Broadband Dipoles—Some New Insights

The search for a broadband 80-meter dipole has continued. Efficiency, often ignored, plays an important role.

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Antennas that provide a good impedance match over a wide frequency range have been a topic of interest to hams for many years. Interest has been highlighted by several recent articles and a recent patent.[1-5] The advantages of a broadband antenna are obvious: fewer adjustments during tune-up, and, for some of the new broadband transceivers and amplifiers, no tune-up at all; after setting the band and frequency, one is "in business."

The topic of broadbanded antennas is fraught with misconceptions and impressive claims. In an attempt to gain further understanding, I wrote a computer program that analyzes the performance of the half-wave dipole with various kinds of bandwidth-broadening schemes. The program, written in BASIC for the Commodore™ 64 computer, is user friendly and well documented. (It may be translated to run on other computers.) Appendix 1 contains a description of the program and the assumptions and equations used in the analysis. Appendix 2 shows how one can design some of the broadband antennas described in this article.

The Half-Wave Dipole

In order to provide a foundation of comparison for some new approaches, first let us briefly review the performance of some previously reported bandwidth-broadening schemes.[6] The reference antenna used for comparison is the uncompensated half-wave dipole. The SWR versus frequency plot for a typical 80-meter horizontal wire dipole is shown in Fig 1. The data shown with the drawing gives the parameters of the dipole.

Most modern amplifiers are designed to match an antenna system with an SWR of better than 2:1. Therefore, in this article, the frequency band over which the SWR is better than 2:1 will be used as the comparison bandwidth. For the typical, uncompensated dipole of Fig 1, the bandwidth is 220 kHz.

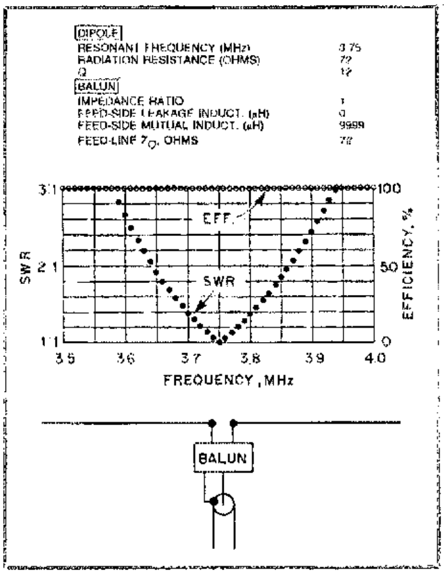

Fig 1—The 80-meter uncompensated half-wave dipole.

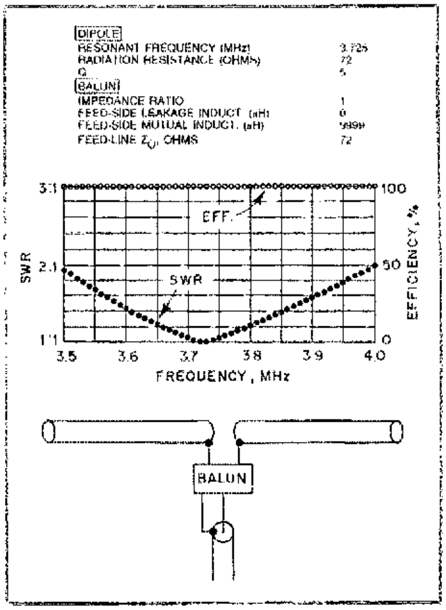

Fig 2—The cage dipole gives very efficient broadbanding, but poses practical problems on 80 and 160 meters.

Cage Dipole

Fig 2 shows what can be done by lowering the Q of a dipole. In practice, this is accomplished by fattening the dipole conductor, as in the case of the cage dipole. This technique is usually not a practical approach for 80 and 160 meters, where bandwidth widening is desired, because of the required cage dimensions. For example, in the case of an 80-meter dipole, it would be necessary to make a cage 3 feet in diameter in order to achieve the characteristics of Fig 2.[7] The bow-tie and fan dipole make use of the same Q lowering principle to obtain increased bandwidth.

Importance of Efficiency

When one augments a dipole with some matching scheme to acquire increased bandwidth, a factor frequently overlooked is the efficiency of the antenna system. When the SWR at the antenna end of the transmission line is less than 2:1, the transmission-line losses are virtually the same as those from the length of a matched line. Thus, for the part of the band over which the SWR is less than 2:1, one need only consider losses in the matching network when computing efficiency. Hence, the efficiency definition used here is:

$$\text{Efficiency (\%)} = \frac{100 \times \text{Power radiated by dipole}}{\text{Total power delivered to the antenna plus matching network}}$$

In what follows, not only will the SWR versus frequency be calculated, but so will the efficiency versus frequency.

Efficiency is related to resistive or ohmic losses in the matching network. The lower the losses, the higher the efficiency. However, ohmic losses in the matching network will broaden the response of a dipole system beyond that possible with a lossless or ideal matching network. Users must

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[1] Notes appear on p 37.

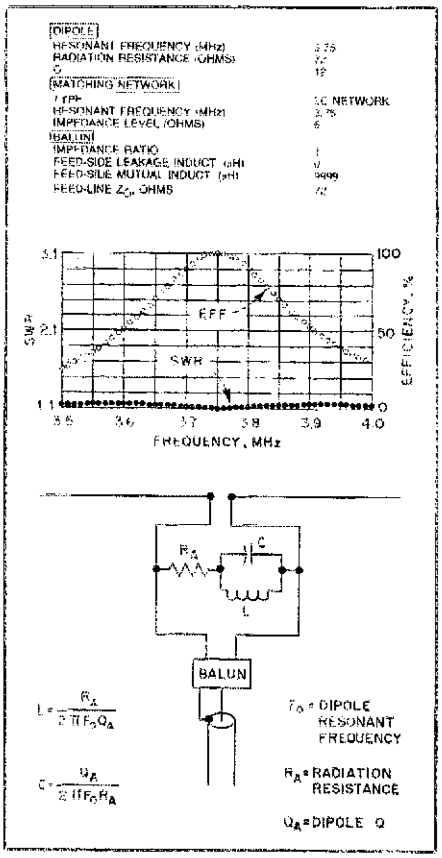


Fig 3—Matching the dipole with a complementary RLC network greatly improves the SWR characteristics, nearly 1:1 across the 80/75-m band. But the relative loss at the band edges is greater than 5 dB.

