A Look at T-Tuners

Bernard G. Huth, W4BGH

The subject of "Antenna Tuners" may be misnamed because they don't actually "tune" the antenna. An antenna tuner is actually an impedance matching device that "matches" the impedance of the source (usually a transceiver with an impedance of $Z_G = 50\ \Omega$) to the antenna. Maximum power is transferred when the modified impedance of the antenna is equal to the complex conjugate of the source impedance (e.g. $Z = Z_G^*$).

In order to start our investigation of impedance-matching circuits it will be useful to consider a graphical interpretation of adding series and parallel reactances to a "load" impedance $Z_L = R + jX$. If we add an inductor or capacitor in series with this impedance, the total impedance becomes $Z' = R + j(X \pm X_S)$ where the plus sign is for an inductive reactance and the minus for a capacitive reactance. This is illustrated in Figure 1 where we can see that adding a series inductance or capacitive reactance moves the impedance up or down vertically in the impedance plane.

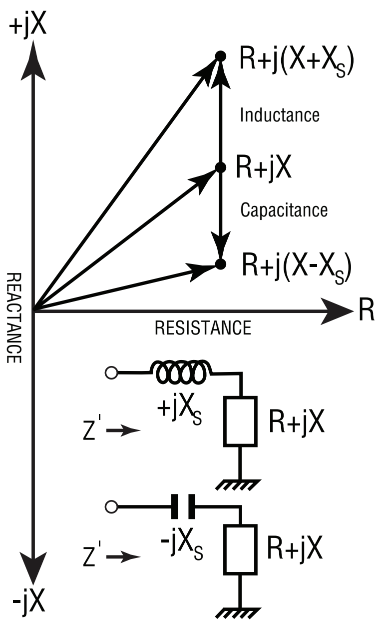


Figure 1

If we consider adding a parallel inductance or capacitance to the "load" impedance, it is useful to deal with the *reciprical* of the impedance, i.e. *1/Z,* known as the ***admittance***.

$$Y_L = \frac{1}{R+jX} = \frac{R-jX}{(R+jX)(R-jX)} = \frac{R}{R^2+X^2} - j\frac{X}{R^2+X^2} \qquad (1)$$

which can be written:

$$\begin{aligned} Y &= G + jB \\ G &= \frac{R}{R^2+X^2} \\ B &= -\frac{X}{R^2+X^2} \end{aligned} \qquad (2)$$

G is called the ***conductance*** and B is called the ***susceptance***. The unit of admittance is the Siemans (symbol **S**) although it is synonymous with the unit mho and the symbol ℧.

When impedances are added in parallel, we may sum their admittances. So if we add a reactance X_P in parallel with an impedance $Z_L = R + jX$, this is the same as a susceptance $B_P = -1/X_P$ in parallel with an admittance $Y = G + jB$ which becomes $Y' = G + j(B + B_P)$. Since the conductance $G = R/(R^2+X^2)$ is constant regardless of the value of X_P, it is true that the conductance is unchanged by the addition of a parallel reactance:

$$G = G' = \frac{R'}{(R'^2 + X'^2)} \qquad (3)$$

which can be rewritten:

$$R'^2 + X'^2 - R'/G = 0 \qquad (4)$$

which is the equation of a circle with the properties that the center of the circle lies at $R_o = 1/(2G), X_o = 0$ and its radius is $1/(2G)$ (see Figure 2)[1]. The circle just touches the origin of the

[1] "Impedance Matching, Part 1: Basic Principles," David Knight G3YNH and Nigel Williams G3GFC, http://www.g3ynh.info/zdocs/z_matcing/part_1.html

graph at $0+j0$ and is called the ***circle of constant conductance***. Using the graphical understanding shown in Figures 1 and 2 we can now follow impedance transformations in the Impedance Plane. This is very similar to the use of a Smith Chart, but uses more familiar Cartestian coordinates.

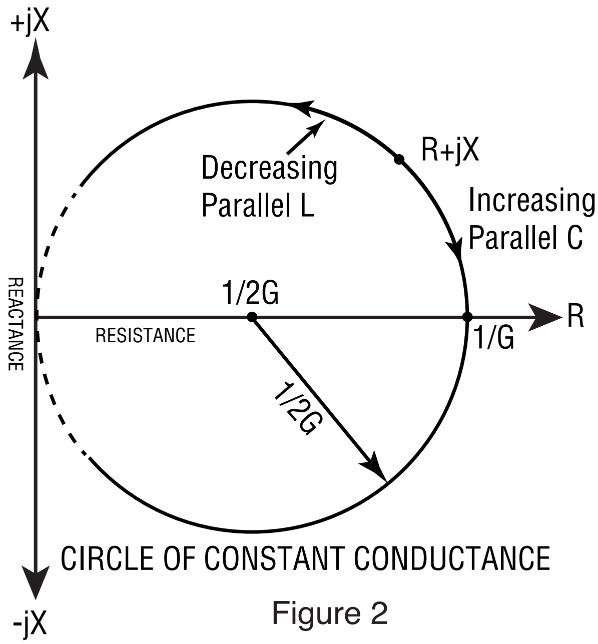


Figure 2

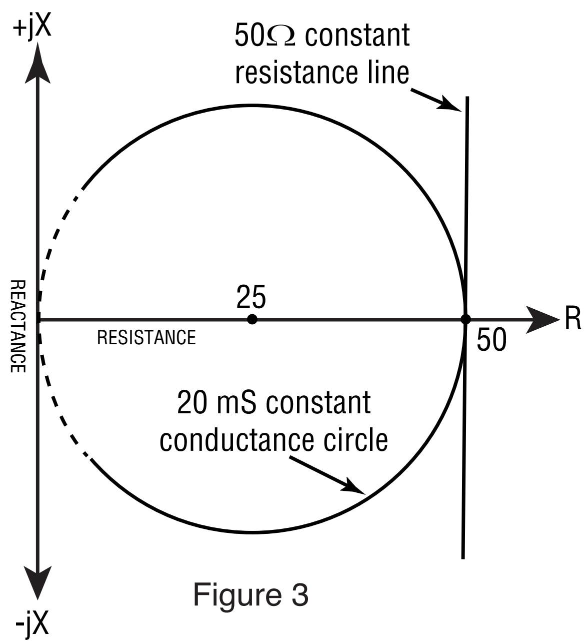


Figure 3

Figure 2 shows that if one adds a parallel reactance, the impedance is transformed around the circle of constant conductance in the clockwise direction when a capacitor is connected across *Z* and in the counter-clockwise direction if an inductor is added across *Z*. Furthermore, if you want to move farther counter-clockwise around the circle reduce the amount of parallel inductance. Similarly, if you wish to move farther clockwise around the circle add a larger amount of parallel capacitance.

