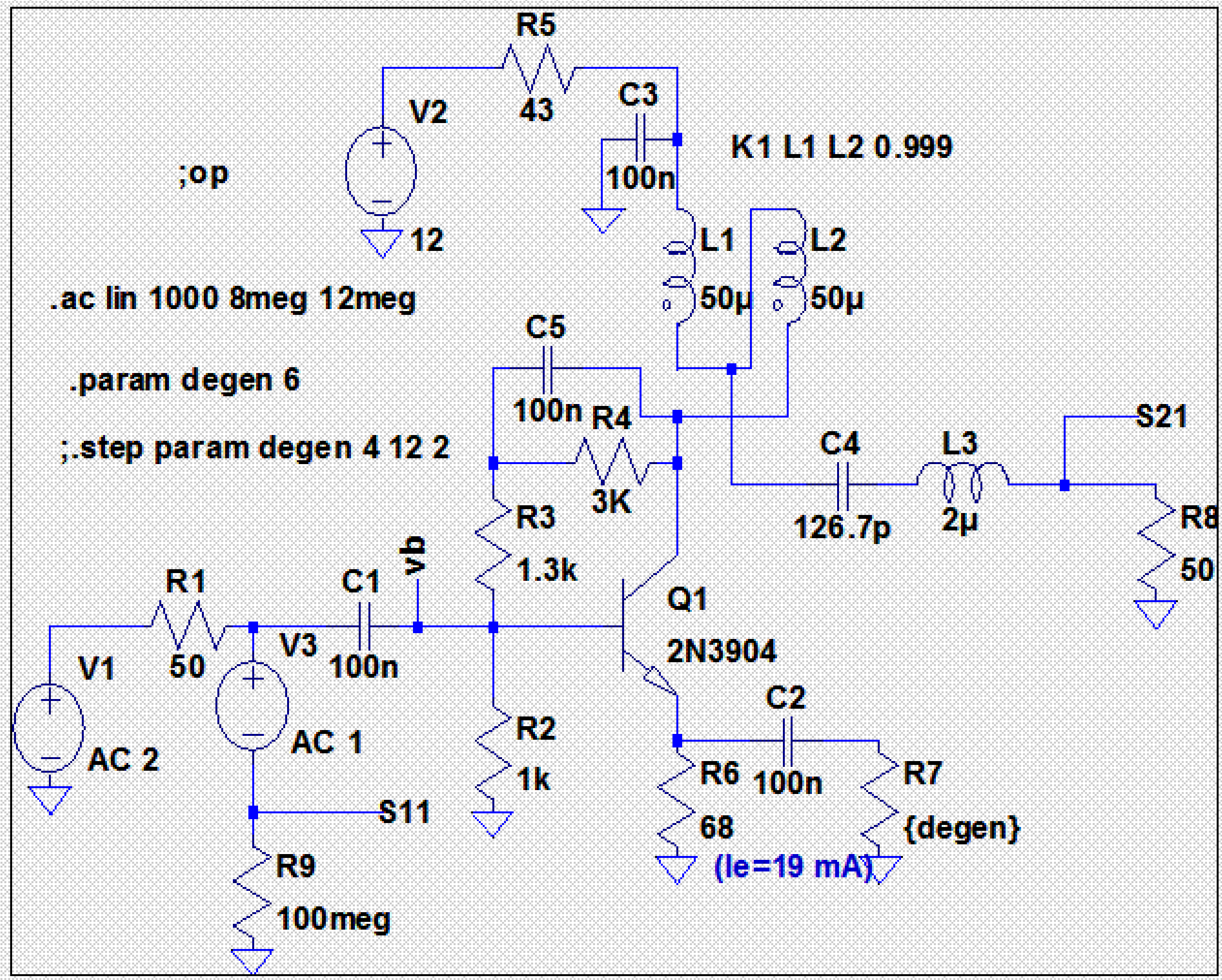
This plot shows the input reflection coefficient plotted over the 1 to 100 MHz region. It is in dB form, which is known as the input scattering parameter, S11, in dB form. The negative of this is ***Return Loss.*** **Return loss, reflection coefficient, and VSWR are all related with some equations presented elsewhere on this web site. Click the blue text.**

The input match, S11, is numerically evaluated by double clicking on the V(S11) from the SPICE program. (Not here -- sorry.) The result at 10 MHz is |S11|= -19.9 dB. (Return loss is, hence, 19.9 dB.) The angle shown at 10 MHz is -112 degrees. The dB return loss value is converted to a voltage reflection coefficient of 0.0102 at -112. (The conversion is done with the methods found in the **blue** link just above.) The reflection coefficient is converted to an impedance with MicroSmith, which yields Z-in at 10 MHz of 49.6 - j 0.939. This corresponds to a 49.6 Ohm resistor paralleled by a 6 pF capacitor.