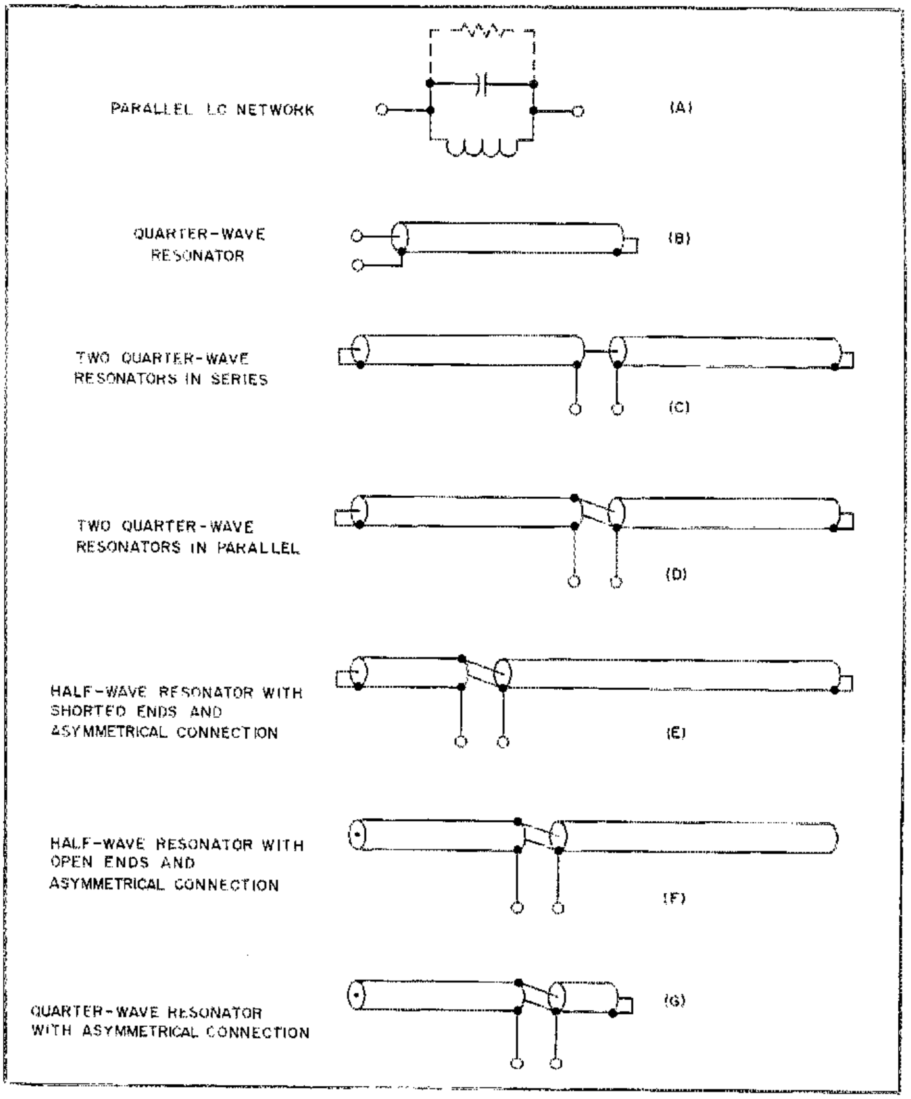


Fig 4—Types of resonators for broadbanding resonant antennas.

decide whether they are willing to accept the lower efficiency in trade for the increased bandwidth.

An extreme degree of bandwidth broadening is illustrated in Fig 3. The broadening is accomplished by adding resistive losses. One may resort to some network theory and derive the RLC (resistor, inductor, capacitor) matching network shown there. The network is called the complement of the antenna impedance. Note that the SWR is virtually 1:1 over the entire band, but the efficiency falls off dramatically away from resonance. Another way of interpreting the band-edge efficiency of 25 or 30% is that the antenna has about 5 dB of loss relative to an ideal dipole:

$$\text{dB (loss)} = -10 \log \frac{\text{Efficiency}}{100}$$

Also note that 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 watts! 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.

In the cases that follow, the designs are based on using reactive components (transmission-line resonators or inductors and capacitors) to achieve bandwidth widening. The nonideal nature of real-world components leads to less than 100% efficiency.

Resonators as Matching Networks

The most practical broadbanding network for a dipole is the parallel LC tuned circuit connected directly across the antenna terminals. This circuit may be realized by using a coil in parallel with a capacitor or by using a coaxial resonator. Fig 4 shows the various resonators used for analysis in this article. For the horizontal dipole cases shown below, the stubs are assumed to be made of RG-58A.

The Double Bazooka

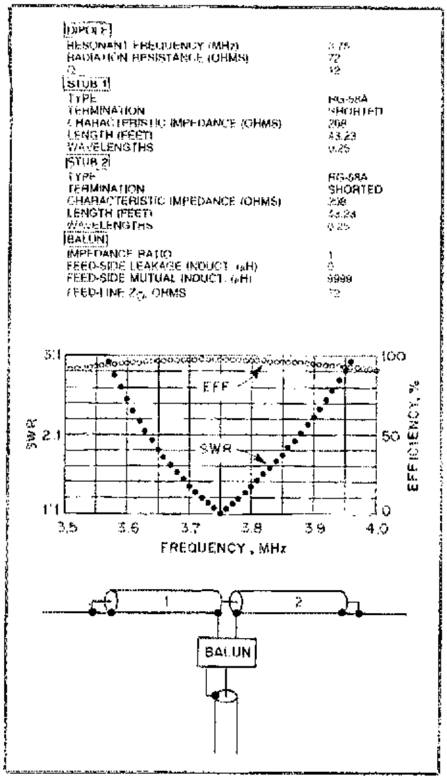
The response of the somewhat controversial double bazooka antenna is shown in Fig 5. This antenna actually consists of a dipole with two quarter-wave coaxial resonators connected in series, Fig 4C. Not much bandwidth enhancement is provided by this resonator connection because the matching network has too high an impedance level. Note the negligibly small bandwidth improvement over the uncompensated dipole of Fig 1. For comparison, it is useful to define *bandwidth improvement factor* (BWIF) as follows:

$$\text{BWIF} = \frac{\text{2:1 SWR bandwidth of broadband antenna}}{\text{2:1 SWR bandwidth of uncompensated half-wave dipole}}$$

The double bazooka BWIF equals 1.14.