So from the ideas in Figures 1 and 2 one can see that the process of matching an impedance presented by an antenna to the transmitter, say 50Ω using inductors and capacitors is to manipulate the "load" impedance, *Z*, to a new point, $Z' = 50 + j0$ by moving along lines of constant resistance or around circles of constant conductance.

Note that the target impedance $50 + j0\Omega$ lies on the 50Ω constant resistance line (see Figure 3). An initial impedance that does not lie on this line can always be brought to it by moving it around a circle of constant conductance, i.e. by placing a reactance in parallel with it. An intermediate impedance that lies on this line can always be brought to $50 + j0$ by placing a reactance in series with it. Therefore, impedance matching can always be carried out in a two step operation in principle.

The constant conductance circle on which $50 + j0\Omega$ lies is known as the 20mS constant conductance circle (i.e. 20 milli-Siemans or $1/\ 50$ Siemans).[1] Its radius is $1/2G = 25\Omega$ and its center lies at $25 + j0\Omega$. It crosses the resistance axis at *0* and at $1/G = 50\Omega$.

If an initial impedance has a resistive component of less than 50Ω, it can always be manipulated onto the constant conductance circle by first placing a reactance in series with it. Then an intermediate impedance which lies on the 20mS circle can then be brought to $50 + j0$ by placing a reactance in parallel with it.

Figure 4 (taken from Reference 1) shows six different regions in the Z-plane, identified according to their relationship to the target impedance, $50 + j0$. It also shows examples of two-component matching networks generally called L-Section networks because of the positioning of the two reactances. The encircled numbers indicate the operations that must be performed, the order in which to perform them, and their effects.

One important thing to note is that when the resistive part of the initial impedance is less than 50Ω,

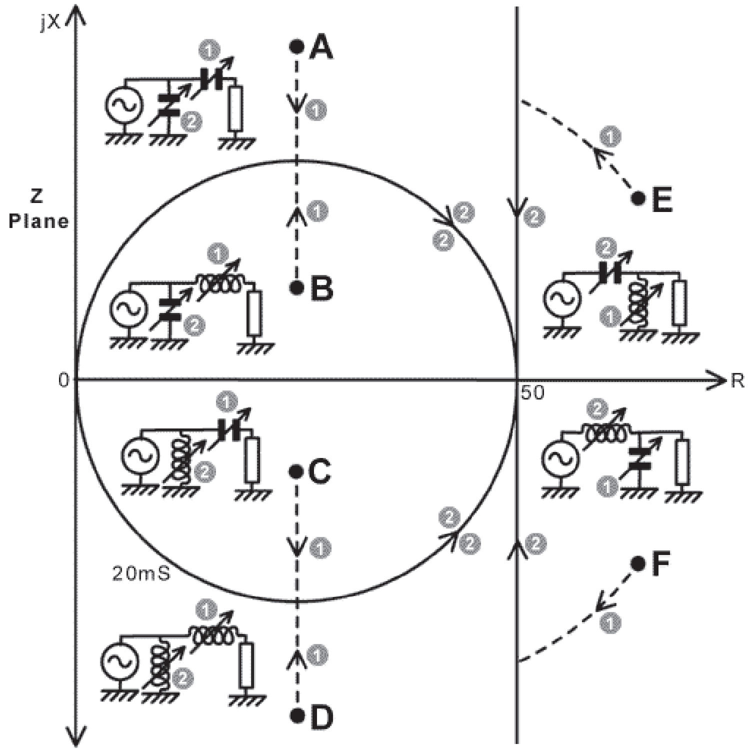


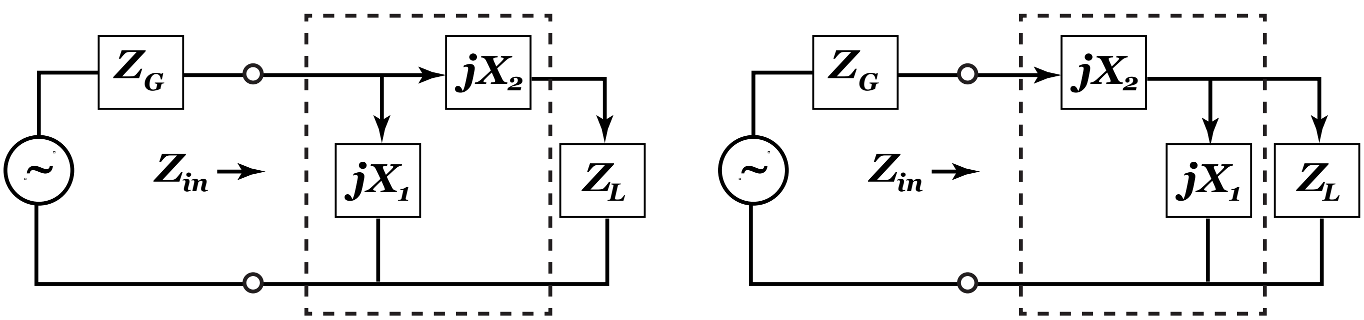
Figure 4
(Taken from Reference 1)

i.e. regions A, B, C, and D in Figure 4, the first matching element is always a *series* reactance. This is commonly called a “normal” L-Section and is required when $R < 50\Omega$.

Similarly, when the resistive part of the initial impedance is greater than 50Ω , i.e. regions E and F in Figure 4, the first matching element is always a *parallel* reactance. This is called a “reversed” L-Section and is required when $R > 50\Omega$.