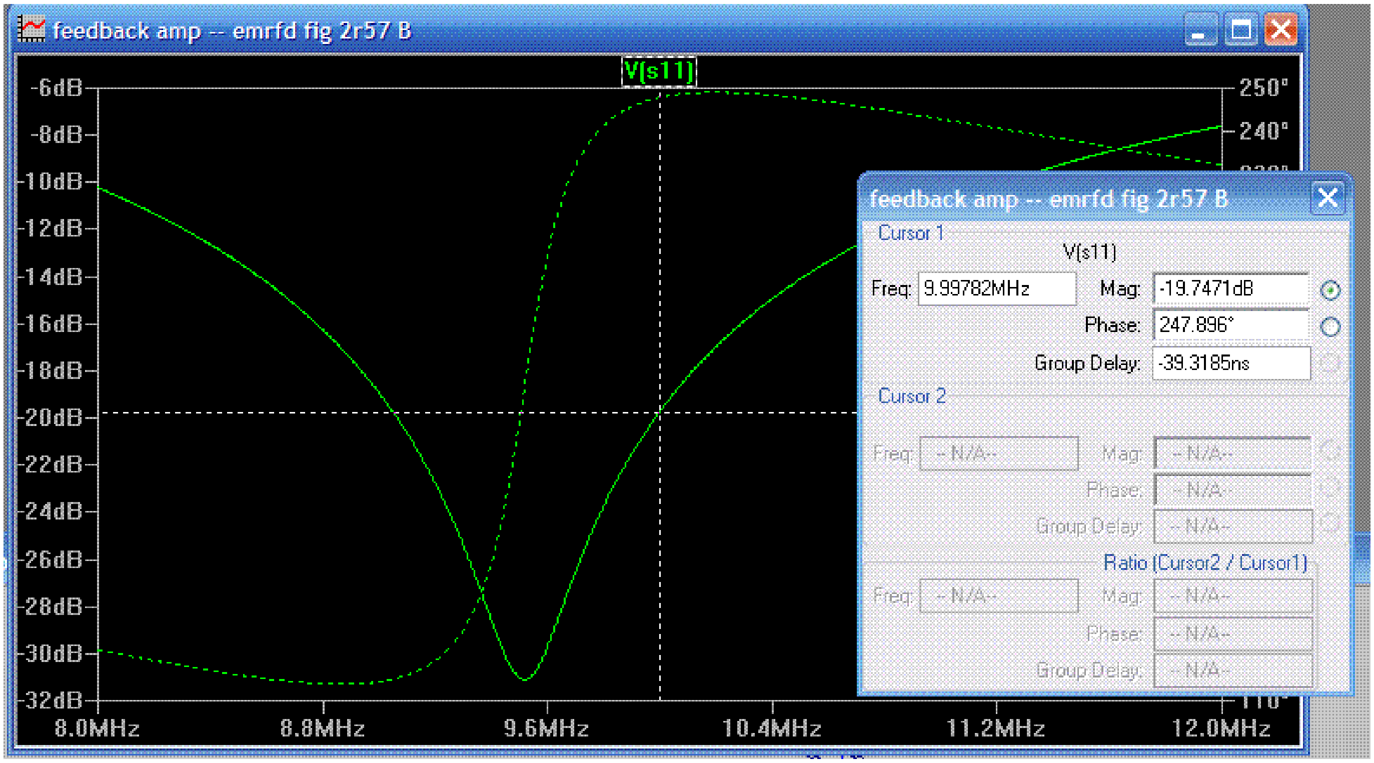
The analysis also shows a forward gain of S21=21.3 dB. The input match, S11, is a function of the feedback elements (R3 and R7) as well as the coupling coefficient K for the windings in the collector transformer.

An Amplifier with a Narrow Band Load.

The analysis above uses a 50 Ohm resistor as the load. It is coupled to the amplifier through a 2:1 turns ratio transformer, usually realized in practice with a bifilar winding on a ferrite toroid. (Typical transformer is 12 bifilar turns #28 on a FT-37-43 toroid core.) The input match is reasonable over the whole HF spectrum. We often want to use an amplifier like this in a receiver to drive a narrow band load, such as a filter. So we ask what happens if a narrow band load is used? This was evaluated by inserting an inductor in series with the load, shown below.