The Crossed Double Bazooka

In earlier articles, the crossed connection of a double bazooka antenna has been described.[5] The response of this antenna is



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DIPOLE
RESONANT FREQUENCY (MHz)              3.75
RADIATION RESISTANCE (OHMS)           72
Q                                     12
STUB 1
TYPE                                  RG-58A
TERMINATION                           SHORTED
CHARACTERISTIC IMPEDANCE (OHMS)       208
LENGTH (FEET)                         43.23
WAVELENGTHS                           0.25
STUB 2
TYPE                                  RG-58A
TERMINATION                           SHORTED
CHARACTERISTIC IMPEDANCE (OHMS)       208
LENGTH (FEET)                         43.23
WAVELENGTHS                           0.25
BALUN
IMPEDANCE RATIO                       1
FEED-SIDE LEAKAGE INDUCT. μH          0
FEED-SIDE MUTUAL INDUCT. μH           9999
FEED-LINE ZO, OHMS                    72
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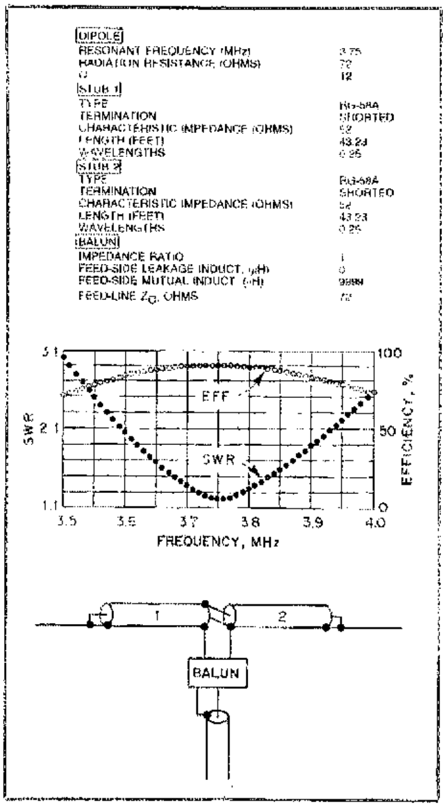
Fig 5—The double bazooka, sometimes called a coaxial dipole. Because the BASIC program assumes that the two stubs are in parallel, the characteristic impedance of the line for each stub is set to 208 ohms. This is equivalent to two 52-ohm stubs connected in series.



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DIPOLE
RESONANT FREQUENCY (MHz)              3.75
RADIATION RESISTANCE (OHMS)           72
Q                                     12
STUB 1
TYPE                                  RG-58A
TERMINATION                           SHORTED
CHARACTERISTIC IMPEDANCE (OHMS)       52
LENGTH (FEET)                         43.23
WAVELENGTHS                           0.25
STUB 2
TYPE                                  RG-58A
TERMINATION                           SHORTED
CHARACTERISTIC IMPEDANCE (OHMS)       52
LENGTH (FEET)                         43.23
WAVELENGTHS                           0.25
BALUN
IMPEDANCE RATIO                       1
FEED-SIDE LEAKAGE INDUCT. μH          0
FEED-SIDE MUTUAL INDUCT. μH           9999
FEED-LINE ZO, OHMS                    72
```

Fig 6—The crossed double bazooka yields bandwidth improvement by using two quarter-wave resonators, parallel connected, as a matching network.

shown in Fig 6. In this case, the impedance level of the matching network, Fig 4D, is reduced to be one-fourth the impedance level of the network of the standard double bazooka. The lower impedance level provides more reactance correction, and hence increases the bandwidth by a noticeable amount. The BWIF equals 1.55. Notice, however, that the efficiency of the antenna drops to about 80% at the 2:1 SWR band edges. The broadbanding, in part, is caused by the resistive losses in the coaxial resonators, which have a remarkably low Q (only 20).

The Asymmetrical Crossed Double Bazooka

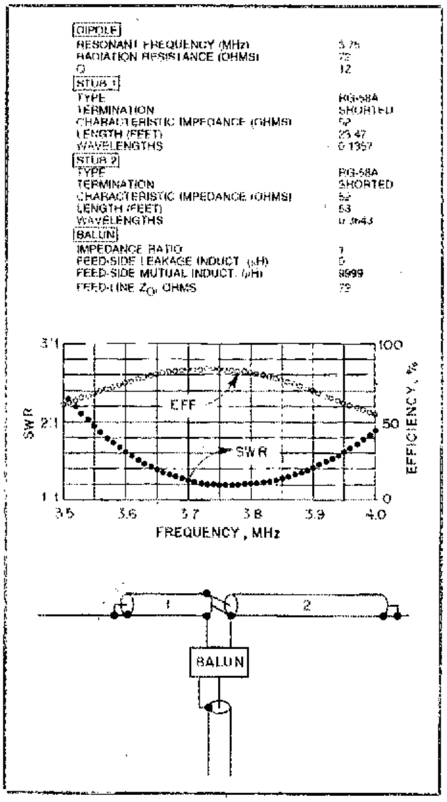
The matching network of the crossed double bazooka may be viewed as a half-wave coaxial resonator with the connection made at the midpoint. The impedance level of the matching network may be reduced to an even lower level by, in effect, tapping the half-wave resonator of the crossed double bazooka at a point off center. This case is shown in Fig 7. Note that by using a lower, more optimum impedance level for the resonator, even more broadbanding is obtained. The BWIF in this case equals 2.16. Notice also that the dimensions of this kind of resonator are still practical since the velocity factor of the coax (0.66 for RG-58 and RG-8) leads to shortened stubs. A variant of this antenna is shown in Fig 8. The half-wave resonator has open ends, permitting even lower matching-network impedance levels to be physically realized. For this case the BWIF equals 2.29.

An even more practical asymmetrical crossed double bazooka antenna may be produced by using a tapped quarter-wave resonator. In this case, one end of the stub is open-circuited while the other end is shorted. This case is shown in Fig 9. The BWIF is again 2.29. Note that the efficiency drops to only about 45 to 55% at the 2:1 SWR limits.

The asymmetrical stubs unbalance the dipole antenna unless the wire portions are made the same diameter as the stubs. The magnitude of this unbalance has not been analyzed, but could be compensated by making the dipole "halves" unequal in length.

It is interesting to note that the effective Q that can be obtained with a quarter-wave resonator is the same as that of either the open- or short-circuited half-wave resonators. The only reason for choosing the longer resonators would be their higher power-handling capability.

Of interest is the Q that can be acquired when resonators are made from coaxial cable. Table 1 summarizes the resonator Q one obtains from different types of coax at 80 and 160 meters. Appendix 2 shows how the Q of coaxial-cable resonators may



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DIPOLE
RESONANT FREQUENCY (MHz)              3.75
RADIATION RESISTANCE (OHMS)           72
Q                                     12
STUB 1
TYPE                                  RG-58A
TERMINATION                           SHORTED
CHARACTERISTIC IMPEDANCE (OHMS)       52
LENGTH (FEET)                         23.47
WAVELENGTHS                           0.1357
STUB 2
TYPE                                  RG-58A
TERMINATION                           SHORTED
CHARACTERISTIC IMPEDANCE (OHMS)       52
LENGTH (FEET)                         63
WAVELENGTHS                           0.3643
BALUN
IMPEDANCE RATIO                       1
FEED-SIDE LEAKAGE INDUCT. μH          0
FEED-SIDE MUTUAL INDUCT. μH           9999
FEED-LINE ZO, OHMS                    72
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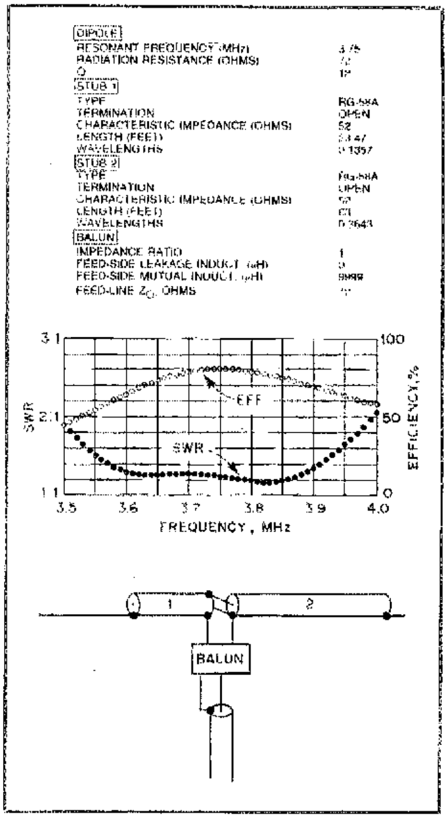
Fig 7—The asymmetrical crossed double bazooka displays even lower matching-network impedances. The resonator may be viewed as a half-wave resonator with shorted ends.



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DIPOLE
RESONANT FREQUENCY (MHz)              3.75
RADIATION RESISTANCE (OHMS)           72
Q                                     12
STUB 1
TYPE                                  RG-58A
TERMINATION                           OPEN
CHARACTERISTIC IMPEDANCE (OHMS)       52
LENGTH (FEET)                         23.47
WAVELENGTHS                           0.1357
STUB 2
TYPE                                  RG-58A
TERMINATION                           OPEN
CHARACTERISTIC IMPEDANCE (OHMS)       52
LENGTH (FEET)                         63
WAVELENGTHS                           0.3643
BALUN
IMPEDANCE RATIO                       1
FEED-SIDE LEAKAGE INDUCT. μH          0
FEED-SIDE MUTUAL INDUCT. μH           9999
FEED-LINE ZO, OHMS                    72
```