Now before we consider T-networks, let’s take a closer look at these L-Sections shown in Figure 5 to derive the conditions for the existence of a matching solution of a particular type[2]. The inputs to the design procedure are the complex load and generator impedances, $Z_L = R_L + jX_L$ and $Z_G = R_G + jX_G$. The outputs are the reactances X_1, X_2. For either type, the matching network transforms the load impedance Z_L into the complex conjugate of the generator impedance:

$$Z_{in} = Z_G^* = R_G - jX_G \qquad (5)$$



normal *L-Section* ($R_G > R_L$) **reversed *L-Section* ($R_G < R_L$)**

Figure 5 L-Section Reactive Conjugate Matching Networks

where Z_{in} is the input impedance looking into the L-Section:

$$Z_{in} = \frac{Z_1(Z_2 + Z_L)}{Z_1 + Z_2 + Z_L} \qquad \text{(normal)}$$

$$Z_{in} = Z_2 + \frac{Z_1 Z_L}{Z_1 + Z_L} \qquad \text{(reversed)} \qquad (6)$$

with $Z_1 = jX_1$ and $Z_2 = jX_2$. Using the equations (6) into the condition in equation (5) and equating the real and imaginary parts of the two sides, we find a system of equations for X1, X2 with solutions

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[2]. Electromagnetic Waves and Antennas, Sophocles J. Orfanidis, ECE Department, Rutgers University, 94 Brett Road, Piscataway, NJ 08854-8058, http://www.ece.rutgers.edu/~orfanidi/ewa/ch12.pdf, November 2002, pp 510-519.

for the two types:

$$X_1 = \frac{X_G \pm R_G Q}{\frac{R_G}{R_L} - 1} \qquad\qquad X_1 = \frac{X_G \pm R_L Q}{\frac{R_L}{R_G} - 1}$$

$$X_2 = -(X_L \pm R_L Q) \quad \text{(normal)} \qquad\qquad X_2 = -(X_G \pm R_G Q) \quad \text{(reversed)} \qquad (7)$$

$$Q = \sqrt{\frac{R_G}{R_L} - 1 + \frac{X_G^2}{R_G R_L}} \qquad\qquad Q = \sqrt{\frac{R_L}{R_G} - 1 + \frac{X_L^2}{R_G R_L}}$$

You can see that the reversed solution is obtained from the normal solution by exchanging Z_L with Z_G. Both solutions assume that $R_G \neq R_L$. If $R_G = R_L$, then for either type the solution is:

$$X_1 = \infty, \quad X_2 = -(X_L + X_G) \qquad (8)$$

Also notice that the Q of the L-Section, and hence the bandwidth, is determined by the Generator and Load resistances. We shall see that using a π- or T-Network will give us an extra degree of freedom that allows us to pick a different Q.

We can write the conditions for real-valued solutions of X_1 and X_2, which are that the Q-factors in Eq. (7) are real-valued (or that the quantities under the square roots are non-negative). The four mutually exclusive cases are[2]:

Existence Conditions	L-Section Types
$R_G > R_L, \quad \lvert X_L \rvert \geq \sqrt{R_L(R_G - R_L)}$	normal and reversed
$R_G > R_L, \quad \lvert X_L \rvert < \sqrt{R_L(R_G - R_L)}$	normal only
$R_G < R_L, \quad \lvert X_G \rvert \geq \sqrt{R_G(R_L - R_G)}$	normal and reversed
$R_G < R_L, \quad \lvert X_G \rvert < \sqrt{R_G(R_L - R_G)}$	reversed only

(9)

So let's look at an example for the design of an L-Section matching netwook for a conjugate match of the load impedance $Z_L = 200 + j50\Omega$ to the generator $Z_G = 50 - j10\Omega$ at $f = 30$ MHz. Note that $\sqrt{R_G(R_L - R_G)} = \sqrt{50(200-50)} = 86.6 > \lvert X_G \rvert$ so that only the last of the four conditions applies. The two solutions from Eq. (7) for the reversed L-Section are:

$$\begin{array}{lll} X_1 = 136.85 & X_2 = -80.13 & Q = 1.803 \\ X_1 = -103.52 & X_2 = 100.13 & Q = 1.803 \end{array} \qquad (10)$$

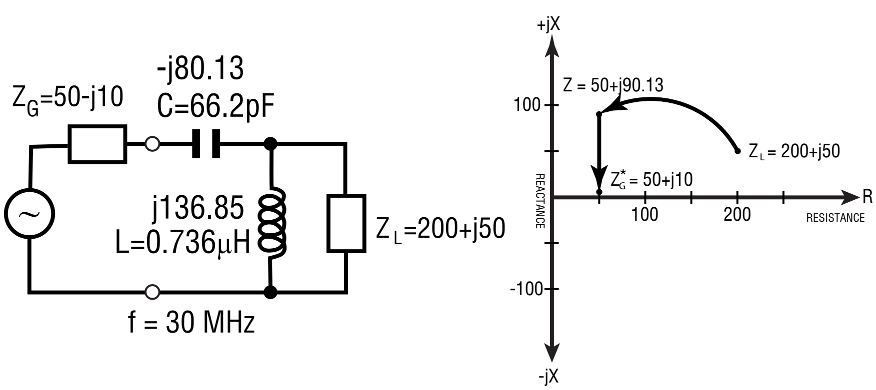


Figure 6 The First L-Section Solution

Figure 6 shows the first solution in Eq. (10). The parallel inductor transforms the load impedance counter-clockwise around the constant conductance circle to a point $Z = 50 + j90.13$, and then the capacitor transforms this intermediate impedance vertically downward to the final impedance of $Z = Z_G^* = 50 + j10$ that is the complex conjugate of the generator impedance.

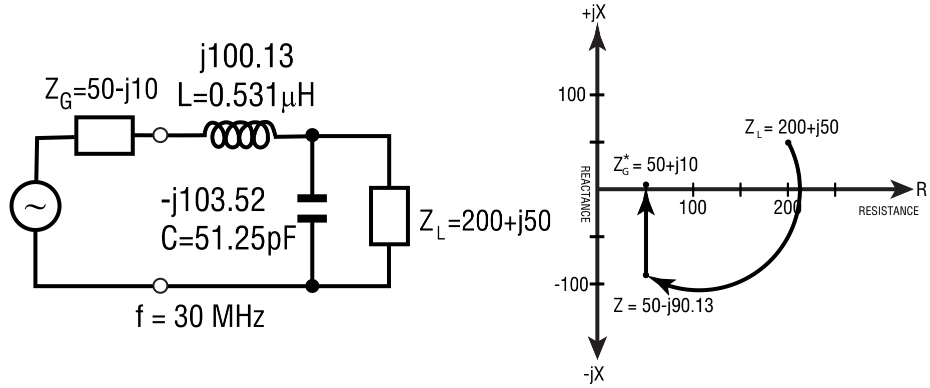


Figure 7 The Second L-Section Solution

Figure 7 shows the second solution in Eq. (10). Here the parallel capacitor transforms the load impedance clockwise around the constant conductance circle in the impedance plane to a point $z = 50 - j90.13$, and then the inductor transforms this impedance vertically upward to the final impedance which, once again, is the complex conjugate of the generator impedance.

Notice the Q of the matching network is the same for both solutions and is determined by the generator and load impedances according to Eq. (7). Of course the analysis up to this point has ignored the resistive losses in both the inductor and capacitor.

