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 \( Z_s = 50 \, \Omega \)) 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. \( Z = Z_s^* \)).

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.

If we consider adding a parallel inductance or capacitance to the “load” impedance, it is useful to deal with the reciprocal of the impedance, i.e. \( 1/Z \), known as the admittance.

\[
Y_L = \frac{1}{R + jX} = \frac{R - jX}{(R + jX)(R - jX)} = \frac{R}{R^2 + X^2} - j \frac{X}{R^2 + X^2} \tag{1}
\]

which can be written:

\[
\begin{align*}
Y &= G + jB \\
G &= \frac{R}{R^2 + X^2} \\
B &= -\frac{X}{R^2 + X^2}
\end{align*} \tag{2}
\]

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 \( \Omega \).

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 \( B_P = -\frac{1}{X_P} \) in parallel with an admittance \( Y = G + jB \) which becomes

\[
Y' = G + j(B + B_P) \tag{1}.
\]

Since the conductance \( G = \frac{R}{(R^2 + X^2)} \) is constant regardless of the value of \( X_P \), it is true that the conductance is unchanged by the addition of a parallel reactance:

\[
G = G' = \frac{R'}{(R'^2 + X'^2)} \tag{3}
\]

which can be rewritten:

\[
R'^2 + X'^2 - \frac{R'}{G} = 0 \tag{4}
\]

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

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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 Cartesian coordinates.

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 + j0Ω 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 + j0Ω lies is known as the 20mS constant conductance circle (i.e. 20 milli-Siemans or \( \frac{1}{50} \) Siemans). Its radius is \( \frac{1}{2} G = 25Ω \) and its center lies at \( 25 + j0Ω \). It crosses the resistance axis at 0 and at \( \frac{1}{G} = 50Ω \).

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,
i.e. regions A, B, C, and D in Figure 4, the first matching element is always a \textit{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 $50\Omega$, i.e. regions E and F in Figure 4, the first matching element is always a \textit{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. 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:

$$Z_{in} = Z_G^* = R_G - jX_G$$  \hspace{1cm} (5)

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

$$Z_{in} = \frac{Z_1 (Z_2 + Z_L)}{Z_1 + Z_2 + Z_L} \quad \text{(normal)}$$  \hspace{1cm} (6)

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

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 $X_1, X_2$ with solutions

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for the two types:

\[ X_1 = \frac{Z_L + R_G Q}{R_L} \]
\[ X_2 = -(Z_L + R_G Q) \] (normal)

\[ X_1 = \frac{Z_G + R_L Q}{R_G} \]
\[ X_2 = -(Z_G + R_L Q) \] (reversed)  \hspace{1cm} (7)

\[ Q = \sqrt{\frac{R_L}{R_G} - 1 + \frac{X_1^2}{R_G R_L}} \]
\[ Q = \sqrt{\frac{R_G}{R_L} - 1 + \frac{X_2^2}{R_G R_L}} \]

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:

\[ X_1 = \infty, \quad X_2 = -(Z_L + X_G) \]  \hspace{1cm} (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 \( \pi \)- 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:

<table>
<thead>
<tr>
<th>Existence Conditions</th>
<th>L-Section Types</th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_G &gt; R_L, \quad</td>
<td>X_L</td>
</tr>
<tr>
<td>( R_G &gt; R_L, \quad</td>
<td>X_L</td>
</tr>
<tr>
<td>( R_G &lt; R_L, \quad</td>
<td>X_G</td>
</tr>
</tbody>
</table>
| \( R_G < R_L, \quad |X_G| < \sqrt{R_G (R_L - R_G)} \) | reversed only |  \hspace{1cm} (9) 

So let’s look at an example for the design of an L-Section matching network for a conjugate match of the load impedance \( Z_L = 200 + j50 \)Ω to the generator \( Z_G = 50 - j10 \)Ω at \( f = 30 \) MHz. Note that

\[ \sqrt{R_G (R_L - R_G)} = \sqrt{50 (200 - 50)} = 86.6 > |X_G| \] so that only the last of the four conditions applies. The two solutions from Eq. (7) for the reversed L-Section are:

\[ X_1 = 136.85 \quad X_2 = -80.13 \quad Q = 1.803 \]
\[ X_1 = -103.52 \quad X_2 = 100.13 \quad Q = 1.803 \]  \hspace{1cm} (10)
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 $Z = 50 + j 90.13$, and then the capacitor transforms this intermediate impedance vertically downward to the final impedance of $Z = Z_G^* = 50 + j 10$ that is the complex conjugate of the generator impedance.