Fig 8—Another form of the asymmetrical crossed double bazooka uses a half-wave resonator with open ends to obtain more optimum impedance levels.

Table 1
Resonator Q for Various Types of Coaxial Cable

Cable Type	*Resonator Q*	
	80 Meters	*160 Meters*
RG-174	6.5	4.5
RG-58A	20.0	15.1
RG-141	22.5	16.1
RG-8	41.0	30.6
½-in Hardline	75.5	53.9
¾-in Hardline	109.1	77.1

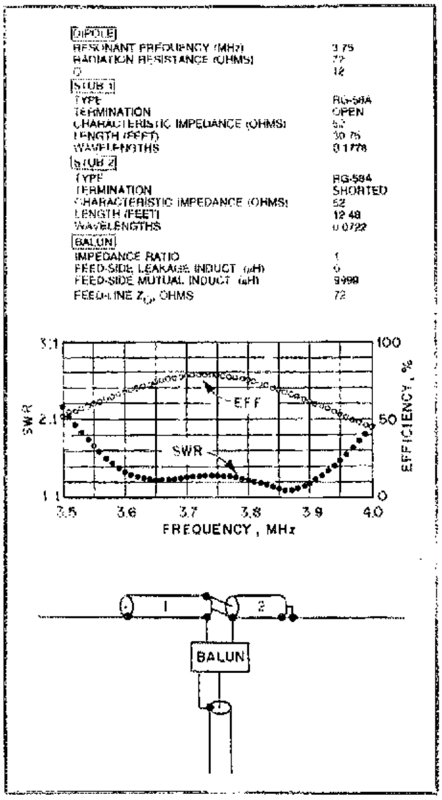


Fig 9—The asymmetrical crossed double bazooka uses a single quarter-wave resonator to achieve essentially the same characteristics as the antenna of Fig 8.

be derived from a knowledge of the cable loss.

Effect of Mismatching

An interesting and useful observation was made during the course of the work described in this paper. The SWR curve of Fig 1 is obtained when the dipole resistance at resonance matches the feed-line characteristic impedance. It is not always convenient to achieve this condition. The feed-line impedance is often either too high or too low. It turns out that mismatching in one direction is far more desirable than in the other. Fig 10 shows how the BWIF is affected by mismatch at resonance. Too high a feed-line impedance actually improves the BWIF up to the point where center-frequency SWR is 1.5. In contrast, a feed-line impedance which is too low, but also yielding an SWR of 1.5 at resonance, causes the dipole bandwidth to drop to ⅔ of the matched bandwidth.