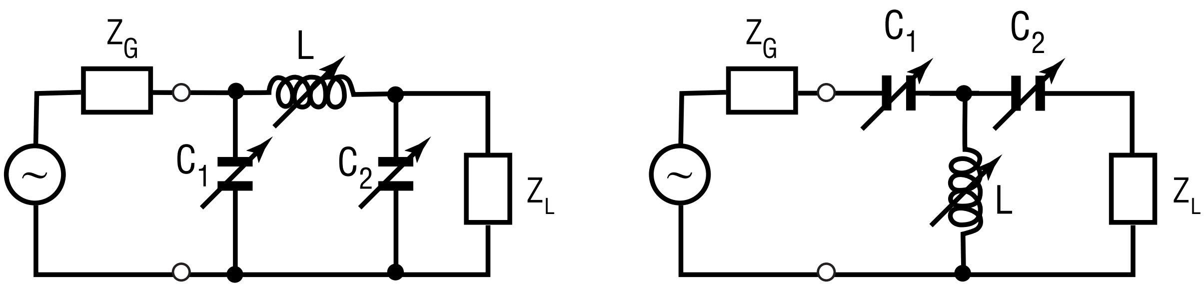
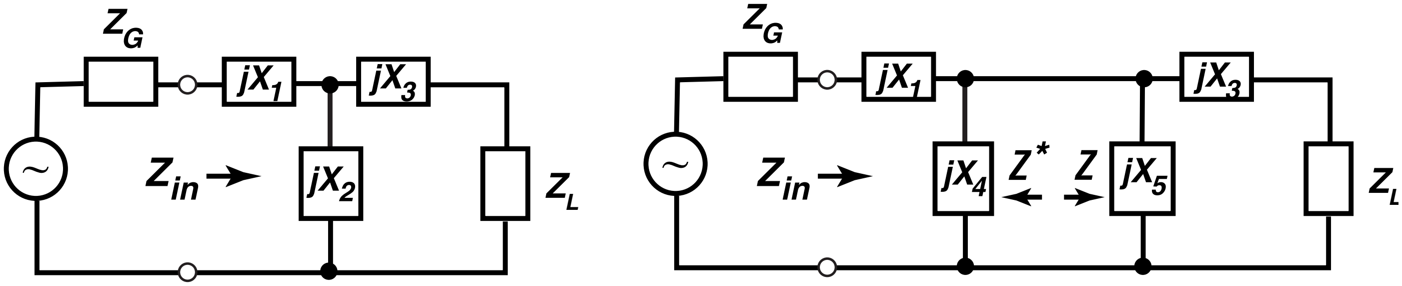


Figure 8 Π-Section and T-Section Networks

One can also consider networks that use three reactive elements for impedance matching. Figure 8 illustrate two that have been used extensively in amateur radio. The first is a Π-Section which was



(a) (b)

Figure 9 A T-Section Network

frequently used with vacuum tube transmitters to couple the final amplifier to the antenna. The second is a T-Section network that is often seen in recent antenna tuners. We shall consider the T-Tuner in more detail.

Let's consider the design procedure suggested in Figure 9. The T-Section in Figure 9(a) can be thought of as two L-Sections arranged back to back as in Figure 9(b), by splitting the parallel reactance into two parts: $.X_2 = X_4 \parallel X_5$. An additional degree of freedom is introduced into the design by an intermediate reference impedance, say $Z = R + jX$, such that looking into the right L-Section the input impedance is *Z*, and looking into the left L-Section, it is *Z**.

In order for the two L-Sections in Figure 9(b) to always have a solution, the resistive part of Z must satisfy the conditions of Eq. (9). So we must choose $R > R_G$ and $R > R_L$, or equivalently:

$$R > R_{\max} \qquad R_{\max} = \max(R_G, R_L) \tag{11}$$

Other than this, the value of *Z* is arbitrary. Since each L-Section has two solutions, there are actually four possible values for X1, X3, X4, and X5, but we will select the two solutions that produce capacitors for X_1 and X_3, and inductors for X_4 and X_5. So let's go back to the previous example with the load impedance $Z_L = 200 + j50$, the generator impedance $Z_G = 50 - j10$, and a frequency $f = 30$ MHz. We arbitrarily choose $Z = 250 + j25$ and $Z^* = 250 - j25$. Solving Eq. (7) twice, we find:

$$\begin{aligned} X_1 &= -90.623 \qquad X_4 = 119.529 \qquad Q = 3.012 \\ X_3 &= -152.47 \qquad X_5 = 612.348 \qquad Q = 0.512 \end{aligned} \tag{12}$$

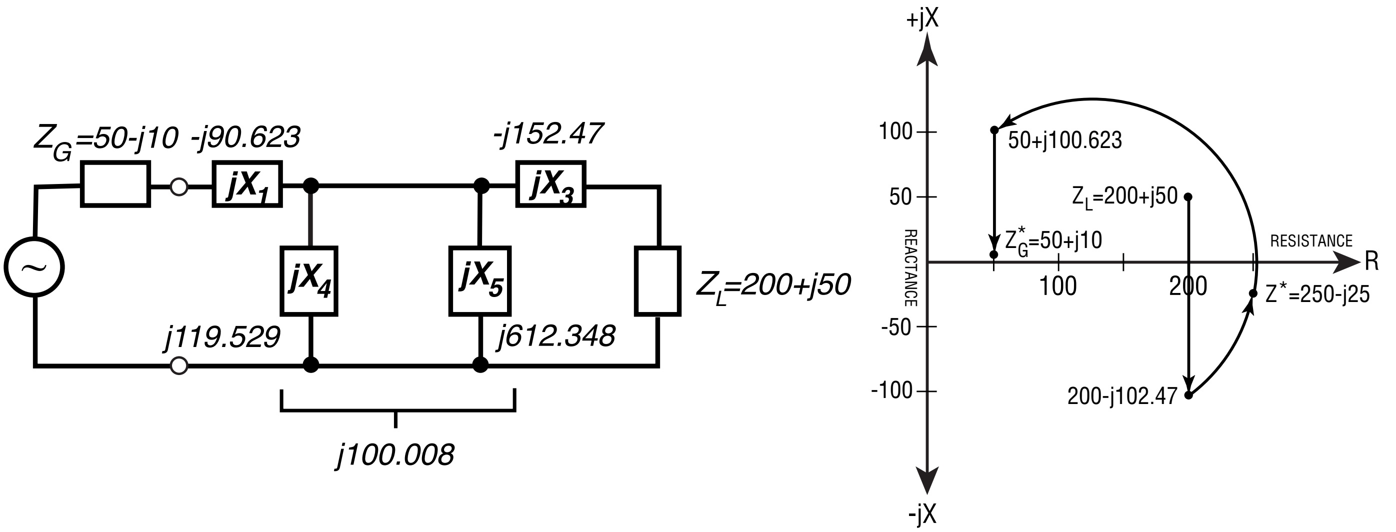


Figure 10 The T-Section Solution for Z=250+j25

Figure 10 illustrates the results for the solutions in Eq. (12). One can follow the transformation of the load impedance in the impedance plane; First the series capacitive reactance X_3 moves the impedance vertically downward followed by the parallel inductive reactance, X_5, moving it counter-clockwize along the constant conductance circle to *Z*=250-j25.* Next the parallel inductive reactance, X4, moves it further counter-clockwise around the constant conductance circle.

