Chapter 9

Broadband Antenna Matching

Antenna systems that provide a good impedance match to the transmitter over a wide frequency range have been a topic of interest to hams for many years. Most emphasis has been focused on the 80-meter band since a conventional half-wave dipole will provide better than 2:1 SWR over only about one-third of the 3.5 to 4.0 MHz band. The advantage of a broadband match is obvious—fewer adjustments during tune-up and an antenna tuner may not be required.

This chapter was written by Frank Witt, AI1H, who has written numerous articles in *QST* and in *The ARRL Antenna Compendium* series on this subject. See the Bibliography for details.

The term *broadband antennas* has frequently been used to describe antenna systems that provide a wideband impedance match to the transmitter. This is something of a misnomer, since most antennas are good radiators over a wide range of frequencies and are therefore "broadband antennas" by definition. The problem is getting the energy to the antenna so it can be radiated. Antenna tuners solve this problem in some cases, although losses in transmission lines, baluns and the antenna tuner itself can be excessive. Also worthy of mention is that antenna directional properties are usually frequency dependent. In this chapter, we discuss broadbanding the impedance match to the transmitter only.

GENERAL CONCEPTS

The Objective

Fig 1 shows a simplified system: the transmitter, an SWR meter and the transmitter load impedance. Modern transceivers are designed to operate properly into a 50-Ω load and they will deliver full power into the load impedance, at their rated level of distortion, when the SWR is less than about 1.5:1. Loads beyond this limit may cause the transceiver to protect itself by lowering the output power.

Many transceivers have built-in automatic antenna tuners that permit operation for loads outside the 1.5:1 SWR range, but the matching range is often limited, particularly on the lower-frequency bands. In practice, barefoot operation is simplified if the load SWR is held to less than 2:1.

For high-power amplifiers, it is also best to keep the load SWR to less than about 2:1, since output tuning components are commonly rated to handle such loads. You can see that the primary function of the SWR meter in Fig 1 is to measure the suitability of the load impedance so far as the transmitter is concerned. Henceforth, we will use the term *load SWR* as a description of the transmitter's load.

The SWR meter is actually reading a circuit condition at a single point in the system. In fact, the meter really measures magnitude of the reflection coefficient, but is calibrated in SWR. The relationships between the complex load impedance, Z_L, the magnitude of the reflection coefficient, $|\rho|$, and the load SWR are as follows:

$$|\rho| = \left| \frac{Z_L - 50}{Z_L + 50} \right| \qquad \text{(Eq 1)}$$

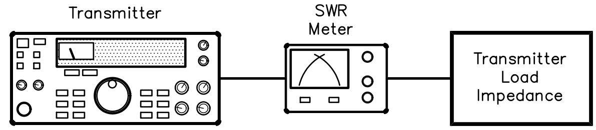

Fig 1—Basic antenna system elements at the output of a transmitter.

$$\mathrm{SWR} = \frac{1+|\rho|}{1-|\rho|} \quad \text{(Eq 2)}$$

where 50 Ω is used as the reference impedance, since the design load impedance for the transmitter is assumed to be 50 Ω.

An important point is that you need not know the output impedance of the transmitter to design a broadband matching network. The issue of the actual value of the output impedance of typical RF power amplifiers is a matter of continuing controversy in amateur circles. Fortunately, this issue is not important for the design of broadband matching networks, since the load SWR is independent of the output impedance of the transmitter. Our objective is to design the matching network so that the load SWR (with a 50-Ω reference resistance) is less than some value, say 2:1, over as wide a band as possible.

Resonant Antennas

The broadband matching techniques described here apply to antennas operating near resonance. Typical resonant antennas include half-wave dipoles, quarter-wave verticals over a ground plane and full-wave loops. To design a broadband matching network, you must know the antenna feed-point impedance near resonance. **Fig 2** shows the antenna-impedance equivalent circuit near resonance. Although the series RLC circuit is an approximation, it is good enough to allow us to design matching networks that significantly increase the band over which a good match to the transmitter is achieved.

Note that the impedance is defined by F_0, the resonant frequency, R_A, the antenna resistance at resonance, and Q_A, the antenna Q. R_A is actually the sum of the radiation resistance and any loss resistance, including conductor losses and losses induced by surrounding objects, such as the ground below the antenna. R_A is frequency dependent, but it is sufficient to assume it is fixed during the matching network design process. Minor adjustments to the matching network will correct for the frequency dependence of R_A.

The antenna resistance and Q depend on the physical properties of the antenna itself, the properties of ground and the height above ground. Consider as an example an 80-meter horizontal half-wave dipole made from #12 wire located over average ground (dielectric constant = 13 and conductivity = 5 mS/m). **Figs 3** and **4** show how feed-point resistance and Q vary with height. It is clear from these figures that there are wide gyrations in antenna parameters. The better these parameters are known, the more successful we will be at designing the broadband matching network. For horizontal dipoles, Figs 3 and 4 may be used to get a good idea of antenna resistance and Q, so long as the height is scaled in proportion to wavelength. For example, a 160-meter dipole at a height of 100 feet would have about the same resistance and Q as an 80-meter dipole at 50 feet.

For optimum results, the resistance and Q can be

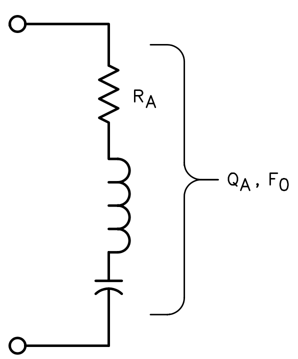

Fig 2—The equivalent circuit of a resonant antenna. The simple series RLC approximation applies to many resonant antenna types, such as dipoles, monopoles, and loops. R_A, Q_A and F_0 are properties of the entire antenna, as discussed in the text.

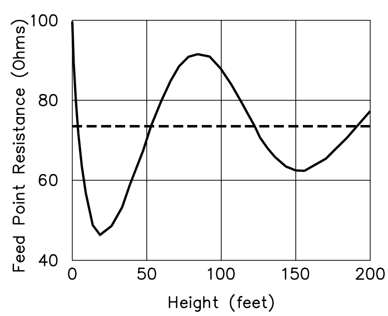

Fig 3—Feed-point resistance (solid line) for 80-meter horizontal dipole at resonance versus height over ground. The free-space value is shown as a dashed line.

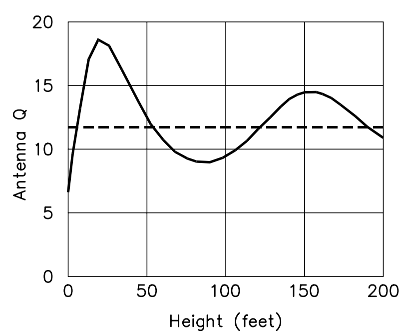

Fig 4—Antenna Q for 80-meter horizontal dipole (solid line) versus height. The free-space value is shown as a dashed line.

determined from computer simulation (using a program like *EZNEC*, for example) or measurement (using a low power SWR/Z meter such as the MFJ-259B or the Autek Research VA1 or RF-1). R_A can be computed using SWR measurements at resonance. **Fig 5** shows the typical bowl-shaped SWR curve as a function of frequency for a resonant antenna. S_0 is the SWR at resonance. You must take into account the loss of the feed line if the measurement must be made at the end of a long 50-Ω transmission line. If the line loss is low or if the measurement is made at the antenna terminals, the formulas given below apply.

For $R_A > 50\ \Omega$

$$R_A = S_0 \times 50 \qquad \text{(Eq 3)}$$

For $R_A < 50\ \Omega$

$$R_A = \frac{50}{S_0} \qquad \text{(Eq 4)}$$

There is an ambiguity, since we do not know if R_A is greater than or less than 50 Ω. A simple way of resolving this ambiguity is to take the same SWR measurement at resonance with a 10-Ω non-inductive resistor added in series with the antenna impedance. If the SWR goes up, $R_A >$ 45 Ω. If the SWR goes down, $R_A <$ 45 Ω. This will resolve the ambiguity.

Q_A may be determined by measuring the SWR bandwidth. Define BW_2 as the bandwidth over which the resonant antenna SWR is less than 2:1 (in the same units as F_0). See Fig 5.

For $R_A > 50\ \Omega$

$$Q_A = \frac{F_0}{BW_2}\sqrt{2.5 S_0 - S_0^{\,2} - 1} \qquad \text{(Eq 5)}$$

For $R_A < 50\ \Omega$

$$Q_A = \frac{F_0}{BW_2 \times S_0}\sqrt{2.5 S_0 - S_0^{\,2} - 1} \qquad \text{(Eq 6)}$$

Again, you must take into account the loss of the feed line if you make the measurement at the end of a long 50-Ω transmission line.

For horizontal and inverted-V half-wave dipoles, a very good approximation for the antenna Q when R_A is known is given by:

$$Q_A = \frac{93.9\left[\ln\dfrac{8110}{D F_0} - 1\right]}{R_A} \qquad \text{(Eq 7)}$$

where D = the diameter of wire, in inches.

Loss

When you build a resonant antenna with some matching scheme designed to increase the bandwidth, you might overlook the loss introduced by the broadband matching components. Loss in dB is calculated from:

$$\text{Loss} = -10\log\frac{\text{Power radiated by antenna}}{\text{Total power from transmitter}} \qquad \text{(Eq 8)}$$

or, alternatively, define efficiency in % as

$$\text{Efficiency} = \frac{100 \times \text{Power radiated by antenna}}{\text{Total power from transmitter}} \qquad \text{(Eq 9)}$$

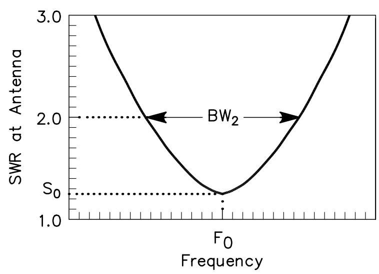

Fig 5—SWR at the antenna versus frequency in the vicinity of resonance.