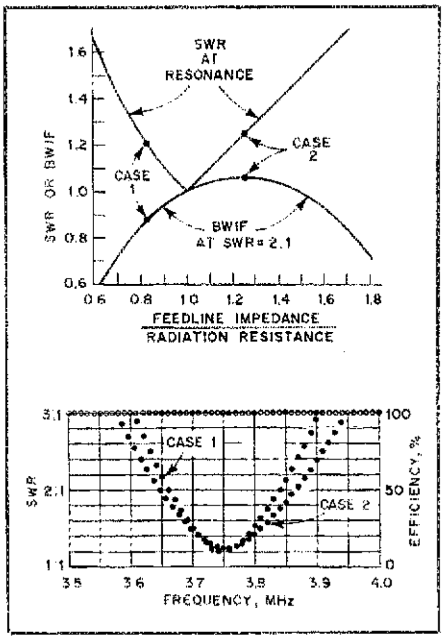


Fig 10—The influence of mismatch on an uncompensated dipole having a radiation resistance of 60 ohms at resonance. Both bandwidth improvement factor (BWIF) and SWR at band center are shown. For Case 1, the feed-line Z_0 is 50 Ω, while for Case 2 it is 75 Ω.

The conclusion to be reached is that greater bandwidth is obtained when an uncompensated dipole is fed with a feed-line impedance that is higher, not lower, than the resistance of the dipole at resonance. For example, a dipole in the inverted-V configuration with a radiation resistance of 60 ohms will have 20% more bandwidth if it is fed with 75-ohm line than if it is fed with 50-ohm line. This result is independent of the dipole Q.

Chebyshev Matching

It is possible to widen the bandwidth further by again resorting to network theory. However, in contrast to the matching with a complementary network (Fig 3), no resistors are introduced. The matching-network parameters are chosen to yield a Chebyshev (often called equi-ripple) approximation. See Appendix 2. The simplest way to make use of this theory for broadbanding the dipole is to deliberately mismatch the dipole at the center of the band by adding a transformer to the matching network. This transformer must provide a voltage step-up between the transmission line and the antenna. The result is a W-shaped SWR characteristic. Low SWR is sacrificed at the band center to obtain wider bandwidth. Incidentally, the Snyder Antenna makes use of this principle.[9] Similar techniques are described in other publications.[10,11]

A broadband dipole using a Chebyshev matching network with a step-up transformer is shown in Fig 11. The transformer can also serve as a balun. Notice that the resonator is the same quarter-wave type as used in Fig 9. The SWR is better than 1.8:1 over the entire 80-meter band, and the BWIF is 2.50—not bad for about 43 feet of RG-58A coax and a slightly modified balun.

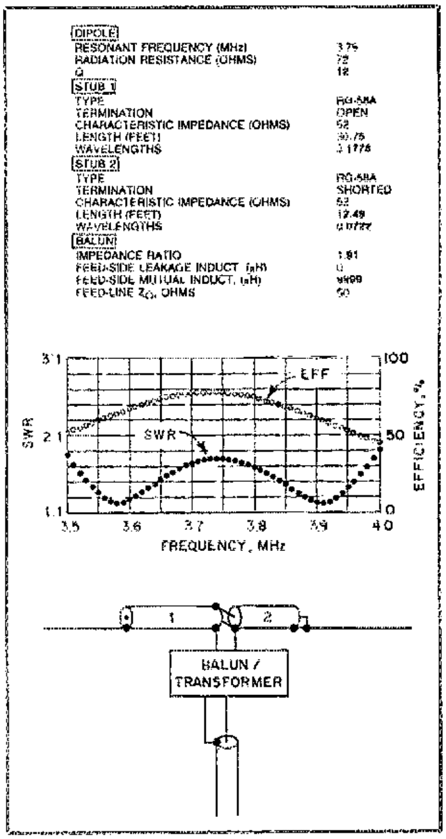


Fig 11—Chebyshev matching provides greater broadbanding by trading midband matching for increased bandwidth. Except for the transformer, this structure is the same as that of Fig 9.

Okay, what's the hitch? Are we getting something for nothing? Not really. Notice that the efficiency in Fig 11 falls to only 45% and 52% at the 80-meter band edges. Only half of the available power is radiated. This low efficiency is directly attributable to the low Q of the coaxial resonator.

LC Matching Network

The efficiency can be improved by using lower loss coax or by using a matching network made up of a high-Q inductor-capacitor parallel-tuned circuit. However, the increase in efficiency is achieved at the expense of bandwidth, as will be seen.

Unfortunately, the very low impedance level required cannot be easily realized with practical inductor-capacitor values. Hence, some form of impedance transformation must be used, as shown in Fig 12. The taps on the coil serve to reduce the impedance level of the matching network, while still permitting the use of practical element values.

The LC network used to provide these impressive broadbanding results utilizes the following: (1) a tuned circuit of very low impedance level, (2) deliberate mismatching at band center to achieve the W-shaped SWR characteristic, and (3) the balun function; that is, balanced (the dipole) to unbalanced (the coax cable) transformation. This network is very similar to one recently proposed, but not tried, by Alan Bloom, N1AL.[12] The difference is that the capacitor is connected across the entire coil in order to obtain practical element values.

