The Incidental Double Tuned Circuit – A Transceiver T/R Design Subtlety.

11 Jan 2004, update August 2007.

A common topology for low power communications transceivers uses a Transmit-Receive (T/R) switch with a single series tuned circuit. Imbedded diode or transistor switches become a “short circuit” during transmit, preventing energy from reaching the receiver. There is, however, an interaction that often occurs when this single tuned circuit is attached to the one normally found in the receiver front-end. The result is an unwanted overall response that could lead to IMD and other overload problems. We illustrate this problem (and a solution) with a design example, a simple transceiver using a NE-602 Gilbert Cell mixer front end.

We begin with the receiver front end, before adding the T/R system. This is outlined in the following figure.

Front-End Design Example

Assume we want a single tuned circuit at 7 MHz to preceed a NE602 or similar receiver front end IC. This part has R in of 3000 Ohms. Assume we want a bandwidth in this tuned circuit of 0.2 MHz. The filter Q is then:

$$Q_F = \frac{7}{0.2} \quad \text{or} \quad Q_F = 35$$

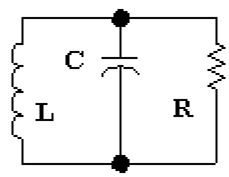
But, our inductor has an unloaded Q of 250.
So the external filter Q will be: $Q_e = 42$

This results from the relationship where two loads establish a net Q.

$$\frac{1}{Q_f} = \frac{1}{Q_u} + \frac{1}{Q_e}$$

We wish to load the tuned circuit equally by the connection on each side, so the external Q will be 84 on each side. The 3000 Ohm input resistance of the IC should create this Q of 84 for the inductor in the tuned circuit. But Q relates to parallel load resistance through

$$R_p = Q \cdot \omega \cdot L$$



Solving for L, we obtain at 7 MHz:

$L = 812$ nH.

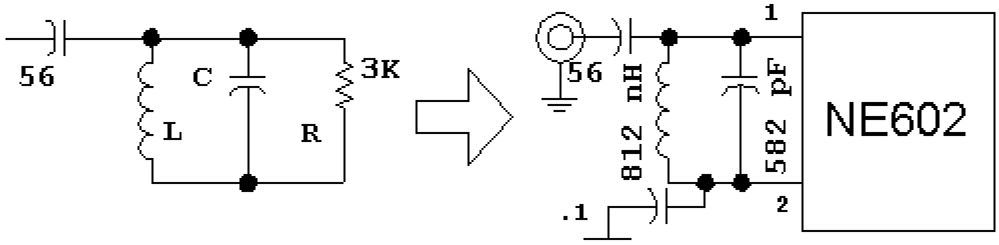
The C required for resonance is calculated from the usual $\omega^2 = \frac{1}{L \cdot C}$

$C = 637$ pF.

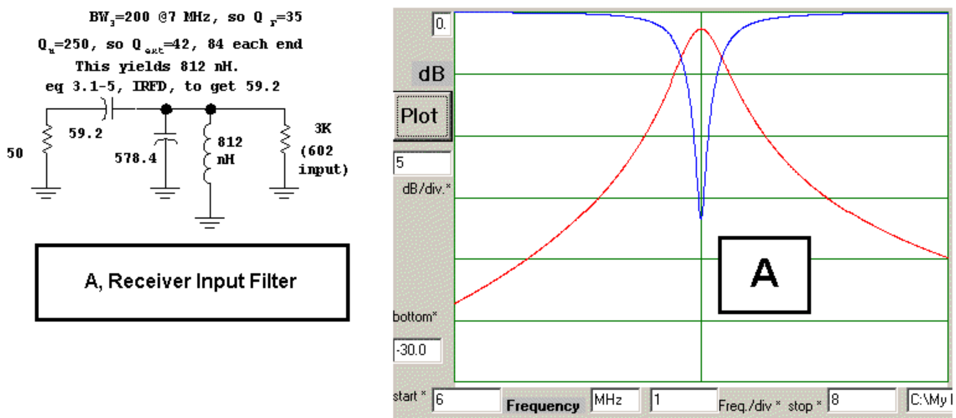
We wish to connect the tuned circuit to a 50 Ohm antenna at the other end through a series capacitor that will cause the 50 Ohms to look like 3000 Ohms at the tuned circuit. The viable relationship is given in Eq. 3.1-5 from Intro. to RF Design. Rp is 3000 and R0 = 50.