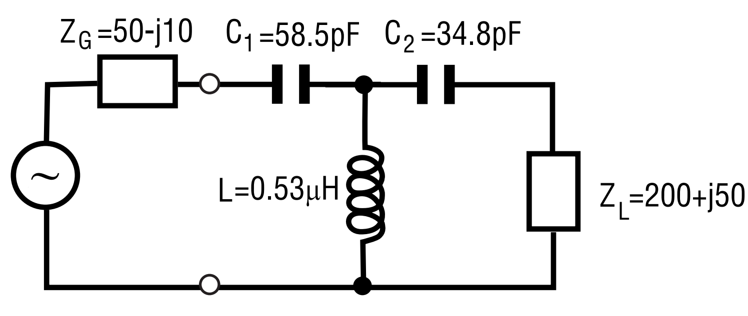


Figure 11 The final solution for the T-Section

Finally, the series capacitive reactance, X1, moves it to the point $Z_G^* = 50 + j10$. The two parallel inductive reactances, X_4 and X_5 can be combined with an effective value of *j100.008*. Knowing the reactances of the three components in the T-Section, one can use *f=30* MHz to compute their capacitance and inductance which are shown in Figure 11. Looking at the transformation of the load impedance depicted in the right side of Figure 10 one can see that any number of choices for *Z* with the *Re(Z)>200* would have produced a similar solution.

Since there is a large number of choices for the three reactive components, what is the best strategy for the user of a T-Section Tuner for selecting their values? The best choice depends on a fact we have ignored at this point which is the power lost in real components rather than the ideal components we have considered in the design. The capacitors used in most antenna tuners use air dielectrics and have very little loss, so we can probably ignore it. Most unwanted loss will come from the inductor which is often a roller-type adjustable coil.

The resistance of a coil will increase with its length, or number of turns. However, the inductance of the coil increases with the square of the number of turns. So as we adjust the coil by increasing its length, the reactance increases faster than the resistance, and the Q will be larger for larger inductance. Some have reported that the unloaded Q of a roller inductor may be in the 100 to 150 range when the number of turns used is large, but may be only in the 20 to 50 range when only a few turns are used (such as is the case when working in the 10-meter band.)[3] Therefore one wants to use the largest inductor possible to minimize the loss.

Since C_1 is in series with the source, all the transmitter current flow through this element. Therefore it is desirable to minimize the impedance by making the capacitor value as large as possible. Finally C_2 can be viewed as "coupling" the network to the antenna. Thus as the value of C_2 is made smaller, the reactance increases and the antenna is further decoupled from the rest of the matching network forcing more current though the inductor which will increase the circuit loss. So to minimize the loss it is best to minimize the reactance of C_2 and maximize the reactance of L.

This, then, indicates the proper method for adjusting a tuner to minimize losses. The procedure can be summarized in the following steps[3,4,5]:

1. Set L to the largest inductance (largest possible X_L)
2. Set C_1 and C_2 to the largest capacitance (smallest possible X_{C1}, X_{C2})
3. Adjust C_1 for best match. If SWR doesn't drop, leave it at maximum capacitance
4. Adjust C_2 for best match. If SWR drops, alternately adjust C_1 and C_2
5. If no acceptable match, reduce L slightly and go to step 2.

Differential T-Section.

A variation of the T-Section antenna tuner is called the "Differential T-Tuner" whose schematic is shown in Figure 12. Here the two capacitors are built as one unit such that as one capacitor increases in value, the other capacitor decreases. Some popular tuners that use this differential T-Section are the MFJ-986, AT-Auto, AT-500, AT-2KD,,and the HF-Auto[6].

The major advantage is that tuning is accomplished with just two adjustable components instead of three, and the minimum VSWR is obtained with just one setting which simplifies its use. The differen-

[3] "Impedance Matching, Part 2: Basic Principles," Section 6. David Knight G3YNH and Nigel Williams G3GFC, http://www.g3ynh.info/zdocs/z_matcing/part_2.html

[4] "Antenna Notes for a Dummy," Walt Fair, Jr., W5ALT, http://www.comportco.com/~w5alt/antennas/notes/ant-notes.php?pg=10

[5] "Getting the Most Out of Your T-Network Antenna Tuner," Andrew S. Griffith, W4ULD, QST Magazine, ARRL, Newington, CT., January 1995, pp 44-47.

[6]. The AT-Auto was originally sold by PalStar and is now supported by Don Kessler Engineering. The AT-500, AT-2KD, and HF-Auto are PalStar's current Tuners. The MFJ-986 is sold by MFJ Enterprises, Inc. Internet links for these vendors are: http://www.mfjenterprises.com/Product.php?productid=MFJ-986, http://kesslerengineeringllc.com/tuners.htm, and http://www.palstar.com.

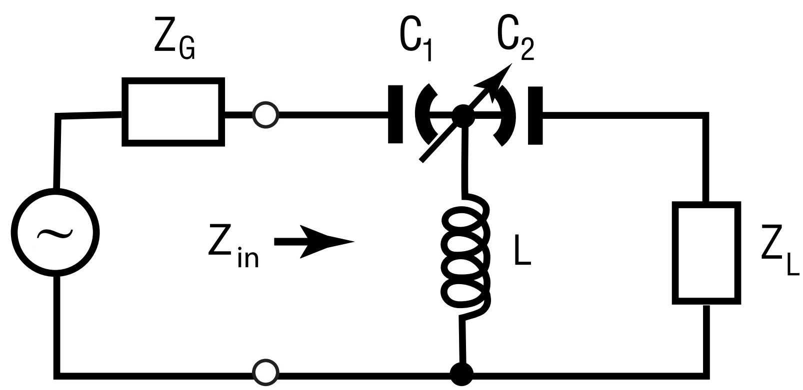


Figure 12 Differential T-Section

tial capacitor removes one degree of freedom in matching the two impedances which suggests that a smaller range of load impedances can be matched with this type of tuner although the AT-Auto specification is an Impedance range of 15 to 1500 Ω from the160m to 6m amateur radio bands. As a practical matter, the instruction manuals for these tuners suggest a change in transmission line length (that changes the impedance presented to the tuner) if an acceptable VSWR cannot be achieved.