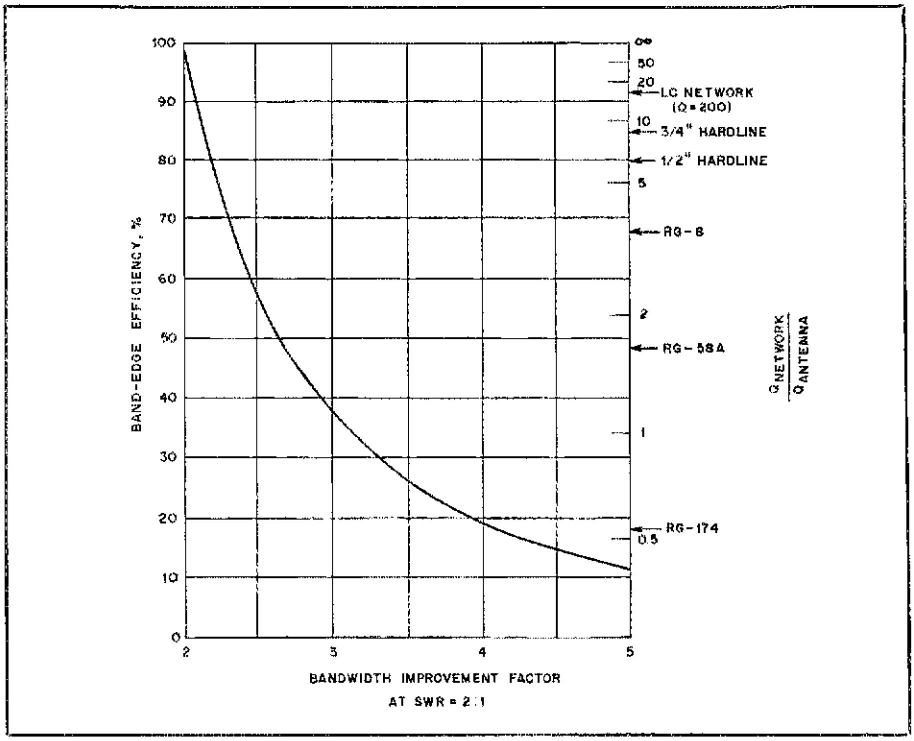


Fig 14—Trade-off between efficiency and bandwidth improvement factor. The points shown for various coaxial cables apply to the case of an 80-meter antenna with a Q of 12.

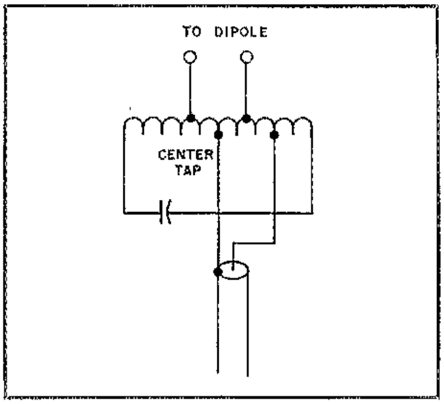
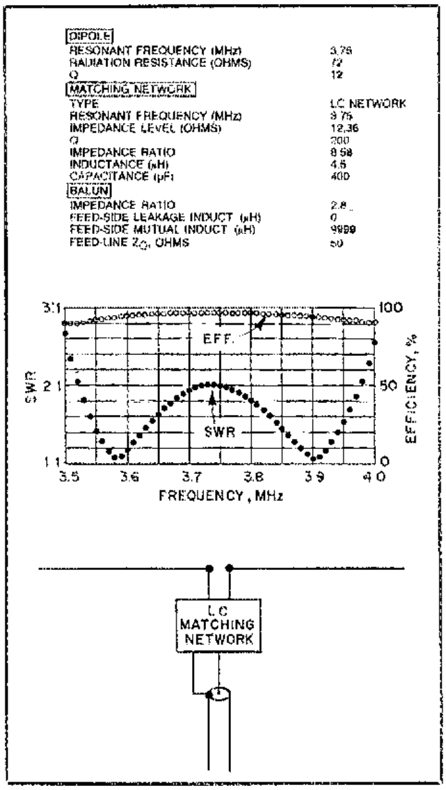


Fig 12—A practical LC matching network which provides reactance compensation, impedance transformation and balun action.



DIPOLE	
RESONANT FREQUENCY (MHz)	3.75
RADIATION RESISTANCE (OHMS)	72
Q	12
MATCHING NETWORK	
TYPE	LC NETWORK
RESONANT FREQUENCY (MHz)	3.75
IMPEDANCE LEVEL (OHMS)	12.36
Q	200
IMPEDANCE RATIO	8.58
INDUCTANCE (μH)	4.5
CAPACITANCE (pF)	400
BALUN	
IMPEDANCE RATIO	2.8
FEED-SIDE LEAKAGE INDUCT (μH)	0
FEED-SIDE MUTUAL INDUCT (μH)	9999
FEED-LINE Z_0, OHMS	50

Fig 13—Efficient broadbanding with an LC matching network.

Fig 13 shows the degree of broadbanding obtainable with a high-Q LC matching network. The BWIF is 2.09. Notice that the bandwidth is not as great as that provided by the coaxial resonator in Fig 11, but the efficiency is better than 90% over the frequency range within the 2:1 SWR bandwidth.

Bandwidth Versus Efficiency Trade-Off

As is apparent in the preceding section, there is clearly a trade-off between bandwidth enhancement and efficiency. This is true because the broadbanding results from two causes: reactance compensation and resistive loading. Pure reactance compensation would be achieved with resonators having infinite Q. The resistive loading caused by nonideal resonators further enhances the bandwidth, but the price paid is that some of the output power heats up the resonator, leading to a loss in efficiency.

The trade-off is clearly depicted in Fig 14, where efficiency versus bandwidth improvement factor is shown for the specific case of Chebyshev matching and a maximum SWR of 2:1. (See Appendix 2.) Note that the best one can do with 100% efficiency is to double the bandwidth. Larger improvements are accompanied by efficiency loss. For example, a tripling of the bandwidth would be obtained with an efficiency of only 38% at the 2:1 SWR band edges.

The graph also shows (on the right hand ordinate) the ratio of the network Q to antenna Q. For the case of an 80-meter dipole with a Q of 12, one can see where on the curve various kinds of commonly used coax would land. For example, RG-8 would yield 68% efficiency and RG-58A gives only 48% efficiency. However, an LC matching network with a Q of 200 would have 92% efficiency.

It should be pointed out that Fig 14 applies for the simple Chebyshev matching that gives a W-shaped SWR plot. It is possible to design a more complex Chebyshev matching network and obtain greater reactance compensation. For example, if an additional inductor and capacitor were added, the bandwidth improvement factor (infinite-Q case) would increase from 2 to 2.8. This rather dramatic improvement could be achieved only at the expense of a more complex matching network which would be more critical to adjust, and the SWR versus frequency plot would have more wiggles.

Experimental Verification

A number of the new ideas presented here were verified by using an inverted-V half-wave dipole with its apex at 60 feet and a 90° included angle. In each case the antenna was made from no. 12 wire. The SWR measurements were made with Bird model 43 and Daiwa model CN-520 (cross needle) SWR meters. Close agreement was obtained with the two meters. All data were corrected for the loss of the transmission line between the SWR meter and the antenna.