$$X_C = \sqrt{R_p \cdot R_0 - R_0^2}$$

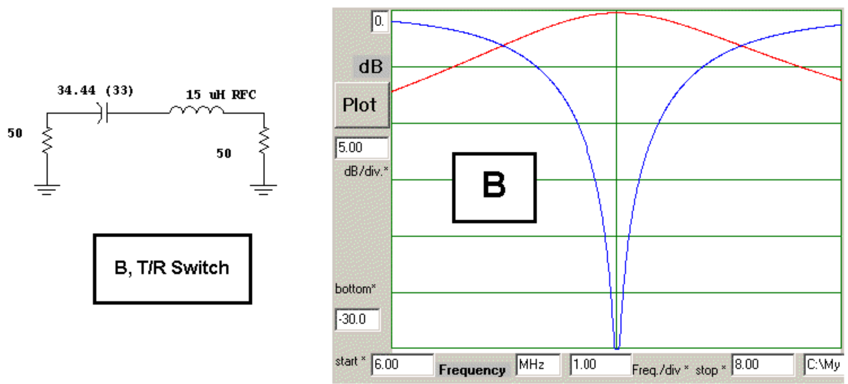
Juggling the numbers produces a series C value of 59 pF. We will use a standard value of 56 pF.



The response (S21) of this single tuned circuit is shown in Fig. A below in red. The blue trace represents the input impedance match, indicated by the magnitude of S11 looking into the receiver. This represents a good input match looking into the receiver.

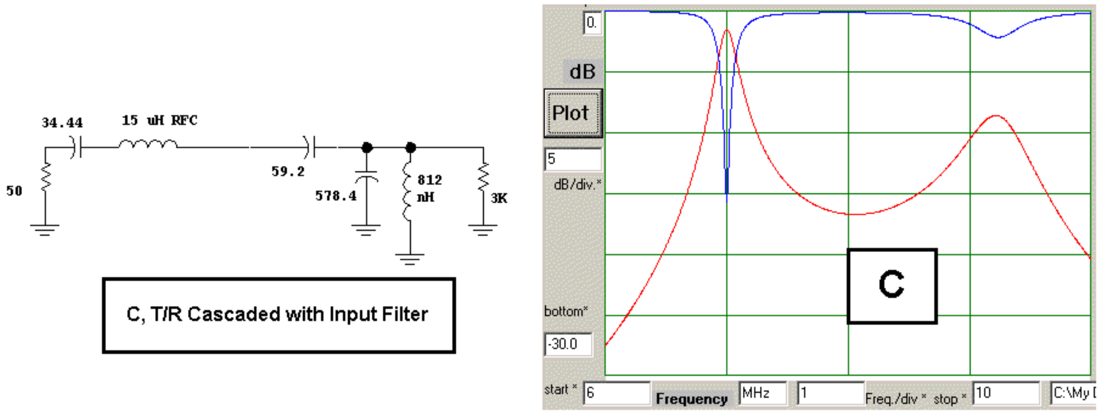


We now design the T/R switch tuned circuit. A 50 Ohm termination is assumed at each end, resulting in Fig. B.

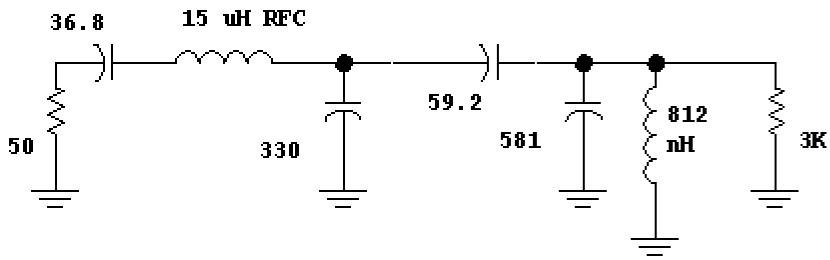


Again, both gain and impedance match are excellent. The 15 uH inductor is assumed here to have the same unloaded Q as was used with the original receiver resonator. The usual design includes a pair of back to back diodes from the junction of the L and C connected to ground. The diodes are transparent, almost open circuits during receive, but turn on during transmit intervals, creating the needed short circuit to ground that protects the receiver. See details in the paper "**Electronic Antenna Switching**," QEX, 1995, including data on insertion loss and IMD introduced into the receive path.