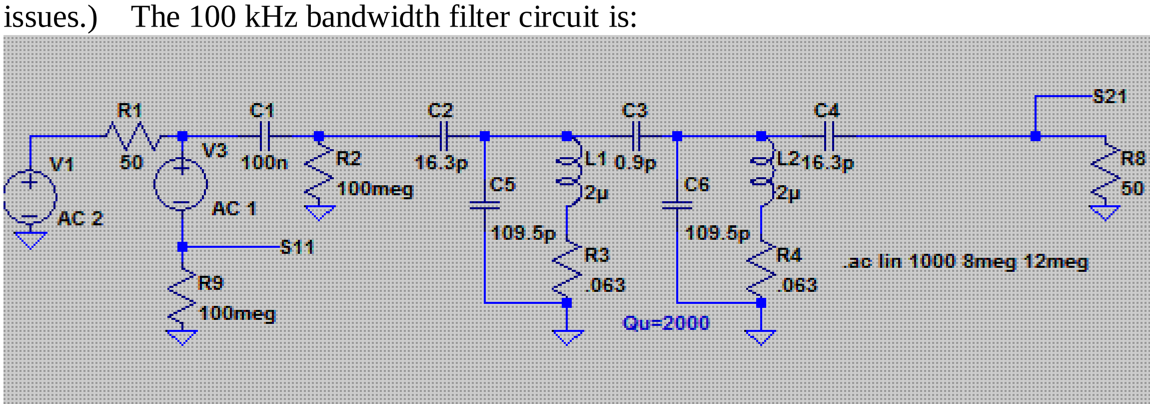
C4 was changed from a 0.1 uF to 126.7 pf, which resonates with L3 at 10 MHz. This results in a drastically different input impedance, shown below, where the sweep range is now reduced to span from 8 to 12 MHz.



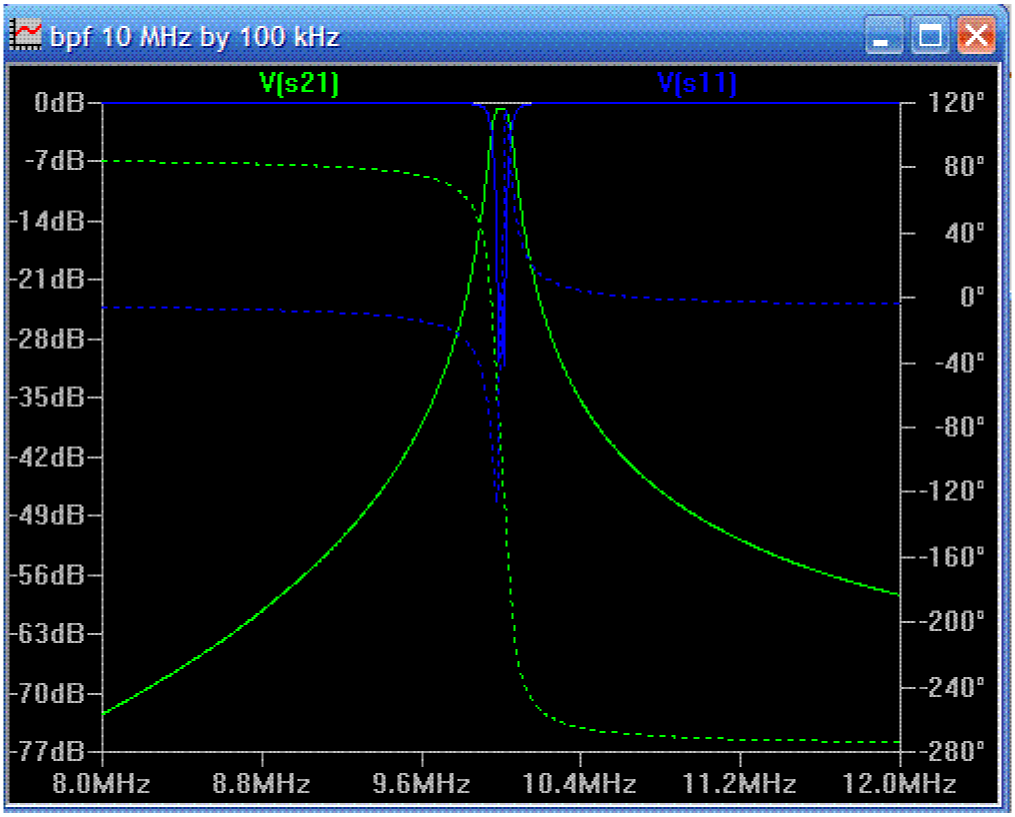
The input match exactly at 10 MHz is changed little from before, a result we would expect. Even the angle is the same. However, it is much different at other frequencies.