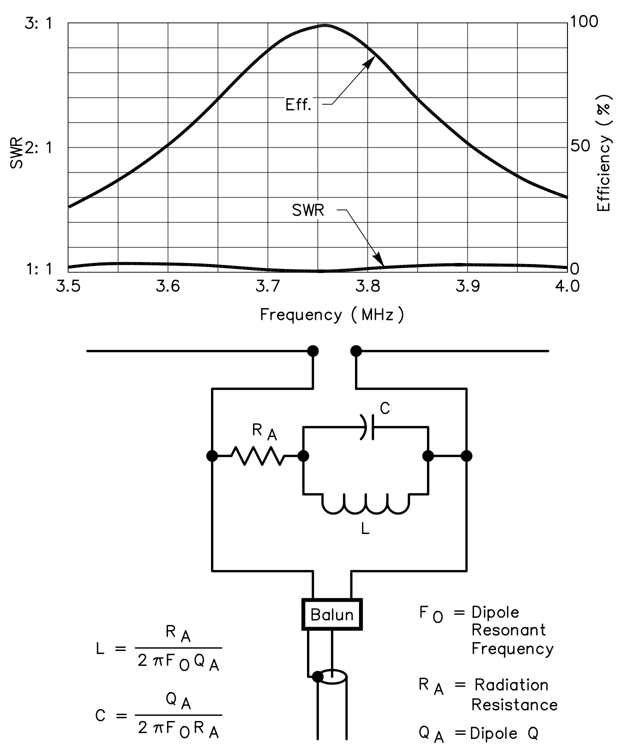

Fig 6—Matching the dipole with a complementary RLC network greatly improves the SWR characteristics, nearly 1:1 across the 3.5-MHz band. However, the relative loss at the band edges is greater than 5 dB.

An extreme degree of bandwidth broadening of an 80-meter dipole is illustrated in **Fig 6**. This approach is not recommended, but it is offered here to make a point. The broadening is accomplished by adding resistive losses. From network theory we obtain the RLC (resistor, inductor, capacitor) matching network shown. The network provides the complement of the antenna impedance. Note that the SWR is virtually 1:1 over the entire band (and beyond), but the efficiency falls off dramatically away from resonance.

The band-edge efficiency of 25 to 30% shown in Fig 6 means that the antenna has about 5 to 6 dB of loss relative to an ideal dipole. At the band edges, 70 to 75% of the power delivered down the transmission line from the transmitter is heating up the matching-network resistor. For a 1-kW output level, the resistor must have a power rating of at least 750 W! Use of an RLC complementary network for broadbanding is not recommended, but it does illustrate how resistance (or losses) in the matching network can significantly increase the apparent antenna SWR bandwidth.

The loss introduced by any broadband matching approach must be taken into consideration. Lossy matching networks will usually provide more match bandwidth improvement than less lossy ones.

SIMPLE BROADBAND MATCHING TECHNIQUES

The Cage Dipole

You can increase the match bandwidth of a single-wire dipole by using a thick radiator, one with a large diameter. The gain and radiation pattern are essentially the same as that of a thin-wire dipole. The radiator does not necessarily have to be solid; open construction such as shown in **Fig 7** may be used.

The theoretical SWR response of an 80-meter cage dipole having a 6-inch diameter is shown in **Fig 8**. BW_2 for this antenna fed with 50-Ω line is 287 kHz, and the Q is approximately 8. Its 2:1-SWR frequency range is 1.79 times broader than a dipole with a Q of 13, typical for thin-wire dipoles.

There are other means of creating a thick radiator, thereby gaining greater match bandwidth. The bow-tie and the fan dipole make use of the same Q-lowering principle as the cage for increased match bandwidth. The broadbanding techniques described below are usually more practical than the unwieldy cage dipole.

Stagger-Tuned Dipoles

A single-wire dipole exhibits a relatively narrow bandwidth in terms of coverage for the 3.5 to 4.0-MHz band. A technique that has been used for years to cover the entire band is to use two dipoles, one cut for the CW portion and one for the phone portion. The dipoles are connected in parallel at the feed point and use a single feeder. This technique is known as *stagger tuning*.

Fig 9 shows the theoretical SWR response of a pair of stagger-tuned dipoles fed with 50-Ω line. No mutual coupling between the wires is assumed, a condition

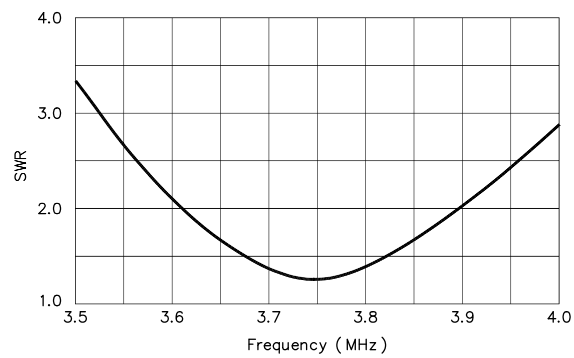

Fig 8—Theoretical SWR versus frequency response for a cage dipole of length 122 feet 6 inches and a spreader diameter of 6 inches, fed with 50-Ω line. The 2:1-SWR bandwidth frequencies are 3.610 and 3.897 MHz, with a resulting BW_2 of 287 kHz.

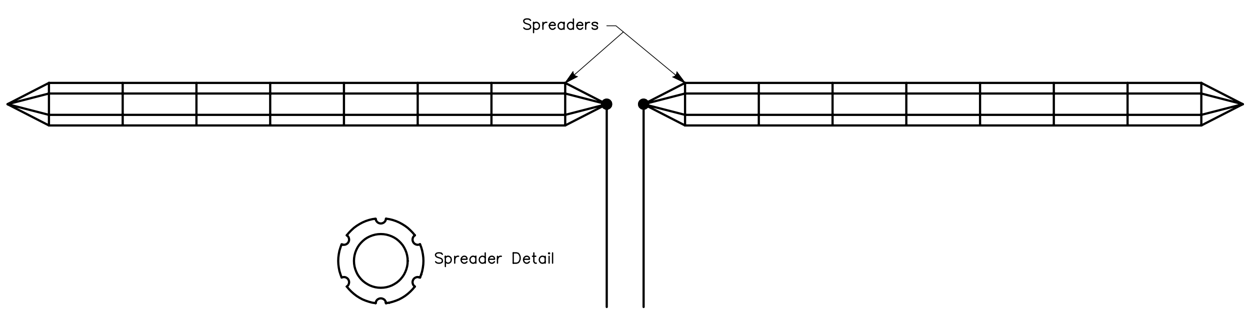

Fig 7—Construction of a cage dipole, which has some resemblance to a round birdcage. The spreaders need not be of conductive material, and should be lightweight. Between adjacent conductors, the spacing should be 0.02 λ or less. The number of spreaders and their spacing should be sufficient to maintain a relatively constant separation of the radiator wires.

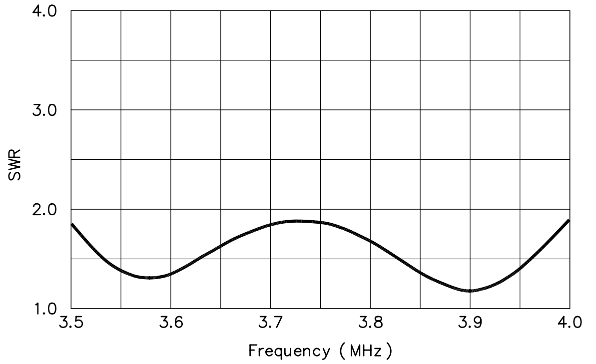

Fig 9—Theoretical SWR response of two stagger-tuned dipoles. They are connected in parallel at the feed point and fed with 50-Ω line. The dipoles are of wire such as #12 or #14, with total lengths of 119 and 132 feet.

that would exist if the two antennas were mounted at right angles to one another. As Fig 9 shows, the SWR response is less than 1.9:1 across the entire band.

A difficulty with such crossed dipoles is that four supports are required for horizontal antennas. A more common arrangement is to use inverted V dipoles with a single support, at the apex of each element. The radiator wires can also act as guy lines for the supporting mast.

When the dipoles are mounted at something other than a right angle, mutual coupling between them comes into play. This causes interaction between the two elements—tuning of one by length adjustment will affect the tuning of the other. The interaction becomes most critical when the two dipoles are run parallel to each other, suspended by the same supports, and the wires are close together. Finding the optimum length for each dipole for total band coverage can become a tedious and frustrating process.

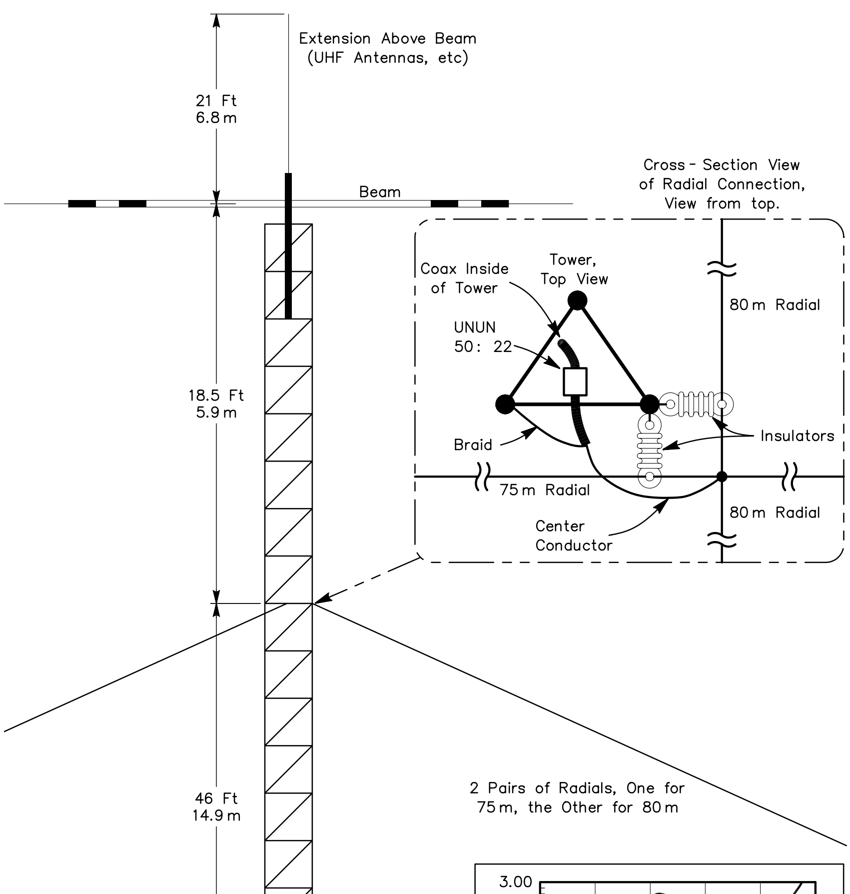

Fig 10—At A, details of the inverse-fed ground-plane at W4PK. The inset shows the feed-point details. Keep the two sets of radials as close to 90· as possible. Don't try to resonate the antenna by adjusting radial lengths. At B, SWR curve of the inverse-fed ground plane antenna with stagger-tuned radials. The objective is to have a low SWR for both the CW and SSB 80/75-meter DX windows.