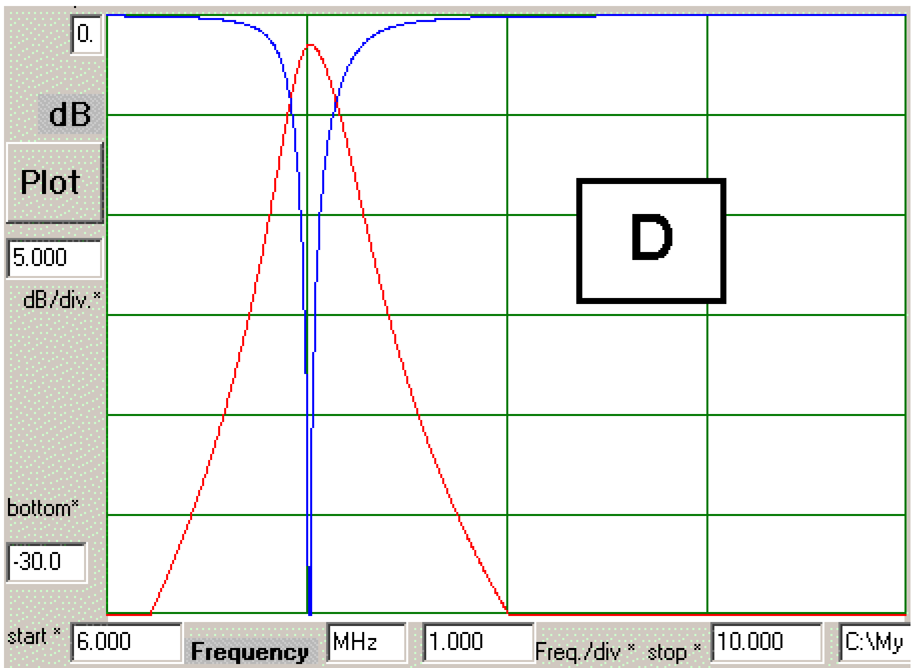
We now connect the two circuits together, only to discover an undesired interaction, shown below in Fig. C. The response is severely compromised with significant energy appearing at the output at both the 7 MHz design frequency and around 9.2 MHz. This is a part of the spectrum containing numerous loud signals that should be suppressed at the receiver mixer. The interaction is a normal one that we should have expected. A double tuned circuit has been formed, creating a double humped response when over coupling is present. (See **"The Double-Tuned Circuit: An Experimenter's Tutorial,"** QST, December, 1991.)



A simple solution is shown in Fig. D. We have added a shunt 330 pF coupling capacitor between the two tuned circuits. This provides an impedance where energy can be shared between resonators, but at an intensity that actually improves the front end selectivity rather than compromising it. The double humped response has been replaced by a well defined, single peak.



D, Fixed Cascade with added coupling capacitor.



We have seen a similar undesired response in other receivers that we have built. In all cases, a careful analysis with a wideband response sweep allows the problem to be diagnosed and fixed with the addition of a single component. But the simplified approach to T/R switch design that we used in earlier designs has been modified.

A recently described transceiver ("**The Micromountaineer Revisited**," QST, July, 2000, pp28-33) suffers from the difficulty. The problem can be solved at 7 MHz by adding a capacitor across J5. Any value from 330 to 470 pF should be fine. C12 and C20 will then need retweaking.

The experimenter should not treat this difficulty as a reason to dismiss the basic T/R method, for it is still sound. Many others have used the topology with great success. (see Burdick, "**NorCal 40, A Club Project**," QRPp, Journal of the Northern California QRP

Club, Vol. 1., No. 3, December 1993; Lewallen, "**An Optimized QRP Transceiver**," QST, Aug, 1980.

About any linear circuit analysis software can be used to analyze and adjust the networks prior to construction. The plots shown were generated with General Purpose Ladder Analysis, part of "LADPAC-2002," a Windows (registered to Microsoft, Inc.) based version of the software now distributed with **Experimental Methods in RF Design.** (as of Feb, 2003).

Regarding the NE602 input impedance: (10Aug07)

Often we see the NE602 input modeled as ideal transistors (which means infinite beta) with 1.5K resistors from each input to "ground." From what I've been able to tell, the circuit is really just a differential pair with a total bias current of about 2 mA (1 mA per transistor.) The equivalent emitter résistance is then re=26/Ie(mA), or 26 Ohms. If the beta is 100, this would put the resistance looking to the base at 2.6 K. If we consider frequencies near or just above the beta cutoff, the net impedance drops down to 1.5K, justifying the traditional model.

But there is one more detail: The base-to-base input impedance of a bipolar differential pair is twice that of the single transistor. This is the result of the common emitter connection that is between the two devices. This point would be a virtual ground if the bases are driven differentially. It is a floating point with half the input voltage if one of the bases is bypass to ground, which is the usual situation with a Gilbert Cell. Hence, the input impedance is around 3K paralleled with a couple of pF of capacitance. This is just what I measured several years ago when I had daily access to a wideband network analyzer.