Adding an Output Filter

Wanting to show the behavior with termination in a narrow filter, we designed a simple double tuned circuit. We picked unusual inductors with an unloaded Q of 2000. While not off the shelf components, the software defined filter is one with low loss, which means that it should present a good load to the amplifier within the passband. (We could have just as well used a crystal filter, but did not want to get into those additional issues.) The 100 kHz bandwidth filter circuit is:

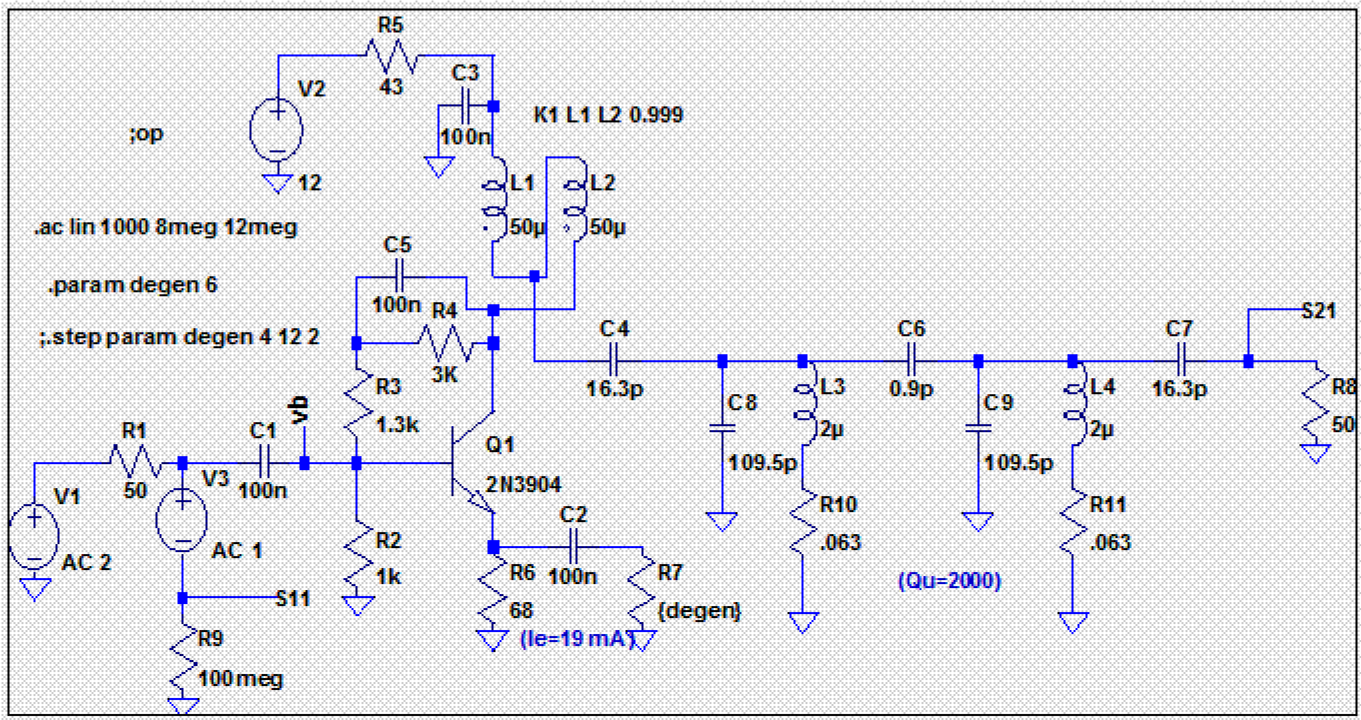


The 0.063 Ohm resistors in series with the 2 uH inductors simulate the unloaded inductor Q. R2 is another no-impact at RF component that keeps SPICE happy. (Don't try to build this filter, for it would be very difficult to find or build inductors with such a high Qu.) The passband response and input match for the filter along are shown below :