Stagger-Tuned Radials

A variation of the use of stagger tuning has been applied by Sam Leslie, W4PK. See his article entitled "Broadbanding the Elevated, Inverse-Fed Ground Plane Antenna" in *The ARRL Antenna Compendium, Vol 6*. An existing tower with a large beam and VHF/UHF antennas on top is used as an 80-meter monopole. See **Fig 10A**. The feed point is part way up the tower, at a point where the metal above it makes up an electrical quarter wavelength. Four elevated quarter-wave radials are used as a ground plane. The radials droop away from the tower. They are joined together at the tower but not connected to the tower. The antenna is fed with 50-Ω coax and an impedance step-down autotransformer (50:22 Ω unun) located at the tower. The shield of the coax is carried through the transformer and connected to the tower. The coax "hot side" is connected to the junction of the radials. A Q-section made from two paralleled RG-59 coax cables ($Z_0 = 75/2 = 37.5\ \Omega$) may be used instead of the autotransformer.

The broadband match is achieved by cutting two opposing radials for one end of the desired match range (75 meters) and cutting the other two for the other end of the range (80 meters). See Fig 10B for the resultant SWR versus frequency characteristic.

FEED-LINE IMPEDANCE MISMATCHING

The simplest broadband matching network is a transformer at the junction of the transmission line and the antenna. See Frank Witt's article entitled "Match Bandwidth of Resonant Antenna Systems" in October 1991 *QST*. Observe that the impedance of the series RLC resonant antenna model of Fig 2 increases at frequencies away from resonance. The result is that more bandwidth is achieved when SWR at resonance is not exactly 1:1. For example, if the antenna is fed with a low-loss 50-Ω transmission line, the maximum SWR bandwidth is obtained if the effective $R_A = 50/1.25 = 40\ \Omega$. The improvement in BW_2, the 2:1 SWR bandwidth, over a perfect match at resonance is only 6%, however. This condition of $R_A = 40\ \Omega$ can be achieved by installing a transformer at the junction of the feed line and the antenna.

If the line loss is larger, you can gain more benefit in terms of bandwidth by deliberately mismatching at the antenna. If the matched-line loss is 2 dB, the improvement rises to 18%. To achieve this requires an $R_A = 28.2\ \Omega$. Again, the improvement over a perfect match at resonance is modest.

It is interesting to compare the 2:1 SWR bandwidth (at the transmitter end of the line) for the case of a lossless line versus the case of 2-dB matched-line loss. BW_2 for the lossy line case is 1.95 times that of the lossless line case, so substantial match bandwidth improvement occurs, but at the cost of considerable loss. The band edge loss for the lossy-line case is 3 dB!

This example is provided mostly as a reference for comparison with the more desirable broadbanding techniques below. It also explains why some installations with long feed lines show large match bandwidths.

PARALLEL-TUNED CIRCUITS AT THE ANTENNA TERMINALS

As mentioned previously, a resonant antenna has an equivalent circuit that may be represented as a series RLC circuit. We can cancel out some of the antenna reactance away from resonance with a parallel-tuned circuit connected across the antenna terminals. The impedance level of the parallel-tuned circuit must be low enough to be effective and must have the same resonant frequency as the antenna. You can use a parallel-tuned circuit made with lumped LC components or with transmission-line segments. A quarter-wave transmission line with a short at the far end, or a half-wave line with an open at the far end, each behave like a parallel-tuned LC circuit.

The Double Bazooka

The response of the controversial *double bazooka* antenna is shown in **Fig 11**. This antenna actually consists of a dipole with two quarter-wave coaxial resonator stubs connected in series. The stubs are connected across the antenna terminals.

Not much bandwidth enhancement is provided by this resonator connection because the impedance of the matching network is too high. The antenna offers a 2:1-SWR bandwidth frequency range that is only 1.14 times that of a simple dipole with the same feeder. And the bandwidth enhancement is partially due to the "fat" antenna wires composed mostly of the coax shield. No

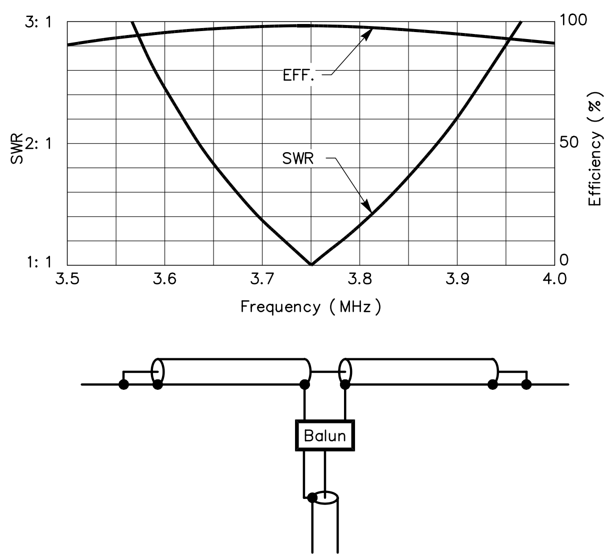

Fig 11—The double bazooka, sometimes called a coaxial dipole. The antenna is self-resonant at 3.75 MHz. The resonator stubs are 43.23-foot lengths of RG-58A coax.

improvement in antenna gain or pattern over a thin-wire dipole can be expected from this antenna.

The Crossed Double Bazooka

A modified version of the double-bazooka antenna is shown in **Fig 12**. In this case, the impedance of the matching network is reduced to one-fourth of the impedance of the standard double-bazooka network. The lower impedance provides more reactance correction, and hence increases the bandwidth frequency range noticeably, to 1.55 times that of a simple dipole. Notice, however, that the efficiency of the antenna drops to about 80% at the 2:1-SWR points. This amounts to a loss of approximately 1 dB. The broadbanding, in part, is caused by the losses in the coaxial resonator stubs (made of RG-58A coax), which have a remarkably low Q (only 20).

MATCHING NETWORK DESIGN

Optimum Matching

You can improve the match bandwidth by using a transformer or a parallel-tuned circuit at the antenna terminals. A combination of these two techniques can yield a very effective match bandwidth enhancement. Again, you can make the impedance transformation and implement the tuned circuit using lumped-circuit elements (transformers, inductors and capacitors) or with distributed-circuit elements (transmission lines).

In 1950, an article by R. M. Fano presented the theory of broadband matching of arbitrary impedances. This classic work was limited to lossless matching networks. Unfortunately, matching networks realized with transmission lines have enough loss to render the Fano results useful as only a starting point in the design process. A theory of optimum matching with lossy matching networks was presented in an article by Frank Witt entitled "Optimum Lossy Broadband Matching Networks for Resonant Antennas" that appeared in April 1990 *RF Design* and was summarized and extended in "Broadband Matching with the Transmission Line Resonator," in *The ARRL Antenna Compendium Vol 4*.

Using a matching network, the 2:1 SWR bandwidth can be increased by a factor of about 2.5. Instead of the familiar bowl-shaped SWR versus frequency characteristic seen in Fig 5, the addition of the matching network yields the W-shaped characteristic of **Fig 13**. The maximum SWR over the band is S_M. The band edges are F_L and F_H. Thus the bandwidth, BW, and the center frequency, F_0, are

$$BW = F_H - F_L \qquad \text{(Eq 10)}$$

$$F_0 = \sqrt{F_H \times F_L} \qquad \text{(Eq 11)}$$

Fig 14 shows the pertinent equivalent circuit for a typical broadband antenna system. The parallel LC matching network is defined by its resonant frequency, F_0 (which is the same as the resonant frequency of the antenna), its Q and its impedance level. The impedance level is the reactance of either the matching network inductance or capacitance at F_0. The optimum transmitter load resistance,

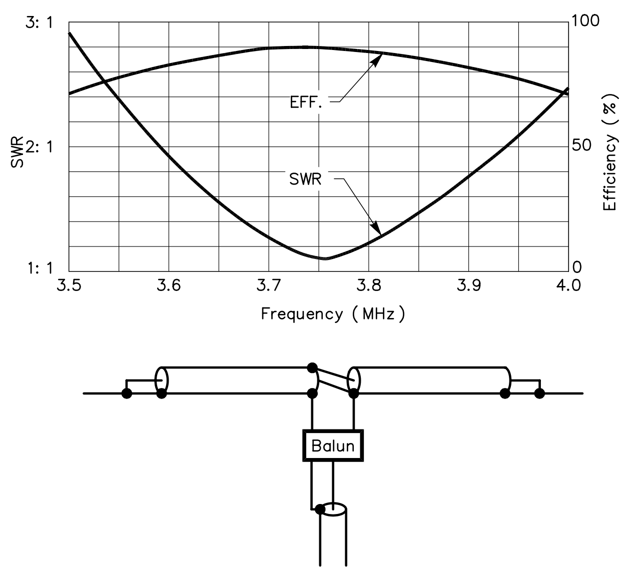

Fig 12—The crossed double bazooka yields bandwidth improvement by using two quarter-wave resonators, parallel connected, as a matching network. As with the double bazooka, the resonator stubs are 43.23-foot lengths of RG-58A coax.