The Uncompensated Inverted-V Dipole

Fig 15 shows the measured data for an

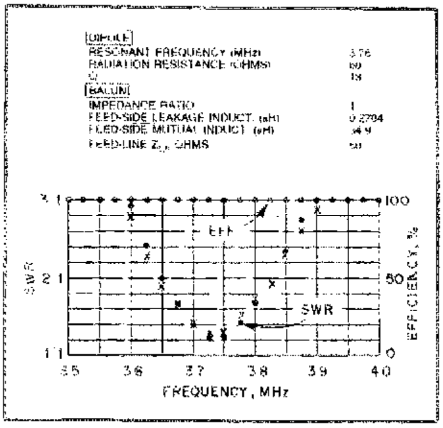


Fig 15—Predicted versus measured SWR for the uncompensated inverted-V dipole. The Xs are plots of the experimental data, while the solid dots are the computer-simulation results.

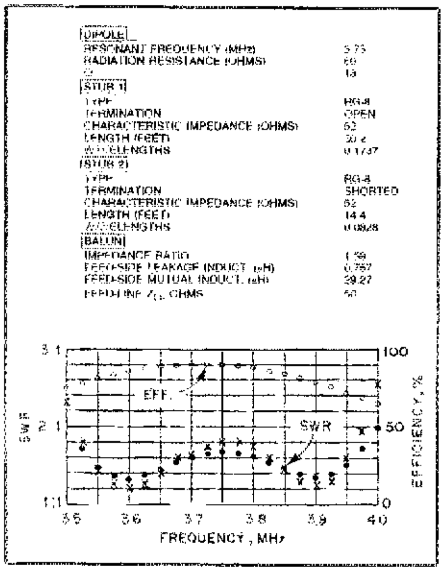


Fig 16—The inverted-V dipole broadbanded with a quarter-wave resonator. In the computer simulation, solid dots, the length of the coax was increased by 3½ inches to account for the physical connection. The Xs are plots of experimental data.

uncompensated inverted-V dipole with a W2AU balun, fed with 50-ohm RG-213 coax. The dipole was 123 feet long. Analysis of these data yielded a dipole Q of 13 and a radiation resistance of 60 ohms at resonance. If this antenna were perfectly matched at resonance, the 2:1 SWR bandwidth would have been 204 kHz.

Inverted-V Dipole with Quarter-Wave Resonator

Shown in Fig 16 is an inverted-V dipole with a quarter-wave resonator. This antenna showed a substantial improvement in bandwidth, to 465 kHz. RG-8 stubs comprised the resonator; a homemade balun transformer with a 1.59:1 impedance ratio was used. A description of the technique for designing and constructing this balun transformer was published earlier.[13]

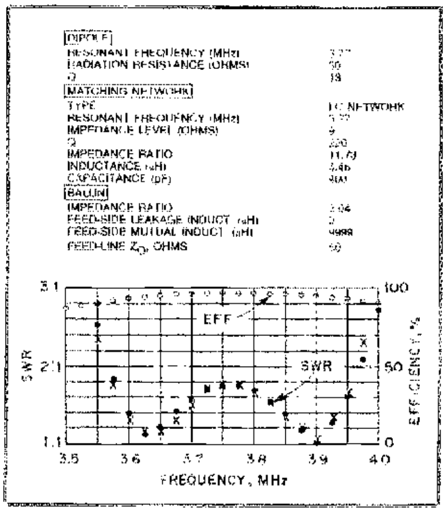


Fig 17—The inverted-V dipole with an LC matching network. Notice the efficiency improvement over that of Fig 16. The solid dots are plots of computer simulation, and the Xs show experimental data.

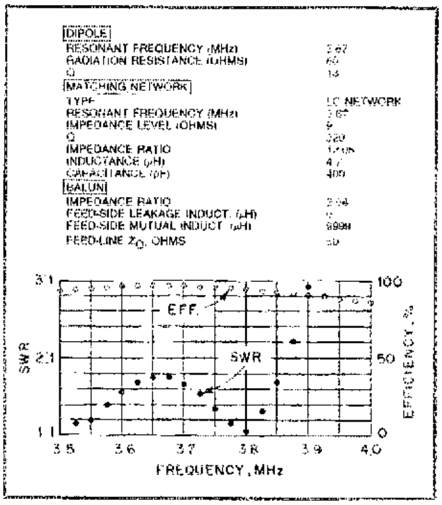


Fig 18—The 80-meter DXer's Delight.

The overall length of the antenna was 130.8 feet. This substantially increased length was necessary to achieve resonance because the coaxial stubs and wire act electrically like an antenna made of tapered material. Taper is known to lead to longer resonance lengths.

No attempt was made to fully optimize the parameters. The computer program was used to calculate SWR and efficiency. The close agreement shows how useful the program is for simulation prior to antenna construction.

Inverted-V Dipole with LC Matching Network

As predicted, the inverted-V dipole with LC matching network, shown in Fig 17, was not as broadband as the coax resonator version. Its 2:1 SWR bandwidth was 405 kHz. The measured Q of the matching network was 220, while the calculated efficiency at the 2:1 SWR band edges was 91%. The antenna length was 122 feet.

Fig 18 shows the calculated SWR and efficiency of a compensated dipole based on a frequency-translated version of the dipole described in Fig 17. It has the SWR minima near 3.5 and 3.8 MHz. This design will hold appeal for 80-meter DXers. A single antenna permits operation with a near-perfect match on interesting parts of the phone and CW portions of the band, with no antenna matching network.

Presented in Fig 19 are photographs of the LC matching network and its weatherproof package. Approximately 8⅓ turns of B&W coil stock, type 3029, are used for the inductor (6 turns per in, 2½-in diam, no. 12 wire). The primary and secondary portions of the coil have 1¾ and 2½ turns, respectively. The coil is resonated at mid-band with a 400-pF capacitor. The capacitor is of the transmitting mica variety, with a breakdown rating of 3000 volts and an RF current rating of 4 amperes.

One must be careful with the selection of a capacitor for this application, especially if high power is to be used. For the capacitor described above, the allowable peak power (limited by the breakdown voltage) is 2450 watts. However, the allowable average power (limited by the RF current rating) is only 88 watts! These limits apply at the SWR = 1.75:1 band edges.