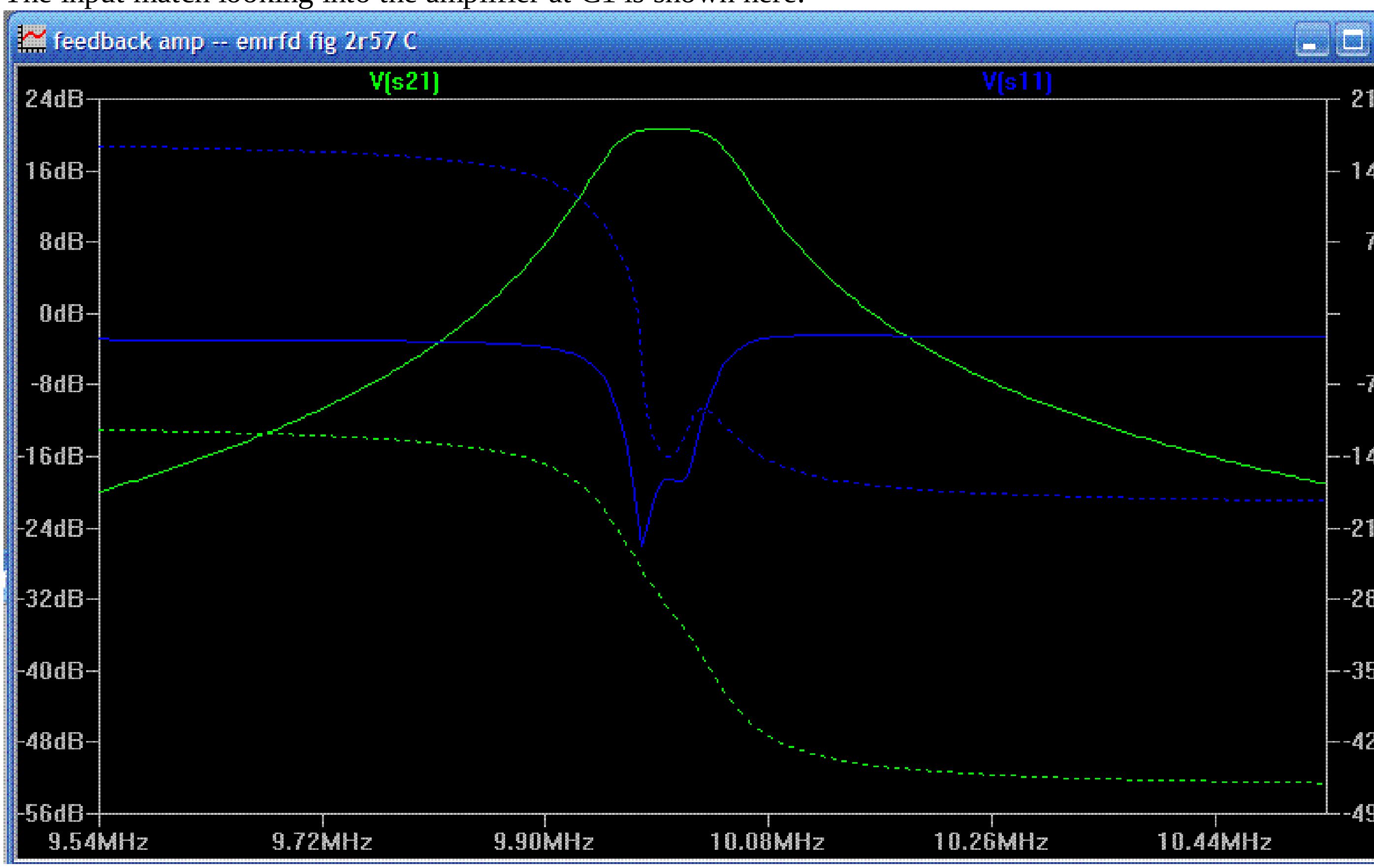


The filter insertion loss is only 0.6 dB. While there is a good input match in the center of the filter, it quickly degrades as you depart from 10 MHz. S11 approaches 1 in the stopband, indicating total reflection. This behavior is typical of virtually all simple LC filters, as well as those with crystals or other similar resonators.

We now add the filter to the output of our amplifier with the total circuit shown below:



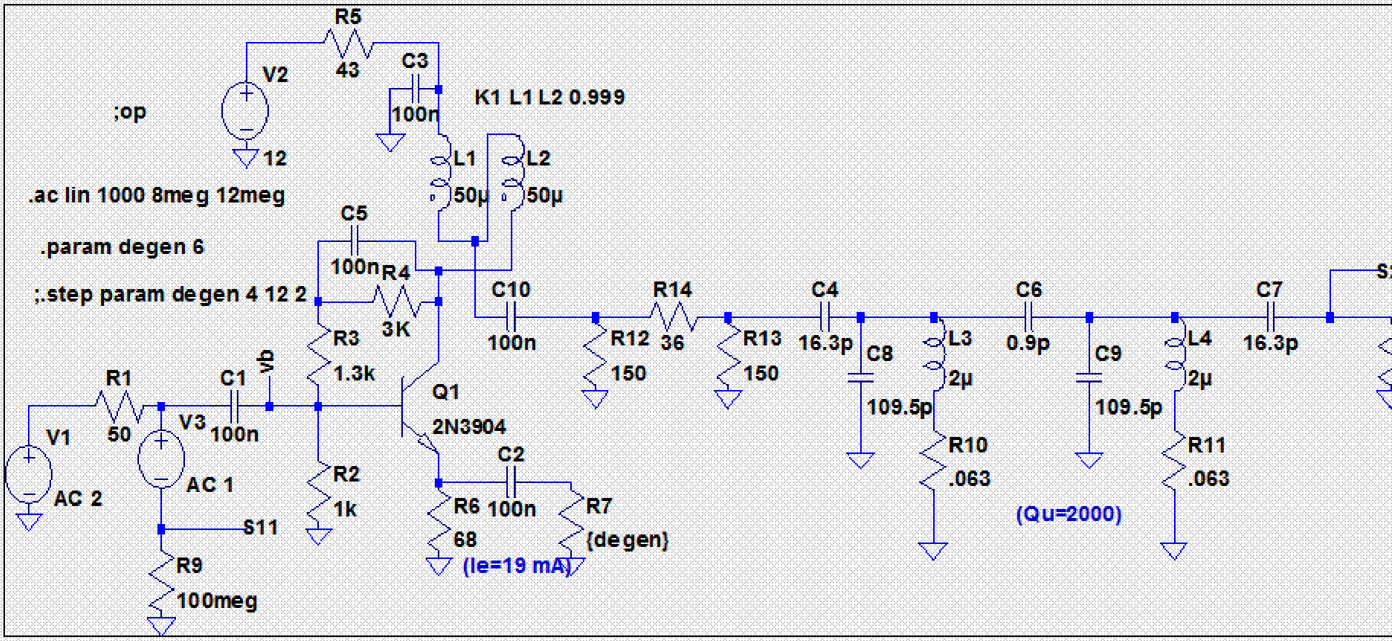
The input match looking into the amplifier at C1 is shown here:



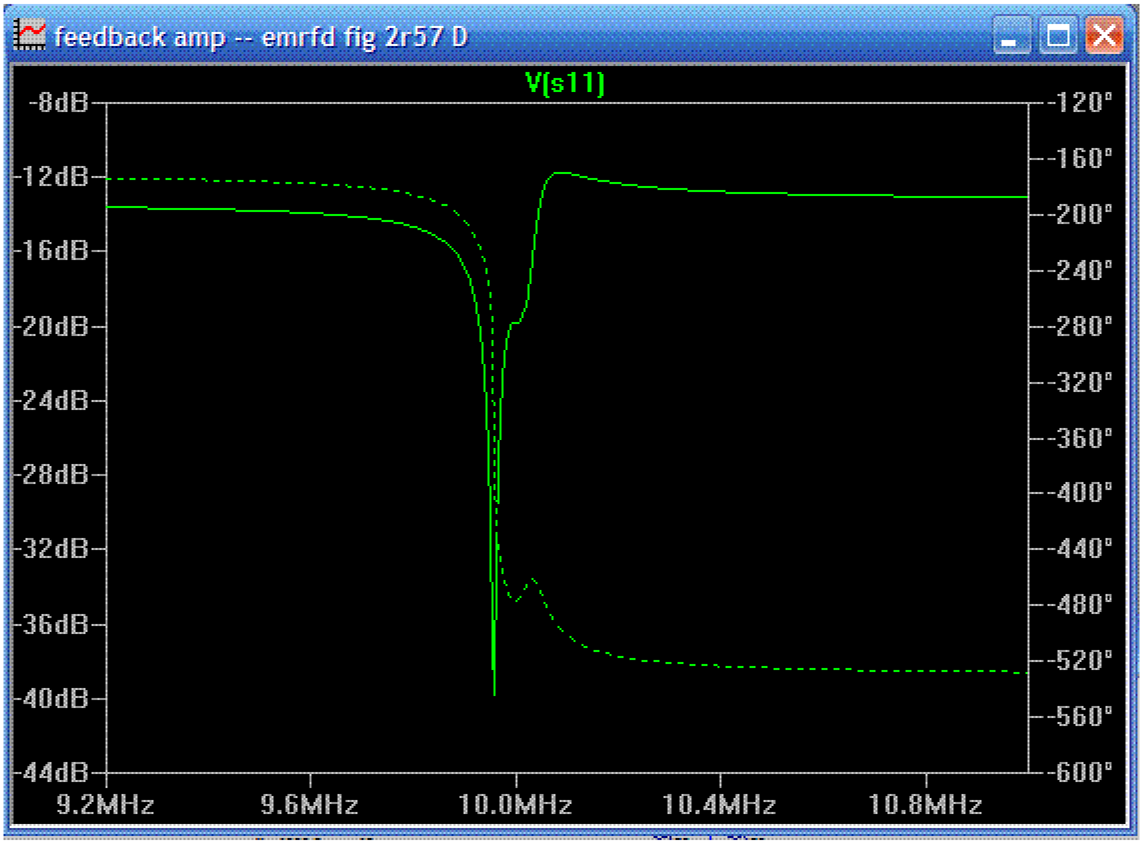
The match is still good at 10 MHz with a return loss of 18.5 dB with an angle of -140 degrees. However, the match degrades to a return loss of only 3 dB in the upper stopband region. This could be a major problem if we tried to use this amplifier to terminate a switching mode mixer, especially one with a diode ring.