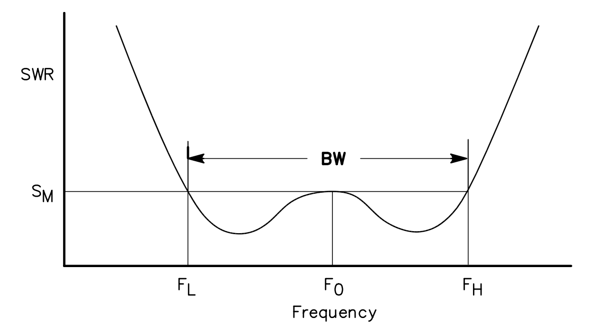

Fig 13—W-shaped SWR versus frequency characteristic that results when a transformer and parallel LC matching network are used.

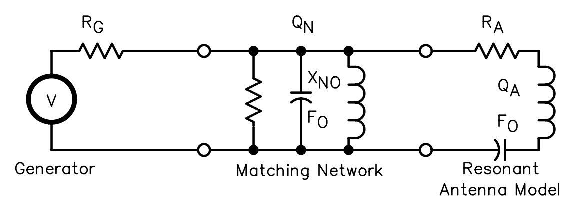

Fig 14—Assumed equivalent circuit for the broadband matching system. The antenna and the matching network have the same resonant frequency.

R_G, is achieved using an impedance transformer between the transmitter and the input to the parallel-tuned circuit. Thus for a transmitter whose optimum load resistance is 50 Ω, the required impedance transformer would have an impedance transformation ratio of 50:R_G.

The goal is to keep the SWR over the band as low as possible. There is a best or minimum value of S_M that may be achieved over the entire band. It depends on the desired match bandwidth and the Q's of the antenna and matching network and is given by the following formula:

$$S_{M\,min} = \frac{\sqrt{B_N^{\,2}+1} + \sqrt{B_N^{\,2}+1+\dfrac{2Q_A}{Q_N}\left(1+\dfrac{Q_A}{2Q_N}\right)}}{2\left(1+\dfrac{Q_A}{2Q_N}\right)} \quad \text{(Eq 12)}$$

where $B_N = \dfrac{BW}{F_0}$ and

Q_A = antenna Q

Q_N = matching network Q.

The reactance at resonance of the parallel tuned LC matching network is given by:

$$X_{N0} = \frac{R_A}{Q_A}\left[\left(1+\frac{Q_A}{2Q_N}\right)S_{M\,min}^{\,2} - \frac{Q_A}{2Q_N}\right] \quad \text{(Eq 13)}$$

where R_A = antenna resistance in Ω.

The transmitter load resistance is calculated from:

$$R_G = \frac{S_{M\,min}\,R_A\,Q_N\,X_{N0}}{R_A + Q_N\,X_{N0}} \quad \text{(Eq 14)}$$

The loss of the matching network is always greatest at the edges of the band. This loss, L_{MNE}, in dB, is given by:

$$L_{MNE} = 10\log\left[1+\frac{R_A}{Q_N\,X_{N0}}\left(B_N^{\,2}+1\right)\right] \quad \text{(Eq 15)}$$

Now you have all the information needed to design an optimal broadband matching network. An 80-meter dipole will serve as an example to illustrate the procedure. The following parameters are assumed:

F_L = 3.5 MHz
F_H = 4.0 MHz
R_A = 57.2 Ω
Q_A = 13
Q_N = 40.65

The values of R_A and Q_A are reasonable values for an 80-meter inverted V with an apex 60 feet above ground. Q_N = 40.65 is the calculated value of the Q of a resonator made from RG-213 or similar cable in the middle of the 80-meter band. The results are:

BW = 0.5 MHz	(from Eq 10)
F_0 = 3.742 MHz	(from Eq 11)
S_{Mmin} = 1.8	(from Eq 12)
X_{N0} = 15.9 Ω	(from Eq 13)
R_G = 94.8 Ω	(from Eq 14)
L_{MNE} = 1.32 dB	(from Eq 15)

Shown in **Fig 15** is the plot of SWR and matching-network loss versus frequency for this example. The reference resistance for the SWR calculation is R_G or 94.8 Ω. Since the desired load resistance for the transmitter is 50 Ω, we need some means to transform from 94.8 Ω to 50 Ω. The examples given in the rest of this chapter will show a variety of ways of achieving the necessary impedance transformation.

The meaning of X_{N0} = 15.9 Ω is that the inductor and the capacitor in the matching network would each have a reactance of 15.9 Ω at 3.742 MHz. Methods for obtaining a particular value of X_{N0} are shown in the fol-

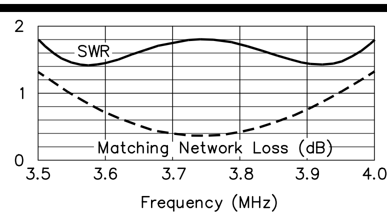

Fig 15—SWR and matching-network loss for example. Note that S_M is exactly 1.8 and the band-edge loss is exactly 1.32 dB, as calculated.

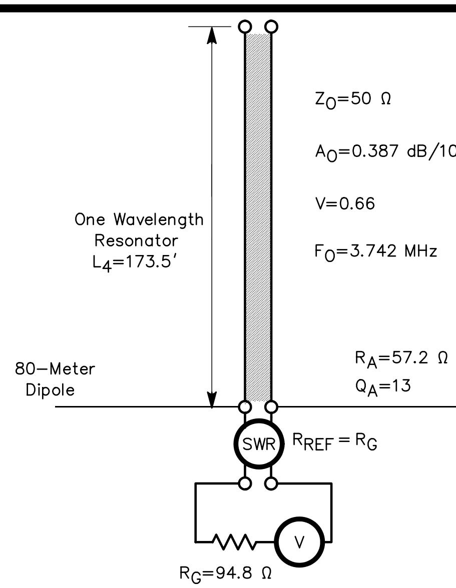

Fig 16—80-meter dipole example with a 1-λ resonator as the matching network.

lowing examples as well. The particular value of X_{N0} of this example may be obtained with a one-wavelength resonator (open-circuited at the far end) made from RG-213. The system described in this example is shown in **Fig 16**.

The broadband antenna system shown in Fig 16 has a desirable SWR characteristic, but the feed line to the transmitter is not yet present. Fortunately, the same cable segment that makes up the resonator may be used as the feed line. A property of a feed line whose length is a multiple of $\lambda/2$ (an even multiple of $\lambda/4$) is that its input impedance is a near replica of the impedance at the far end. This is exactly true for lossless lines at the frequencies where the $\lambda/2$ condition is true. Fortunately, practical lines have low-enough loss that the property mentioned above is true at the resonant frequency. Off resonance, the desired reactance cancellation we are looking for from the resonant line takes place. The very same design equations apply.

In **Fig 17**, the antenna shown in Fig 16 is moved to the far end of the 1λ resonator. The SWR and loss for this case are shown in **Fig 18**, which should be compared with Fig 15. The SWR is only slightly degraded, but the loss is about the same at the band edges. We have picked up a feed line that is 1λ long (173.5 feet) *and* broadband matching. You can make the SWR curve virtually the same as that of Fig 15 if the transmitter load resistance is increased from 94.8 Ω to 102 Ω. From a practical point of view, this degree of design refinement is unnecessary, but is instructive to know how the SWR characteristic may be controlled. You may apply the same approach when the resonator is an odd multiple of a wavelength, but in that case the transmitter or antenna must be connected $\lambda/4$ from the shorted end.

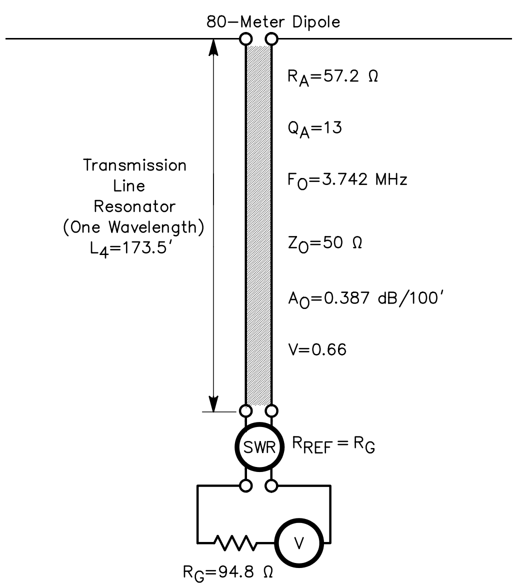

Fig 17—The antenna is at the far end of the 1-λ line. The line does double duty—broadband matching and signal transport.