![Figure 7 The Second L-Section Solution](image)

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 $z = 50 - j 90.13$, 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 II-Section and T-Section Networks](image)

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 II-Section which was

![Figure 9 A T-Section Network](image)
frequently used with vacuum tube transmitters to couple the final amplifier to the antenna. The sec-
ond 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 reac-
tance into two parts: \( \cdot X_2 = X_4 \parallel X_5 \). 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 \( R > R_0 \) and \( R > R_L \), or equivalently:

\[
R > R_{\text{max}} \quad R_{\text{max}} = \max(R_0, R_L)
\]  

Other than this, the value of \( Z \) is arbitrary. Since each L-Section has two solutions, there are actually 
four possible values for \( X_1, X_3, X_4, \) and \( X_5 \), but we will select the two solutions that produce capaci-
tors 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 \( Z_0 = 50 - j10 \), and a frequency \( f = 30 \text{ MHz} \). 
We arbitrarily choose \( Z = 250 + j25 \) and \( Z^* = 250 - j25 \). Solving Eq. (7) twice, we find:

\[
\begin{align*}
X_1 &= -90.623 \quad X_4 = 119.529 \quad Q = 3.012 \\
X_3 &= -152.47 \quad X_5 = 612.348 \quad Q = 0.512
\end{align*}
\]  

Figure 10 illustrates the results for the solu-
tions in Eq. (12). One can follow the transfor-
mation 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-clockwise 
along the constant conductance circle to 
\( Z^* = 250 - j25 \). Next the parallel inductive reac-
tance, \( X_4 \), moves it further counter-clockwise 
around the constant conductance circle.

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

Figure 11 The final solution for the T-Section
Finally, the series capacitive reactance, X1, moves it to the point \( Z' = 50 + j10 \). 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). 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-

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\(^6\) 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.
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 the 160m 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:

\[
X_{C1}(x) = \frac{1}{\omega(C_{\text{min}} + C_{\Delta}x)}
\]

\[
X_{C2}(x) = \frac{1}{\omega(C_{\text{min}} + C_{\Delta}(1-x))}
\]

\[
X_t(L) = \omega L
\]

where \( \omega = 2\pi f \text{ MHz} \), \( C_{\text{min}} = 14 \text{pF} \), and \( C_{\Delta} = 326 \text{pF} \) in the case of the AT-Auto. Using \( Z_{c1}(x) = -jX_{c1} \), \( Z_{c2}(x) = -jX_{c2} \), and \( Z_t(L) = jX_t(L) \) we may write an expression for the impedance looking into the Differential T-Section:

\[
Z_{in}(x,L) = Z_{c1}(x) + \frac{Z_t(L) \cdot (Z_{c2}(x) + Z_L)}{Z_t(L) + Z_{c2}(x) + Z_L}
\]  

(14)

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

\[
\rho(x,L) = \frac{Z_{in}(x,L) - Z_G}{Z_{in}(x,L) + Z_G}
\]

\[
VSWR(x,L) = \frac{1 + |\rho(x,L)|}{1 - |\rho(x,L)|}
\]  

(15)

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

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 the constant conductance circle followed by the final vertical drop from $C_1$ 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 $L=1.971\mu H$. 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 $Q_e$

$$Q_e = \left| \frac{X_e(L)}{R_e} \right|$$  \hspace{1cm} (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:

$$Q_e = 50 + 4L \hspace{1cm} (17)$$

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

$$Z_e(L) = \left| \frac{X_e(L)}{50 + 4L} \right| + jX_e(L) \hspace{1cm} (18)$$

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):

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

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

$$I_{c1}(x,L) = \frac{V_{in}(x,L)}{Z_{in}(x,L)}$$  \hspace{1cm} (19)

$$I_{c2}(x,L) = I_{c1}(x,L) \cdot \frac{Z_e(L)}{Z_e + Z_{c2}(x) + Z_{in}}$$

$$I_e(x,L) = I_{c1}(x,L) - I_{c2}(x,L)$$

(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 $I_e$ is the complex conjugate of $I_e$)

$$P_e(x,L) = \text{Re}[V_e(x,L)I_e(x,L)]$$

$$P_e(x,L) = \text{Re}[I_e(x,L)Z_e(x,L)I_e^*(x,L)]$$

$$P_e(x,L) = |I_e(x,L)|^2 \cdot \text{Re}[Z_e(x,L)]$$  \hspace{1cm} (20)

and we can calculate the efficiency, $\eta$:

$$\eta(x,L) = \frac{P_{in} - P_e(x,L)}{P_{in}}$$  \hspace{1cm} (21)

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

$$|V_{c1}^{\text{Peak}}(x,L)| = |I_{c1}(x,L) \cdot X_{c1}(x,L)| \cdot \sqrt{2}$$

$$|V_{c2}^{\text{Peak}}(x,L)| = |I_{c2}(x,L) \cdot X_{c2}(x,L)| \cdot \sqrt{2}$$

$$|V_e^{\text{Peak}}(x,L)| = |I_e(x,L) \cdot X_e(x,L)| \cdot \sqrt{2}$$  \hspace{1cm} (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**

<table>
<thead>
<tr>
<th>$f$ (Mhz)</th>
<th>Load $Z_L$</th>
<th>$x$</th>
<th>$L$ ($\mu H$)</th>
<th>Efficiency</th>
<th>$V_{c1}$</th>
<th>$V_{c2}$</th>
<th>$V_L$</th>
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</thead>
<tbody>
<tr>
<td>1.8</td>
<td>10+j0</td>
<td>0.336</td>
<td>22.131</td>
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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 $Z_L=10\,\Omega + j50\,\Omega$. 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

<table>
<thead>
<tr>
<th>$f$ (Mhz)</th>
<th>Load $Z_L$</th>
<th>$x$</th>
<th>$L$ ($\mu$H)</th>
<th>Efficiency</th>
<th>$Vc1$</th>
<th>$Vc2$</th>
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In summary, 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 arbitrary 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 frequently 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 arbitrary 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 impossible to match some combination of impedances, and some matches may result in large losses with low efficiencies.