Adding a Pad

We now adopt the often seen topology of an amplifier followed by a pad. The circuit with a 6 dB pad (R12, R13, and R14) is shown below:



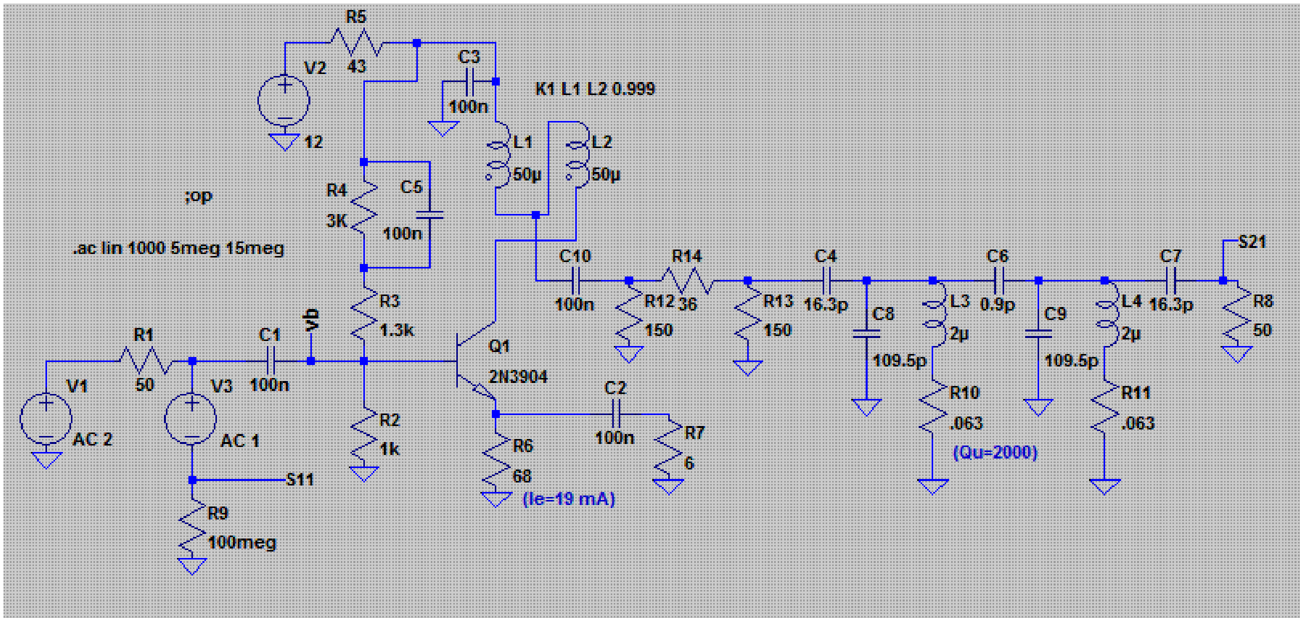
Gain (S21) at 10 MHz, which is within the filter passband, is down to 14.7 dB. A sweep of input S11 is shown below:



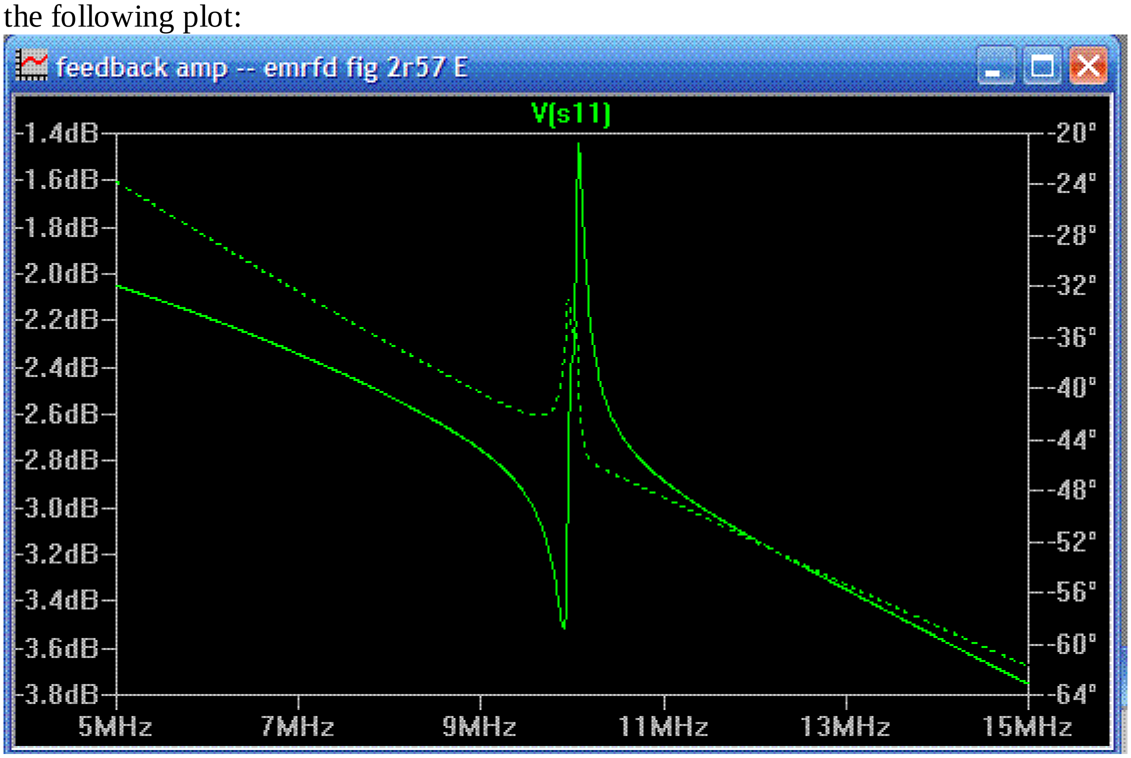
The input match has the narrow band characteristics of the output filter. But they are modified. The in-band match at 10 MHz is a return loss of 19.8 dB. The out of band match is 11.8 dB return loss or better for all frequencies. This is less than a 2:1 VSWR and is good enough to terminate most of the out of band components from a diode ring mixer that cause severe IMD. There is more to be gained with even better terminations, but this represents a first step.