Chebyshev Matching

Although it is not as efficient from the standpoint of bandwidth improvement and loss, Chebyshev matching is useful for some purposes. This arrangement yields a transmitter load impedance of 50 + *j* 0 Ω at two frequencies in the band. The matching network circuit is the same as the optimum matching case, but the parameters are different. In this case, we assume that the antenna parameters, F_0, R_A and Q_A, the network Q and the maximum SWR over the band are specified. The bandwidth, impedance level and generator resistance are given by:

$$BW = \frac{F_0}{Q_A Q_N} \left\{ \frac{(Q_A + Q_N)\left[(Q_A + 2Q_N)S_M - Q_A\right](S_M - 1)}{S_M} \right\}^{\frac{1}{2}} \qquad \text{(Eq 16)}$$

$$X_{N0} = \frac{R_A}{Q_A}\left[S_M + \frac{Q_A}{Q_N}(S_M - 1)\right] \qquad \text{(Eq 17)}$$

$$R_G = R_A\left(S_M - \frac{Q_A}{Q_A + Q_N}\right) \qquad \text{(Eq 18)}$$

The loss at the band edges is given by Eq 15. The frequencies where the transmitter load is 50 + *j* 0 Ω, F_L and F_H, are given by:

$$F_L = \sqrt{F_0^{\,2} + F_M^{\,2}} - F_M \qquad \text{(Eq 19)}$$

where

$$F_M = \frac{F_0}{2Q_A}\left[(S_M - 1)\left(1 + \frac{Q_A}{Q_N}\right)\right]^{\frac{1}{2}} \qquad \text{(Eq 20)}$$

$$F_H = \frac{F_0^{\,2}}{F_L} \qquad \text{(Eq 21)}$$

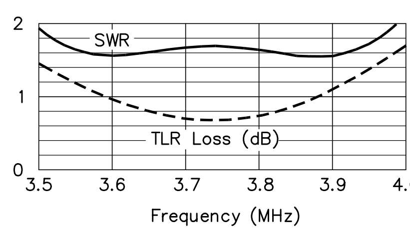

Fig 18—SWR and loss for a 1-λ 50-Ω feed line. The antenna system model is the one shown in Fig 17.

LC-MATCHING NETWORKS

Before the theory of optimal matching above was developed, Frank Witt, AI1H, described in his article "Broadband Dipoles—Some New Insights" in October 1986 *QST* the basic principle and some applications of LC matching networks. **Fig 19** shows an LC matching network that is a modified version of one originally proposed by Alan Bloom, N1AL. Note that the basic ingredients are present: the parallel-tuned circuit for reactance cancellation and a transformer.

The network also functions as a voltage balun by connecting the shield of the coaxial feed line to the center tap of the inductor. The capacitor is connected across the entire coil to obtain practical element values. The SWR response and efficiency offered by a network of lumped components is shown in **Fig 20**. The 2:1-SWR bandwidth with 50-Ω line is 460 kHz. The design is a Chebyshev match, where SWR = 1:1 is achieved at two frequencies. For the same configuration, more match bandwidth and less loss could have been realized if the optimal-match theory had been applied.

The efficiency at the band edges for the antenna system shown in Fig 20 is 90% (Loss = 0.5 dB). This low loss is due to the high Q of the LC matching network ($Q_N \approx 200$). The very low-impedance level required ($X_{N0} = 12.4\ \Omega$) cannot be easily realized with practical inductor-capacitor values. It is for this reason that the coil is tapped and serves as an autotransformer with multiple taps. The required impedance transformation ratio of 1:2.8 is easily achieved with this arrangement.

80-Meter DXer's Delight

A version of an antenna system with an LC broadbanding network is dubbed the *80-meter DXer's Delight*. This antenna has SWR minima near 3.5 and 3.8 MHz. A single antenna permits operation with a transmitter load impedance of about 50 Ω in the DX portions of the band, both CW and phone. The efficiency and SWR characteristic are shown in **Fig 21**.

Fig 22 shows the method of construction. The selection of a capacitor for this application must be made carefully, especially if high power is to be used. For the capacitor described in the caption of Fig 22, the allowable peak power (limited by the breakdown voltage) is 2450 W. However, the allowable average power (limited by the RF current rating of 4 A) is only 88 W! These limits apply at the 1.8:1 SWR points.

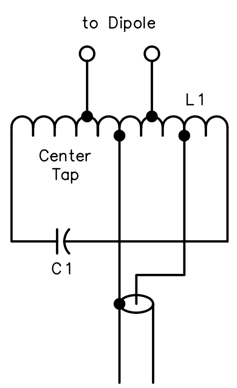

Fig 19—A practical LC matching network that provides reactance compensation, impedance transformation and balun action.

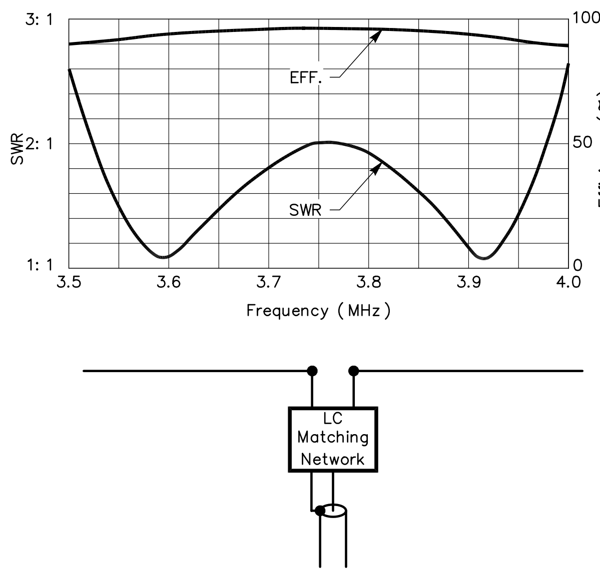

Fig 20—Efficient broadband matching with a lumped element LC network. The antenna in this example has a resistance of 72 Ω and a Q of 12. The matching network has an impedance level of 12.4 Ω, and a Q of 200. The feed line is 50-Ω coax, and the matching network provides an impedance step-up ratio of 2.8:1.

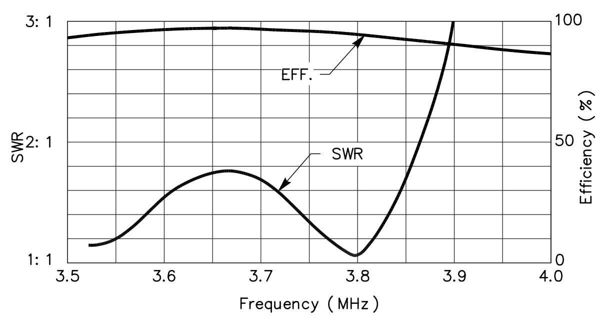

Fig 21—The 80-meter DXer's Delight permits operation with a 50-Ω transmitter load in the DX portions of the band, both CW and phone. This network was designed to provide a broadband match for an inverted V with F_0 = 3.67 MHz, an antenna resistance of 60 Ω and a Q of 13. The matching network has an impedance level of 9 Ω, and a Q of 220. The feed line is 50-Ω coax, and the matching network provides an impedance step-up ratio of 2.0:1.

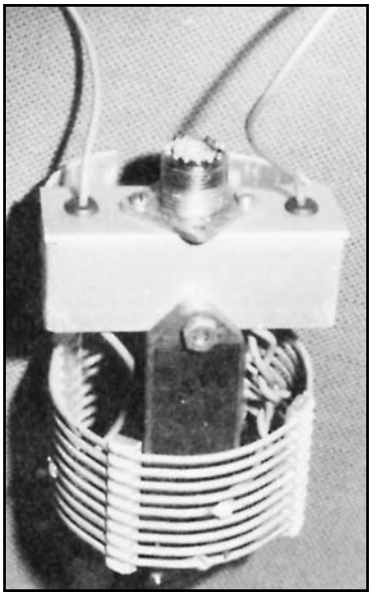

Fig 22—A method of constructing the LC broadband matching network. Components must be chosen for high Q and must have adequate voltage and current ratings. The network is designed for use at the antenna feed point, and should be housed in a weather-resistant package. The component values are as follows:
C1—400 pF transmitting mica rated at 3000 V, 4 A (RF)
L1—4.7 µH, 8 1/2 turns of B&W coil stock, type 3029 (6 turns per in., 2 1/2-in. dia, #12 wire).
The LC circuit is brought to midband resonance by adjusting an end tap on the inductor. The primary and secondary portions of the coil have 1 3/4 and 2 1/2 turns, respectively.

PROPERTIES OF TRANSMISSION-LINE RESONATORS

One way around the power limitations of LC broadband matching networks is to use a transmission line as the resonator. Transmission line resonators, TLRs, typically have higher power-handling capability because the losses, though higher, are distributed over the length of the line instead of being concentrated in the lumped-LC components. Transmission-line resonators must be multiples of a quarter wavelength. For a parallel-tuned circuit, the odd-multiple λ/4 lines must be shorted at the far end and the even-multiple λ/4 lines must be open circuited.

The length of the matching network resonator is n electrical quarter wavelengths. The impedance level, Q, and line length are given by:

$$X_{N0} = \frac{4 Z_0}{n \pi}$$

or, solving for Z_0,

$$Z_0 = \frac{n \pi X_{N0}}{4} \qquad \text{(Eq 22)}$$

$$Q_N = \frac{2.774 F_0}{A_0 \; V} \qquad \text{(Eq 23)}$$

$$L_N = \frac{245.9 \; n \; V}{F_0} \qquad \text{(Eq 24)}$$

where

Z_0 = line characteristic impedance in Ω
A_0 = matched line loss in dB per 100 feet
V = velocity factor, and
L_N = line length in feet.

These equations may be used to find the properties of the transmission line resonator in the systems shown in Figs 16 and 17. The line is RG-213 with the following properties:

Z_0 = 50 Ω
V = 0.66
A_0 = 0.4 dB/100 feet at 4 MHz

We need the matched loss at 3.742 MHz. Since the line loss is approximately proportional to $\sqrt{\text{frequency}}$,

$$A_{0F2} = A_{0F1} \sqrt{\frac{F_2}{F_1}} \qquad \text{(Eq 25)}$$

where A_{0F1} and A_{0F2} are the matched line losses at frequencies F_1 and F_2, respectively. Hence,

A_0 = 0.387 dB/100 feet at 3.742 MHz
Q_N = 40.65 (from Eq 23)

This is the Q of any resonator made from RG-213 for a frequency of 3.742 MHz. It does not matter how many quarter wavelengths (n) make up the cable segment. Earlier we calculated that the required impedance level, X_{N0}, is 15.9 Ω. Therefore, the required impedance level is:

$$Z_0 = \frac{n \pi \times 15.9}{4} = 12.5 \, n \, \Omega \qquad \text{(from Eq 22)}$$

This means that you could use a quarter-wavelength segment with a short at the far end if Z_0 = 12.5 Ω, or a half-wavelength segment with a open at the far end if Z_0 = 25 Ω. Our example used an open-circuited full-wavelength resonator (n = 4) with a Z_0 of 50 Ω. This example was crafted to make this happen (so that it fits RG-213 properties), but it illustrates that some juggling is required to obtain the right resonator properties when it is made from a transmission line, since we are stuck with the cable's characteristic impedance.