The other disadvantage is that a less than optimum value for the inductance may be required for the impedance match that could increase the power losses in the network with a loss in efficiency.

The AT-Auto specifies a 340pF - 14pF - 340pF differential capacitor, and the HF-Auto specifies a 470pF - 10pF - 470pF capacitor. A 26 μH roller-inductor is specified for the AT-Auto. In order to model the Differential T-Section the reactances of the two capacitors can be written:

$$X_{C1}(x) = \frac{1}{\omega(C_{\min} + C_{\Delta} x)}$$
$$X_{C2}(x) = \frac{1}{\omega[C_{\min} + C_{\Delta}(1 - x)]} \qquad (13)$$
$$X_{\ell}(L) = \omega L$$

where ω*=2πf MHz*, *C_{min}=14pF,* and *C_{Δ}=326pF* in the case of the AT-Auto. Using $Z_{C1}(x) = -jX_{C1}$, $Z_{C2}(x) = -jX_{C2}$, and $Z_{\ell}(L) = jX_{\ell}(L)$ we may write an expression for the impedance looking into the Differential T-Section:

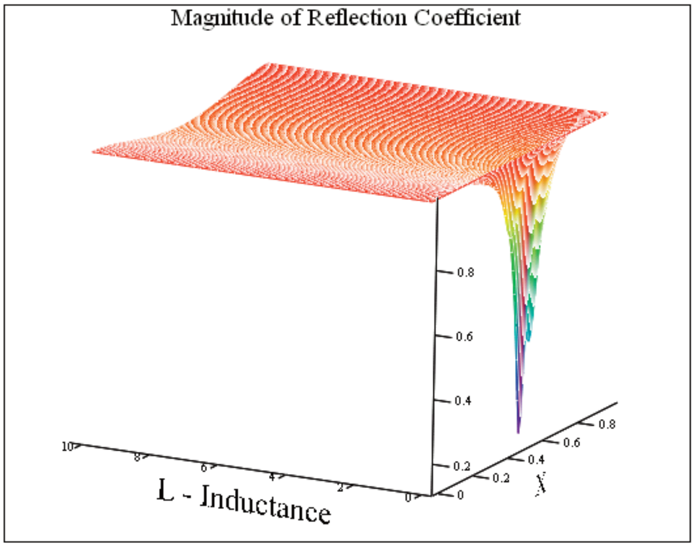
$$Z_{in}(x,L) = Z_{C1}(x) + \frac{Z_{\ell}(L) \cdot (Z_{C2}(x) + Z_L)}{Z_{\ell}(L) + Z_{C2}(x) + Z_L} \qquad (14)$$

Next we may solve for the reflection coefficient, ρ*(x,L)*, and the standing wave ratio, *VSWR(x,L)*:

$$\rho(x,L) = \frac{Z_{in}(x,L) - Z_G}{Z_{in}(x,L) + Z_G}$$
$$VSWR(x,L) = \frac{1 + |\rho(x,L)|}{1 - |\rho(x,L)|} \qquad (15)$$

In principle we can solve Eqs. (13) - (15) analytically, but fortunately we can simplify this using the computational methods of programs like Mathcad[7] and search through the variable ranges: $0 \le x \le 1$ and $0.1\mu H \le L \le 26\mu H$ (for the AT-Auto example) for a minimum VSWR. Figure 13 show a 3D plot of the magnitude of the reflection coefficient for a $Z_L = 5 + j50$, and one can see a sharp minimum giving a VSWR=1 for a value of x=0.577 and L=0.219μH. Figure 14 illustrates the three movements in the

[7] PTC, 140 Kendrick Street, Needham, MA 02494, (781) 370-5000, http://www.ptc.com/product/mathcad.



B Z_G=50+j0 Z_L=5+j50 *f=14.1 MHz*

x=0.577 L=0.219µH

Figure 13 *|ρ(x,L)| vs. x and L*

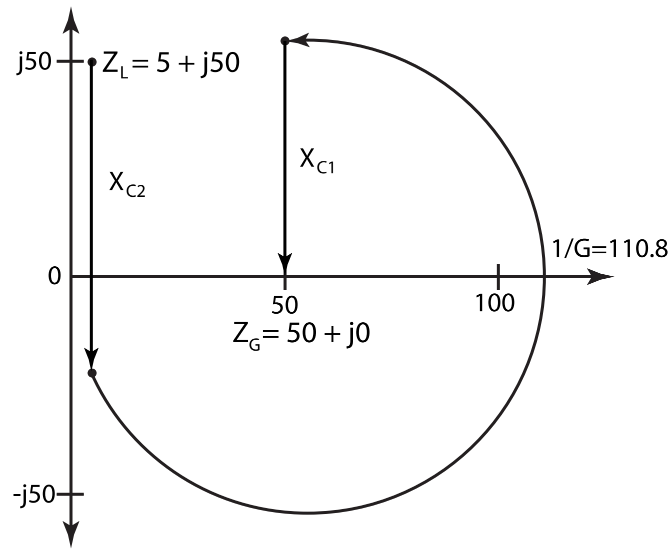
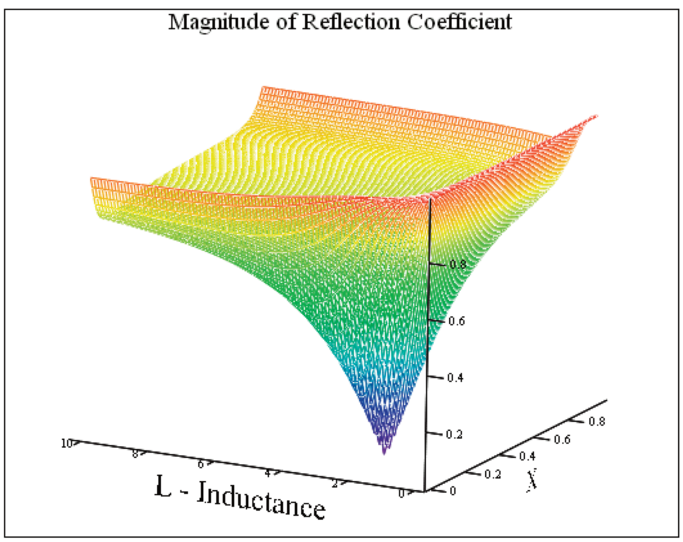