Removing Amplifier Feedback

Finally, we ask what would happen if we modified the amplifier to remove the parallel feedback. This is shown in the following :



In this circuit, we have moved the connection of the bias and feedback network (R3, R4, C5) from the collector to the bypassed and decoupled power supply point. This leaves the DC biasing unchanged, retaining an emitter current of 19 mA. The emitter degeneration is left unchanged. The result is an amplifier with a 10 MHz gain of almost 21 dB, even with the 6 dB pad in place. The input match is shown from 5 to 15 MHz in the following plot:



We see a smooth downward trend in S11, modified by the output tuned filter. At 10 MHz the input impedance is 84.5 - j85.3, which has the equivalent of a 171 Ohms in parallel with 94 pF. (S11= -2.38 dB at -35.6 degrees.)

Further Thoughts and Conclusions

1. Pads or attenuators can be extremely useful in RF systems for establishing a stable, wideband impedance. The worst case match looking into a pad (with the output termination being a short or open circuit) is a return loss equal to twice the attenuation. Hence, a 6 dB pad will establish a return loss looking into the pad of 12 dB or better. This is intuitively reasonable. Envision a pulse of energy launched toward the 6 dB pad. The signal will undergo a 6 dB attenuation. It then encounters a discontinuity at the open or short circuit. All energy is then reflected with the reflected wave going back toward the pad. It then returns to the source where it is again attenuated by 6 dB. The wave returns 12 dB weaker than it was when launched. (Another term we might used is "out and back" loss.)

2. Amplifiers are usually desired in our communications systems. An bipolar transistor amplifier should ideally have modest to high standing current, for the high power capability means that it can amplify small signals without excess distortion. But high current also means high gain. This gain can be reduced with negative feedback while maintaining the high current needed for low distortion.

3. Emitter degeneration is a form of negative feedback that increases input impedance. This is why followers have increased input resistance.

4. Parallel feedback in the form of resistance between collector and base reduces both input and output impedance.

5. Combinations of the two feedback forms can be combined to establish a desired gain and low S11 and S22. A good starting point for feedback elements is (R-fb)(R-degen) = (R-source)(R-load) In our example amplifier, the source is 50 Ohms while the load (at the collector) is 200 Ohms. Hence, with a feedback resistor of 1300 Ohms, we would start with a degeneration R of 50*200/1300=7.7 Ohms. We used a value of 6 in parallel with the bias R of 68 Ohms in series with the r-e of the biased transistor (26/19=1.37) for a net of 6.9 Ohms, which is close.

6. Decreasing parallel feedback R while increasing degeneration move toward lower gain.

7. The feedback amplifier shown is a low power design. This is limited to much less than a Watt output. (Recall the equation for output load resistance for a given power: R-load = (Vcc-Ve)^2/(2*Pout) .) If higher output power is desired, the transformer should be eliminated.

8. Any form of narrow band output termination will be reflected back to an amplifier input. This occurs to a greater extent with negative feedback amplifiers. (Problem: Simulate a lower gain amplifier (more feedback) than the one we have used and examine the effect of narrow band terminations.) Transformer feedback amplifiers (EMRFD

Fig 2.81) are useful for low noise and low distortion with excellent impedance matches. However, they can be especially bad in reflecting an output load back to the input.

9. The problems of amplifiers reflecting impedances from one port to the other can often be fixed to some extent with a pad. In the example we used, we showed that a pad at the output improved the amplifier input match. It also improves the output match seen looking back from a point "to the right" (in our examples) of the pad. When we used a 6 dB pad, the worst case mismatch seen when looking back is 12 dB return loss. Hence, a filter just after the pad is going to see a reasonable termination.

10. Just one pad in a multiple stage amplifier can serve to preserve match in more than one stage.

11. It is often convenient to merge biasing and feedback resistors in one design, realizing fewer components in an amplifier. It is just as possible to build circuits where the bias elements and feedback components are well isolated. See EMRFD, Fig 2.66.

12. Although not a "universal truth," negative feedback tends to enhance amplifier stability with regard to parasitic oscillation.