The TLR Transformer

A way around the limitation created by available characteristic impedance values is to use the *transmission-line resonator transformer*. The basic idea is to make the connection to the TLR at an intermediate point along the line instead of at the end of the line. It is analogous to using taps on a parallel-tuned LC resonant circuit to achieve transformer action (See Fig 19). The nature of the transformer action is seen in **Fig 23**, where the ends of the transformer are designated 1-1 and 2-2. The impedance ratio of the transformer, N_Z, is approximated by

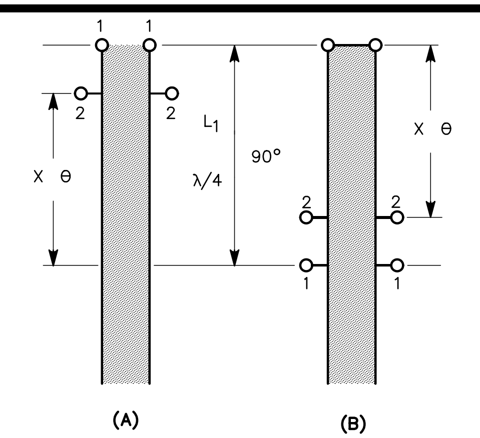

Fig 23—Transformer action in the TLR. The definition of transformer terminals depends on whether the TLR end is open- or short-circuited. θ is the distance between the minimum of the voltage standing wave (at resonance) and the connection point, expressed as an electrical angle. The distance x is that same distance expressed in feet.

$$N_Z = \sin^2 \theta \qquad \text{(Eq 26)}$$

The distance in feet, x, is given by

$$x = \frac{\theta}{90} \times \frac{\lambda}{4} = \frac{\theta}{90} \times \frac{245.9\ V}{F_0} \qquad \text{(Eq 27)}$$

This approach may be used for the connection of the TLR to the transmitter or to the antenna or both. An example will help make the power of the TLR transformer clear. We want to broadband match an 80-meter dipole (where F_0 is 3.742 MHz, R_A is 65 Ω and Q_A is 13). The feed line is RG-213 50-Ω cable, and the total distance between the antenna and transmitter is less than 100 feet.

Since we know that Z_0 = 50 Ω, we will design a prototype system with a ¾ λ TLR (n=3). Since n is odd, the TLR must be shorted at one end, and we will place the short at the transmitter end of the line. The intermediate step of using a prototype system will enable us to design a system with the final TLR, but without transformer action. Then we'll calculate the proper connection points along the TLR. **Fig 24** shows the prototype system and the system with built-in transformers.

Note that the prototype antenna is connected to the open end of the TLR. The prototype generator (transmitter) is connected at a multiple of λ/2 from the antenna. Here, at resonance, an approximate replica of the antenna will appear. We will use primed variables to indicate prototype antenna system elements. The prototype antenna differs from the actual antenna in impedance level, so only R_A will change. F_0 and Q_A remain the same.

Starting with the ¾ λ TLR, we have:

X_{N0} = 21.2 Ω (from Eq 22)

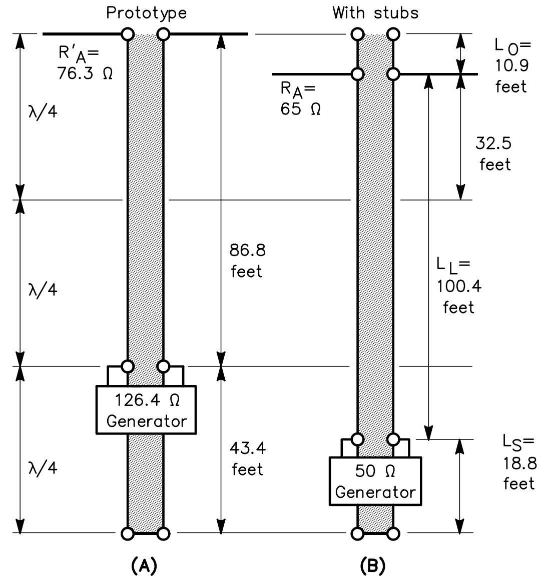

Fig 24—An optimized antenna system with a ¾ λ TLR made from RG-213 coaxial cable. The prototype system at A is a convenient intermediate step in the design process. At B is the configuration with TLR transformers. It has an open stub, L_O, a link, L_L (which serves as the feed line), and a shorted stub, L_S.

The prototype antenna resistance at resonance, R'_A, is calculated by rearranging Eq 13.

$$R'_G = \frac{X_{N0}\, Q_A}{\left(1 + \frac{Q_A}{2Q_N}\right) S_{M\,min}{}^2 - \frac{Q_A}{2\,Q_N}} \qquad \text{(Eq 28)}$$

Earlier we calculated that Q_N = 40.65 and S_{Mmin} = 1.8. Now we find

R'_A = 76.3 Ω (from Eq 28)
R'_G = 126.4 Ω (from Eq 14)

The electrical angles (in degrees) and connection locations (in feet) are given by

$$\theta_A = \sin^{-1}\sqrt{\frac{R_A}{R'_A}} \qquad \text{(Eq 29)}$$

$$\theta_G = \sin^{-1}\sqrt{\frac{R_G}{R'_G}} \qquad \text{(Eq 30)}$$

$$X_A = \frac{\theta_A}{90} L_1 \qquad \text{(Eq 31)}$$

$$X_G = \frac{\theta_G}{90} L_1 \qquad \text{(Eq 32)}$$

where

$$L_1 = \frac{\lambda}{4} = \frac{245.9\ V}{F_0}$$

The lengths of the stubs and link of Fig 24 are given by

$L_O = L_1 = x_A$ (Eq 33)

$L_S = x_G$ (Eq 34)

$L_L = 3L_1 - L_O - L_S$ (Eq 35)

Since L_1 = 43.4 feet,

$\theta_A = 67.4°$ (from Eq 29)

$\theta_G = 39.0°$ (from Eq 30)

L_O = 10.9 feet (from Eq 33)

L_S = 18.8 feet (from Eq 34)

L_L = 100.4 feet (from Eq 35)

These dimensions are summarized in Fig 24. If required, additional 50-Ω cable may be added between the transmitter and the junction of the shorted stub and the link. A remarkable aspect of the TLR transformer, as demonstrated in this example, is that it is a transformer with multiple taps distributed over a great distance, in this case over 100 feet.

The calculated SWR and loss of the antenna system with the TLR transformer are shown in **Fig 25A**. We mentioned earlier that the stub calculations are an approximation. Fig 25 bears this out, but the changes necessary to achieve the optimized results are relatively small. By changing the lengths and dipole resonant frequency slightly, the target optimized SWR characteristic of Fig 25B results. See **Table 1** for a summary of the changes required.

To demonstrate the significant improvement in match bandwidth that the TLR transformer provides, the SWR and loss of the same dipole fed with 100.4 feet of RG-213 cable is shown in Fig 25C. The loss added by the two stubs used to obtain match bandwidth enhancement is negligible (0.2 to 0.5 dB).

A TLR transformer may be used by itself to achieve a low-loss narrow-band impedance transformation, functioning like a λ/4 Q-section. It is different in nature from a Q-section, in that it is a true impedance transformer, while a Q section is an *impedance inverter*. The TLR transformer may be designed into the antenna system and is a useful tool when a narrow-band impedance transformation is needed.

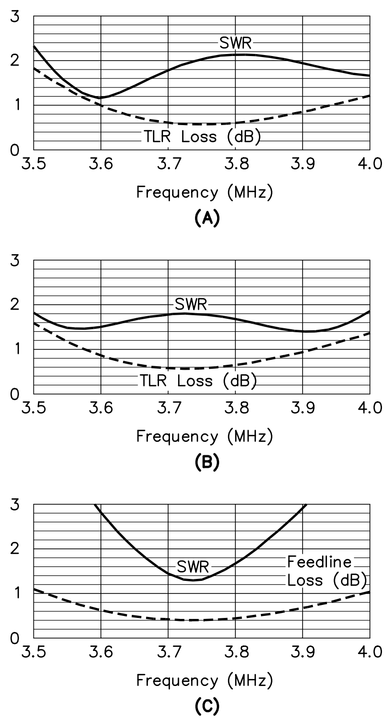

Fig 25—The SWR and loss of the TLR transformer antenna system. At A, the results are shown with the calculated line lengths. At B, the improvement after the lengths were optimized is shown in Table 1. At C, for comparison, SWR and loss for the same setup as at A, but with the stubs removed.

Table 1
Calculated and Optimized Parameters Using TLR Transformers

	Calculated	*Optimized*
Dipole resonant frequency (F_0)	3.742 MHz	3.710 MHz
Open stub (L_O)	10.9 feet	13.1 feet
Shorted stub (L_S)	18.8 feet	20.1 feet
Link (L_L)	100.4 feet	101.0 feet

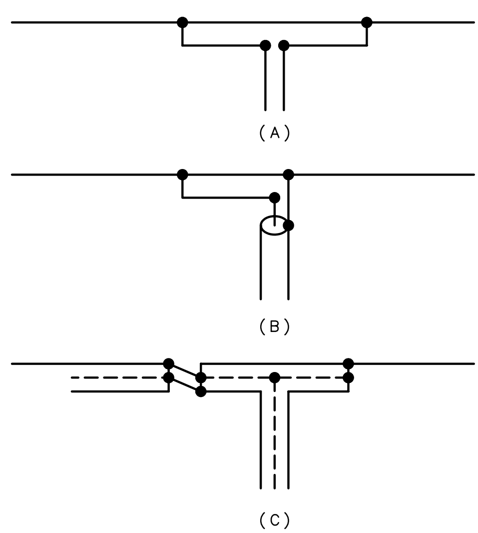

Fig 26—Dipole matching methods. At A, the T match; at B, the gamma match; at C, the coaxial resonator match.