Figure 14 Impedance Plane for Z_L=5+j50

impedance plane. Starting at a $Z_L = 5 + j50$ capacitor C_2 transforms the impedance straight downward intersecting a constant conductance circle with a $\frac{1}{G} = 110.8\Omega$. Next the parallel inductive reactance moves the impedance counter-clockwise around



B Z_G=50+j0 Z_L=500+j100 *f=14.1 MHz*

x=0.181 L=1.971µH

Figure 15 *|ρ(x,L)| vs. x and L*

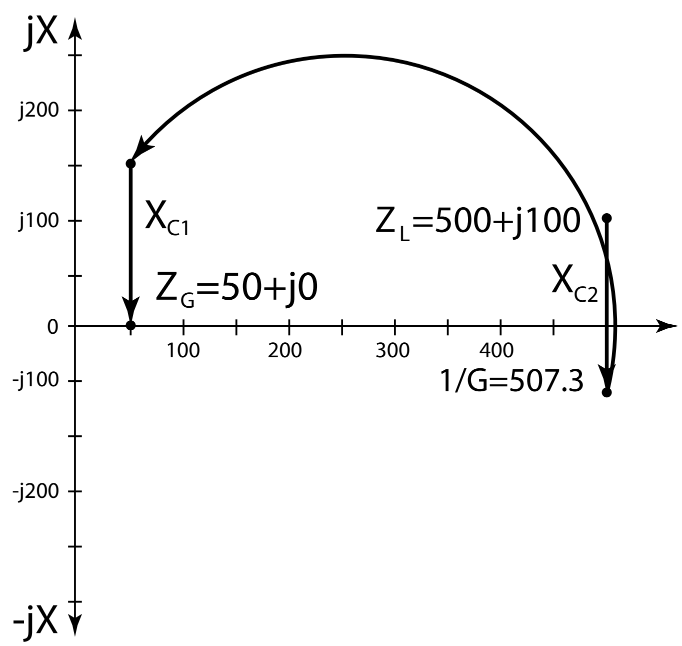


Figure 16 Impedance Plane for Z_L=500+j100

the constant conductance circle followed by the final vertical drop from C1 to a value of $Z = 50\Omega$. Another example is shown in Figure 15 for a $Z_L = 500 + j100$. Now the values for a VSWR=1 are x=0.181 and L=1.971µH. *Figure 16 show the movements in the impedance plane that this time involve a constant conductance circle with a* $\frac{1}{G} = 507.3\Omega$.

Most of the losses present in T-Section tuners are in the inductor and we can make a change to our equations to take this into account. The effective series resistance of the inductor, R_ℓ is related to the inductive reactance by Q_ℓ

$$Q_\ell = \frac{|X_\ell(L)|}{R_\ell} \tag{16}$$

So referring to Reference (3) which says, "...that the unloaded Q of a roller inductor may be in the 100 to 150 range when the number of turns used is large, but may be only in the 20 to 50 range when only a few turns are used," we will assume:

$$Q_\ell = 50 + 4L \tag{17}$$

where L is in $\mu\mathrm{H}$ for a maximum $Q_\ell^{Max} \approx 150$. Now use the following expression for the inductive impedance.

$$Z_\ell(L) = \frac{|X_\ell(L)|}{50+4L} + jX_\ell(L) \tag{18}$$

If P_{in} is the power delivered by the "generator" in Figure 12, we may solve for the currents in the three components (note that $I_{in}^*(x,L)$ is the complex conjugate of the input current):

$$P_{in} = \mathrm{Re}[V_{in}(x,L) I_{in}^*(x,L)] = \mathrm{Re}[V_{in}(x,L)\frac{V_{in}^*(x,L)}{Z_{in}^*(x,L)}] = |V_{in}(x,L)|^2 \mathrm{Re}[\frac{1}{Z_{in}^*(x,L)}] = \frac{|V_{in}(x,L)|^2}{|Z_{in}(x,L)|^2}\mathrm{Re}[Z_{in}(x,L)]$$

$$|V_{in}(x,L)| = |Z_{in}(x,L)| \cdot \sqrt{\frac{P_{in}}{\mathrm{Re}[Z_{in}(x,L)]}}$$

$$I_{C1}(x,L) = \frac{V_{in}(x,L)}{Z_{in}(x,L)} \tag{19}$$

$$I_{C2}(x,L) = I_{C1}(x,L) \cdot \frac{Z_\ell(L)}{Z_\ell + Z_{C2}(x) + Z_L}$$

$$I_\ell(x,L) = I_{C1}(x,L) - I_{C2}(x,L)$$

(Note that we are taking $V_{in}(x,L) = |V_{in}(x,L)| \angle 0^\circ$ as the reference phase for the power calculations.)
Next we can solve for the power dissipated in the inductor (where I_ℓ^* is the complex conjugate of I_ℓ)

$$P_\ell(x,L) = \mathrm{Re}[V_\ell(x,L) I_\ell^*(x,L)]$$

$$P_\ell(x,L) = \mathrm{Re}[I_\ell(x,L) Z_\ell(x,L) I_\ell^*(x,L)] \tag{20}$$

$$P_\ell(x,L) = |I_\ell(x,L)|^2 \cdot \mathrm{Re}[Z_\ell(x,L)]$$

and we can calculate the efficiency, η:

$$\eta(x,L) = \frac{P_{in} - P_\ell(x,L)}{P_{in}} \tag{21}$$

Finally, we can solve for the magnitudes of the capacitor and coil peak voltages:

$$\begin{aligned} |V_{C1}^{Peak}(x,L)| &= |I_{C1}(x,L) \cdot X_{C1}(x,L)| \cdot \sqrt{2} \\ |V_{C2}^{Peak}(x,L)| &= |I_{C2}(x,L) \cdot X_{C2}(x,L)| \cdot \sqrt{2} \\ |V_\ell^{Peak}(x,L)| &= |I_\ell(x,L) \cdot X_\ell(x,L)| \cdot \sqrt{2} \end{aligned} \tag{22}$$

Table 1 shows a number of calculations with different frequencies and load impedances for the Differential T-Section with $Z_G = 50 + j0$ and $P_{in} = 1500$ Watts. Several trends in the calculations can be seen:

Table 1 Differential T-Section with Z_G=50+j0 and P_{in}=1500 Watts

f (Mhz)	*Load Z_L*	*x*	*L (μH)*	*Efficiency*	*V_{c1}*	*V_{c2}*	*V_L*
1.8	10+j0	0.336	22.131	69.9	5544	5556	5558
	20+j0	0.410	22.173	78.9	4650	4659	4664
	50+j0	0.521	22.304	86.4	3736	3739	3756
	100+j0	0.608	22.500	89.8	3253	3235	3276
	200+j0	0.685	22.871	92.2	2882	2816	2913
	500+j0	0.765	23.783	94.1	2572	2321	2607
	1000+j0	0.785	25.436	95.3	2427	1781	2456
3.5	10+j0	0.336	5.894	70.8	2845	2877	2880
	20+j0	0.410	5.932	79.6	2836	2407	2417
	50+j0	0.520	6.062	87.0	1921	1926	1959
	100+j0	0.601	6.259	90.4	1679	1641	1722
	200+j0	0.667	6.619	92.6	1522	1382	1570
	500+j0	0.693	7.697	94.3	1470	949	1522
	1000+j0	0.586	9.942	94.7	1719	515	1762
7	10+j0	0.327	1.504	79.4	1460	1503	1511
	20+j0	0.406	1.549	86.0	1203	1244	1264
	50+j0	0.514	1.681	91.3	970	976	1044
	100+j0	0.577	1.888	93.6	872	793	954
	200+j0	0.582	2.303	95.0	864	571	947
	500+j0	0.422	3.534	95.0	1161	268	1223
	1000+j0	0.287	4.977	93.8	1639	155	1684
14	10+j0	0.334	0.407	88.0	717	799	815
	20+j0	0.421	0.454	92.1	583	659	699
	50+j0	0.511	0.586	95.2	488	495	623
	100+j0	0.492	0.802	96.4	505	340	636
	200+j0	0.353	1.162	96.3	682	192	784
	500+j0	0.196	1.806	94.7	1132	98	1196
	1000+j0	0.123	2.499	92.9	1625	63	1670
21	10+j0	0.362	0.205	91.6	445	566	589
	20+j0	0.457	0.252	94.5	360	472	529
	50+j0	0.511	0.383	96.7	325	333	506
	100+j0	0.384	0.578	97.2	422	190	573
	200+j0	0.228	0.806	96.5	664	109	769
	500+j0	0.117	1.213	94.6	1128	60	1192
	1000+j0	0.068	1.667	92.5	1621	40	1666
28	10+j0	0.408	0.136	93.4	299	460	489
	20+j0	0.508	0.181	95.7	245	391	458
	50+j0	0.512	0.312	97.5	244	251	457
	100+j0	0.296	0.465	97.6	398	126	555
	200+j0	0.162	0.616	96.6	660	75	765
	500+j0	0.077	0.913	94.5	1127	43	1191
	1000+j0	0.040	1.251	92.3	1619	29	1664
50	10+j0	-	-	-	-	-	-
	20+j0	0.676	0.109	96.9	105	321	401
	50+j0	0.517	0.250	98.6	135	143	410
	100+j0	0.154	0.387	97.8	384	60	545
	200+j0	0.072	0.035	96.6	657	38	762
	500+j0	0.024	0.514	94.4	1125	23	1190
	1000+j0	0.004	0.700	92.0	1616	16	1661

1. Matches were achieved at all frequencies for real load impedances from 10Ω to 1000Ω except for f=50 MHz which failed at 10Ω.
2. The required inductance is larger for lower frequencies.
3. The efficiency is lower for lower load resistances. In fact at the lowest efficiencies the inductors could be dissipating several hundred Watts and could be a concern.
4. The voltages across the capacitors and inductor can be several thousand volts and adequate insulation should be provided to prevent breakdown at high powers.

Although there is an infinite number of complex load impedances to consider, Table 2 shows two ranges calculated at 7 MHz. Once again we see the efficiency suffers for loads with small resistive values, and no matching solution was found for a $Z_L=10+j500\Omega$. Also, the peak component voltages are generally larger for the load with $R_L=10\Omega$.

Table 2 Differential T-Section showing result with reactive loads, $Z_G=50+j0$ and $P_{in}=1500$ Watts

f (Mhz)	*Load Z_L*	*x*	*L (μH)*	*Efficiency*	*Vc1*	*Vc2*	*V_L*
7	10-j500	0.038	7.814	46.2	6692	817	6702
	10-j200	0.104	3.969	60.3	3676	1000	3692
	10-j100	0.168	2.671	68.7	2555	1145	2584
	10-j50	0.229	2.050	73.9	1988	1275	2025
	10+j0	0.327	1.504	79.4	1460	1503	1511
	10+j50	0.465	1.119	84.0	1064	1914	1132
	10+j100	0.598	0.921	86.6	844	2527	929
	10+j200	0.760	0.780	88.7	672	4023	778
	10+j500	-	-	-	-	-	-
7	1000-j500	0.243	5.201	92.3	1890	145	1929
	1000-j200	0.275	4.940	93.3	1699	152	1742
	1000-j100	0.282	4.934	93.5	1661	154	1706
	1000-j50	0.285	4.950	93.7	1648	154	1693
	1000+j0	0.287	4.977	93.8	1639	155	1684
	1000+j50	0.288	5.017	93.9	1635	155	1680
	1000+j100	0.288	5.073	94.0	1634	155	1679
	1000+j200	0.285	5.220	94.1	1645	155	1690
	1000+j500	0.261	5.963	94.3	1776	150	1818

In sumary, we considered a number of impedance matching circuits involving reactive components. First were normal and reversed L-Sections, and we saw that one or the other of these could match an arbitraty load impedance to a generator impedance. The choice between these solutions depended on which impedance had the larger resistive component. The impedance transformation from the load to the generator could be visualized in the impedance plane with series reactances causing a vertical movement and a parallel reactance moved around a constant conductance circle.

Next we considered a C-L-C T-Section because this is frequencly used in modern amateur radio antenna tuners. The third reactive component gives an extra degree of freedom that not only can match arbitrary load impedances to arbitrary generator impedances, but also allow a choice of bandwidth or Q. Although a match of arbritary impedances is possible, such a match may be limited by the adjustment ranges of realistic inductors and capacitors. Three adjustable components makes tuning more complex because there is more than one set of component values that accomplish an impedance match, and there is usually one that provides a smaller loss, e.g. higher efficiency. A method was described to achieve an impedance match with the minimum loss.

Finally, we looked at a Differential T-Section in which the two capacitors are built together in such a way that as one capacitor increases in value the other one decreases. This has the advantage of simplifying the tuning procedure because there are only two components to adjust, and there will only be one possible set of values to achieve the impedance match. Calculations showed that a wide range of impedances may be matched to a generator of $Z_G = 50\Omega$. The disadvantages are that it may not be imposible to match some combination of impedances, and some matches may result in large losses with low efficiencies.