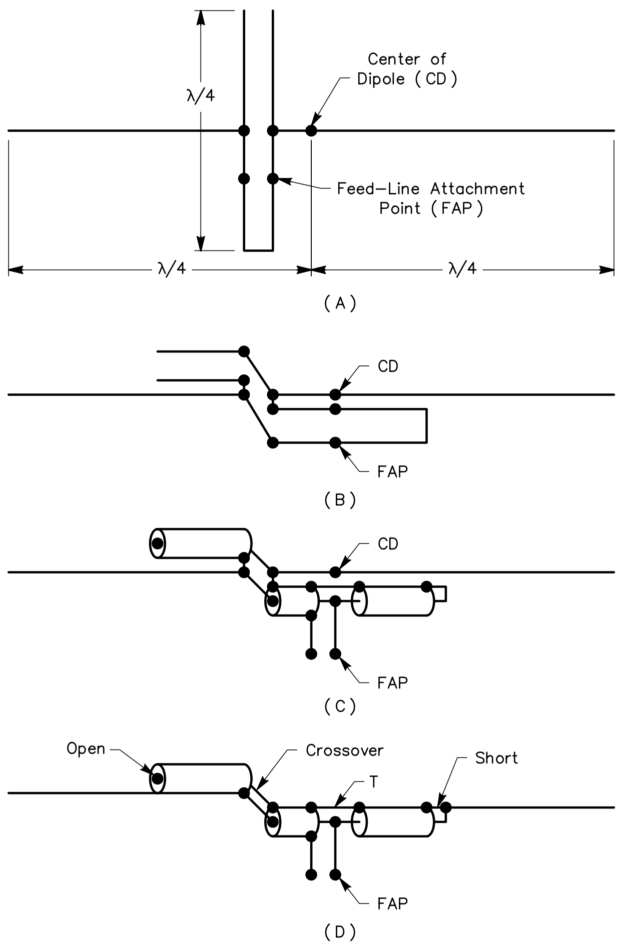

Fig 27—Evolution of the coaxial-resonator-match broadband dipole. At A, a TLR transformer is used to match the feed line to the off-center-fed dipole. The match and dipole are made collinear at B. At C, the balanced transmission-line TLR transformer of A and B is replaced by a coaxial version. Because the shield of the coax can serve as a part of the dipole radiator, the wire adjacent to the coax match may be eliminated, D.

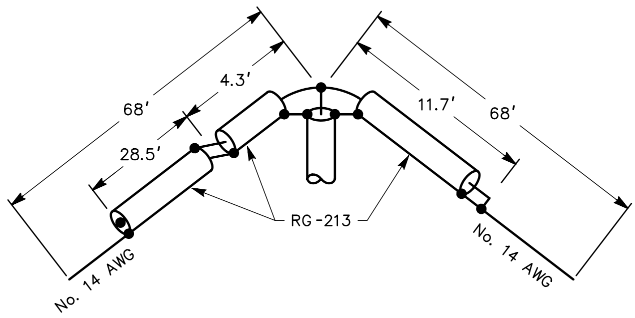

Fig 28—Dimensions for the 80-Meter MHz DX Special, an antenna optimized for the phone and CW DX portions of the 3.5-MHz band. The coax segment lengths total to one quarter-wavelength. The overall length is the same as that of a conventional inverted V dipole.

TRANSMISSION-LINE RESONATORS AS PART OF THE ANTENNA

The Coaxial-Resonator Match

This material is condensed from articles that appeared in April 1989 *QST* and *The ARRL Antenna Compendium, Vol 2*. The coaxial-resonator match, a concept based on the TLR transformer, performs the same function as the T match and the gamma match, that is, matching a transmission line to a resonant dipole. These familiar matching devices as well as the coaxial resonator match are shown in **Fig 26**. The coaxial resonator match has some similarity to the gamma match in that it allows connection of the shield of the coaxial feed line to the center of the dipole, and it feeds the dipole off center. The coaxial resonator match has a further advantage: It can be used to broadband the antenna system while providing an impedance match.

The coaxial resonator match is a resonant transformer made from a quarter-wave long piece of coaxial cable. **Fig 27** shows the evolution of the coaxial-resonator match. Now it becomes clear why coaxial cable is used for the quarter-wave TLR transformer—interaction between the dipole and the matching network is minimized. The effective dipole feed point is located at the crossover, which is an off-center feed point. In effect, the match is physically located "inside" the dipole. Currents flowing on the inside of the shield of the coax are associated with the resonator; currents flowing on the outside of the shield of the coax are the usual dipole currents. Skin effect provides a degree of isolation and allows the coax to perform its dual function. The wire extensions at each end make up the remainder of the dipole, making the overall length equal to one half-wavelength.

A useful feature of an antenna using the coaxial-resonator match is that the entire antenna is at the same dc potential as the feed-line potential, thereby avoiding charge buildup on the antenna. Hence, noise and the potential of lightning damage are reduced.

Fig 28 shows the detailed dimensions of an 80-meter inverted-V dipole using a coaxial-resonator match. This design provides a load SWR better than 1.6:1 from 3.5 to 3.825 MHz and has been named the *80-Meter DX Special*. The coax is an electrical quarter wavelength, has a short at one end, an open at the other end, a strategically placed crossover, and is fed at a tee junction. (The crossover is made by connecting the shield of one coax segment to the center conductor of the adjacent segment and by connecting the remaining center conductor and shield in a similar way.) The antenna is constructed as an inverted-V dipole with a 110° included angle and an apex at 60 feet. The measured SWR versus frequency is shown in **Fig 29**. Also in Fig 29

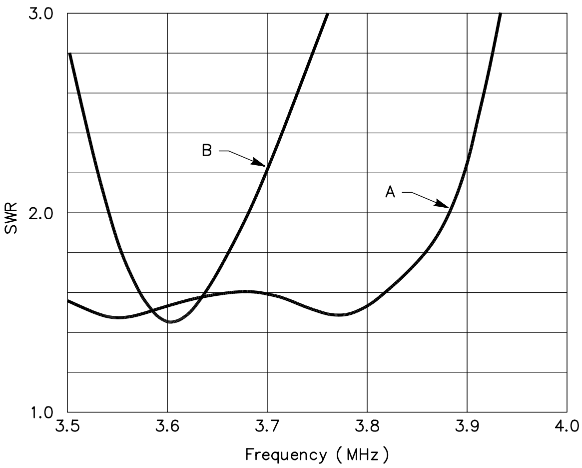

Fig 29—Measured SWR performance of the 80-Meter DX Special, curve A. Note the substantial broadbanding relative to a conventional uncompensated dipole, curve B.

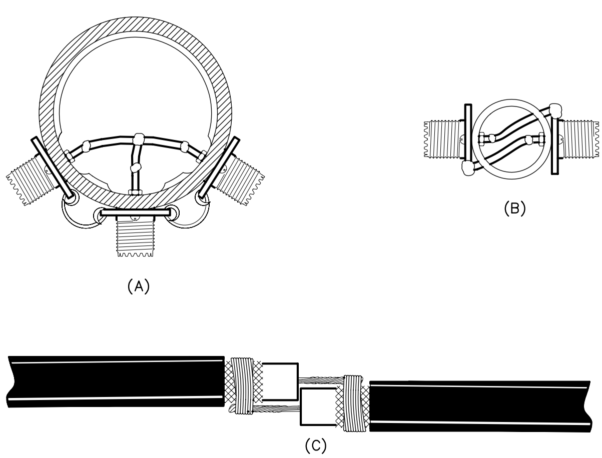

Fig 30—T and crossover construction. At A, a 2-inch PVC pipe coupling can be used for the T, and at B, a 1-inch coupling for the crossover. These sizes are the nominal inside diameters of the PVC pipe that is normally used with the couplings. The T could be made from standard UHF hardware (an M-358 T and a PL-258 coupler). An alternative construction for the crossover is shown at C, where a direct solder connection is made.

is the SWR characteristic for an uncompensated inverted-V dipole made from the same materials and positioned exactly like the broadband version.

The 80-Meter DX Special is made from RG-213 coaxial cable and #14 AWG wire, and is fed with 50-Ω coax. The coax forming the antenna should be cut so that the stub lengths of Fig 28 are within $^1/_2$ inch of the specified values. PVC plastic-pipe couplings and SO-239 UHF chassis connectors can be used to make the T and crossover connections, as shown in **Fig 30A** and 30B. Alternatively, a standard UHF T connector and coupler can be used for the T, and the crossover may be a soldered connection (Fig 30C). RG-213 is used because of its ready availability, physical strength, power handling capability, and moderate loss.

Cut the wire ends of the dipole about three feet longer than the lengths given in Fig 28. If there is a tilt in the SWR-frequency curve when the antenna is first built, it may be flattened to look like the shape given in Fig 29 by increasing or decreasing the wire length. Each end should be lengthened or shortened by the same amount.

A word of caution: If the coaxial cable chosen is not RG-213 or equivalent, the dimensions will have to be modified. The following cable types have about the same characteristic impedance, loss and velocity factor as RG-213 and could be substituted: RG-8, RG-8A, RG-10, RG-10A and RG-215. If the Q of the dipole is particularly high or the radiation resistance is unusually low because of different ground characteristics, antenna height, surrounding objects and so on, then different segment lengths will be required.

What is the performance of this broadband antenna relative to that of a conventional inverted-V dipole? Aside from the slight loss (about 0.75 dB at band edges, less elsewhere) because of the non-ideal matching network, the broadband version will behave essentially the same as a dipole cut for the frequency of interest. That is, the radiation patterns for the two cases will be virtually the same.

The Snyder Dipole

A commercially manufactured antenna utilizing the principles described in the Matching Network Design section is the Snyder dipole. Patented by Richard D. Snyder in late 1984 (see Bibliography), it immediately received much public attention through an article that Snyder published. Snyder's claimed performance for the antenna is a 2:1 SWR bandwidth of 20% with high efficiency.

The configuration of the Snyder antenna is like that of the crossed-double bazooka of Fig 12, with 25-Ω line used for the cross-connected resonators. In addition, the antenna is fed through a 2:1 balun, and exhibits a W-shaped SWR characteristic like that of Fig 13. The SWR at band center, based on information in the patent document, is 1.7 to 1. There is some controversy in professional circles regarding the claims for the Snyder antenna.

The Improved Crossed-Double Bazooka

The following has been condensed from an article by Reed Fisher, W2CQH, that appeared in *The ARRL Antenna Compendium Volume 2*. The antenna is shown in **Fig 31**. Note that the half-wave flat-top is constructed of sections of RG-58 coaxial cable. These sections of coaxial cable serve as quarter-wave shunt stubs that are essentially connected *in parallel* at the feed point in the crossed double-bazooka fashion. At an electrical quarter

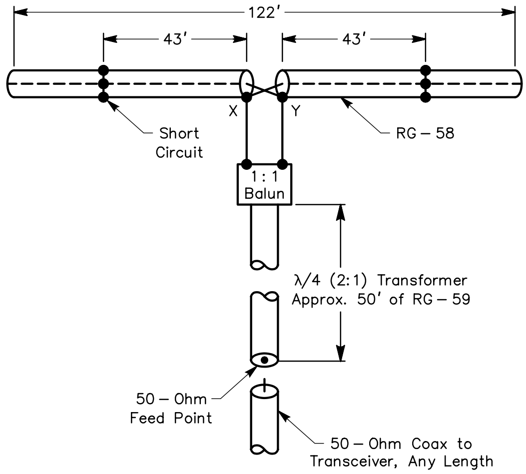

Fig 31—Details of the W2CQH broadband-matched 80-meter dipole.

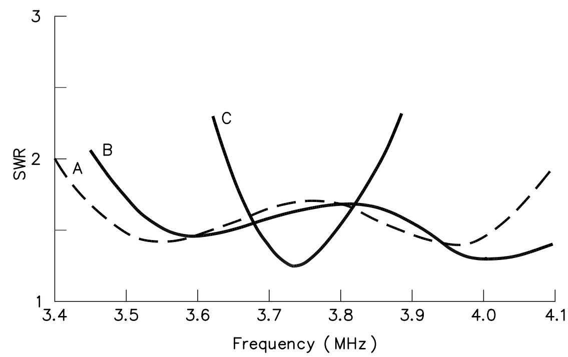

Fig 32—SWR curves for the improved double bazooka. Curve A, the theoretical curve with 50-Ω stubs and a λ/4, 75-Ω matching transformer. Curve B, measured response of the same antenna, built with RG-58 stubs and an RG- 59 transformer. Curve C, measurements from a dipole without broadbanding. Measurements were made at W2CQH with the dipole horizontal at 30 feet.

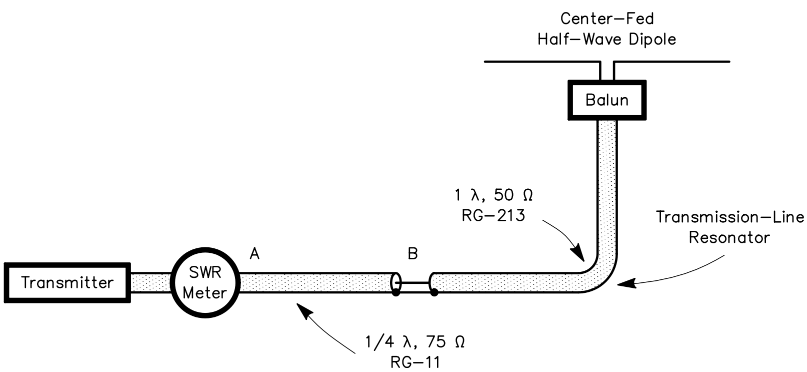

Fig 33—This simple broadband antenna system resembles a conventional 80-meter dipole except for the λ/4-wavelength 75-Ω segment. The lengths of the Q-section, TLR (including balun) and dipole are 43.3 feet, 170.5 feet, and 122.7 feet, respectively.

wavelength (43 feet) from each side of the feed point X-Y, the center conductor is shorted to the braid of the coaxial cable.

The parallel stubs provide reactance compensation. The necessary impedance transformation at the antenna feed point is provided by the quarter-wave Q-section constructed of a 50-foot section of 75-Ω coaxial cable (RG-59). A W2DU 1:1 current balun is used at the feed point. See **Fig 32** for the SWR versus frequency for this antenna with and without the broadband matching network.

This antenna is essentially that of the crossed double bazooka antenna shown in Fig 12, with the addition of the 75-Ω Q-section. The improvement in match bandwidth is substantial. The antenna at W2CQH is straight and nearly horizontal with an average height of about 30 feet.

TRANSMISSION-LINE RESONATORS AS PART OF THE FEED LINE

A Simple Broadband Dipole for 80 Meters

This system was described in an article in September 1993 *QST*. It has the advantage that the radiator is the same as that of a conventional half-wave wire dipole. Thus the antenna is light weight, easy to support and has small wind and ice loading. The broadband matching network is integrated into the feed line.

The system is shown in **Fig 33**. It is a variation of the example shown in the Matching Network Design section of this chapter as shown in Fig 17. The TLR is a 1λ length of RG-213 50-Ω coax and the impedance transformation is accomplished with a 75-Ω Q-section

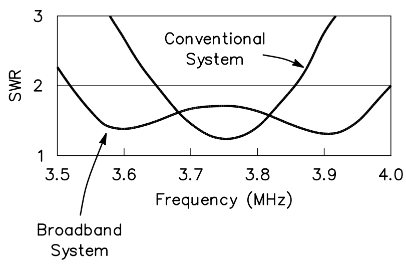

Fig 34—Measured SWR versus frequency for the broadband and conventional antenna systems.

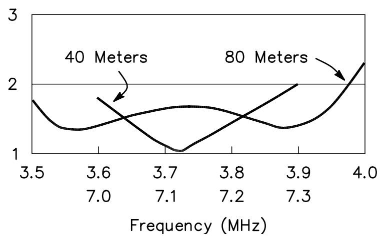

Fig 35—Measured SWR for the 80- and 40-meter multiband antenna system.

made of RG-11. The antenna is an inverted-V with a 140° included angle and an apex height of 60 feet. The wire size is #14, but is not critical. The balun is a W2DU 1:1 current type. It easily handles the legal power limit over the entire 80-meter band. The measured SWR is shown in **Fig 34**.

A unique advantage of this antenna is that by paralleling a 40-meter dipole at the feed point and sharing the feed line, operation on both 80 and 40 meters is possible. The reason for this is that the lengths of the Q-section and the TLR are multiples of a half-wavelength on 40 meters. To minimize the interaction, the two dipoles should be spaced from each other away from the feed point. First tune the 80-meter antenna and then the 40-meter one. **Fig 35** shows the result of adding a 40-meter dipole to the system shown in Fig 33. Each 40-meter dipole leg is 34.4 feet long. Note that the SWR on 80 meters changes very little compared to Fig 34. No change was made to the 80-meter dipole or to the transmission line.

80-Meter Dipole with the TLR Transformer

The total length of the feed line in the system just described above in Fig 33 is quite long, about 214 feet. For installations with shorter runs, the TLR transformer concept may be employed to advantage. An example provided in "Optimizing the 80-Meter Dipole" in *The ARRL Antenna Compendium, Vol 4* is shown in **Fig 36**. Here the TLR is $^{3}/_{4}$ λ long overall and is made from RG-213 50-Ω cable. The design applies for an 80-meter horizontal dipole, 80 feet above average ground, with an

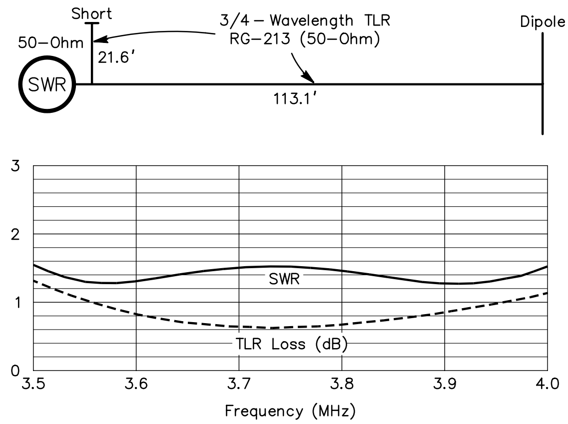

Fig 36—Broadbanding with the TLR transformer. The dimensions shown apply for an 80-meter horizontal dipole 80 feet above average ground.

assumed antenna F_0, R_A, and Q_A of 3.72 MHz, 92 Ω and 9, respectively. (See Figs 3 and 4). The lengths must be revised for other dipole parameters.

For this design, the feed line is 113.1 feet long and the shorted stub at the transmitter end is 21.6 feet long. You can add any length of 50-Ω cable between the transmitter and the junction of the shorted stub and the feed line. The shorted stub gives a dc path to ground for both sides of the dipole, preventing charge buildup and some lightning protection. For multiband operation with an antenna tuner, the stub should be removed.

The general application of the TLR transformer concept has a tap at both the transmitter end and a tap at the antenna end. (See Fig 24.) In this case, the concept was applied only at the transmitter end, since very little impedance transformation was required at the antenna end. The location of the TLR transformer is not obvious. Consider a $^{3}/_{4}$ λ TLR with a short at one end, and an open at the other end where the antenna is connected. The generator for no impedance transformation would be located λ/4 (43.6 feet @ 3.72 MHz) from the short. By connecting the transmitter 21.6 feet from the short, the required impedance transformation for a transmitter optimized for 50-Ω loads is achieved.

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