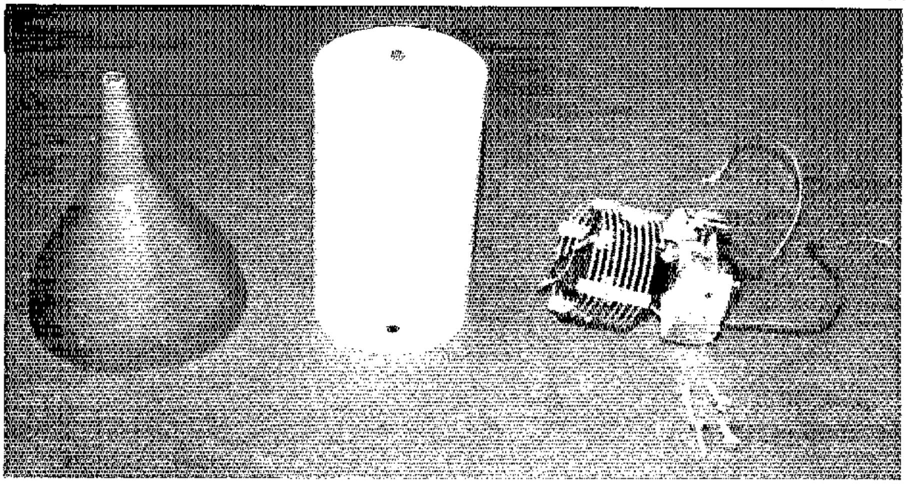
Inverted-V Dipole with Transmatch

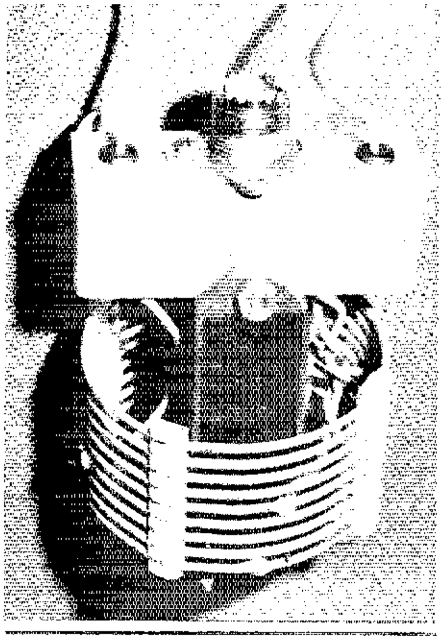
For purposes of comparison, the efficiency of an antenna system using no matching at the antenna, but with a lossless Transmatch at the transmitter end of the feed line, has been calculated. See Fig 20. Here, the efficiency definition is different because the source of loss is different. Remember that in all the cases considered earlier, the efficiency was of concern primarily over the part of the band for which the SWR was less than 2:1. Hence, in those cases the SWR on the feed line was low enough so as not to affect the results significantly. Now it is necessary to take into account the losses caused by the high SWR on the feed line away from antenna resonance.[14]

The particular case considered here is an 80-meter inverted-V dipole fed with 250 feet of RG-213 coax, which has about 1 dB of loss in the matched (SWR = 1:1) case. In this situation, the only lossy element in the antenna system is the feed line. Hence the efficiency of interest is defined as:

$$\text{Efficiency (\%)} = 100 \times \frac{\text{Total power delivered by transmitter minus power lost because of high SWR}}{\text{Total power delivered by transmitter}}$$

This efficiency, plotted in Fig 20, which may be compared with the other efficien-





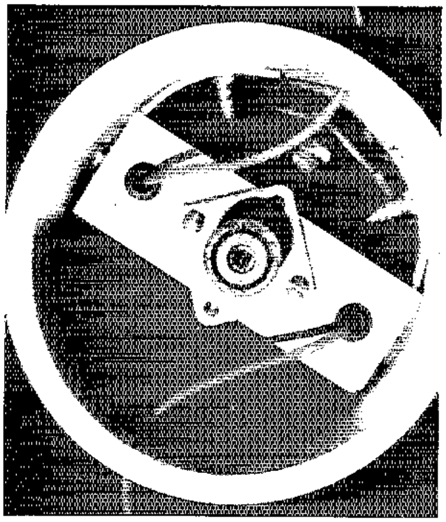


Fig 19—The AI1H matching network follows the circuit diagram of Fig 12. Components must be chosen for a high Q and must have adequate voltage and current ratings. See text for component information.

cies calculated earlier (even though the definition is different), drops to about 74% at the band edges. The result is not bad, especially when compared to the dipoles that use coaxial resonators for broadbanding. Of course, in making the comparison one must recognize the operational inconvenience of having to adjust the Transmatch. Furthermore, some baluns will not handle the large mismatch at the edges of the band and will be destroyed when high power is applied.

Other Bands and Antennas

The computer program is designed to analyze antennas at arbitrary frequencies. For example, a 160-meter half-wave dipole with an LC matching network for broadbanding is shown in Fig 21. Because of the smaller percentage bandwidth of the 160-meter band (compared to the 80-meter band), it is possible to obtain excellent SWR and efficiency performance over the entire band with an LC matching network.

Other resonant antenna systems, such as monopoles and full-wave loops, may be broadbanded by using the same procedures described in this article. For the analysis to apply, it is necessary only that the model

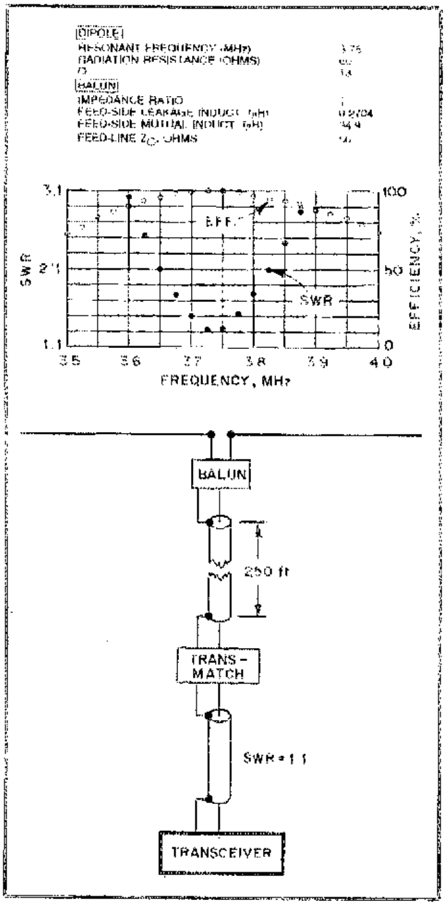


Fig 20—The uncompensated inverted-V dipole fed with 250 ft of RG-213 coax, used with a Transmatch for operation over the entire 80-meter band.

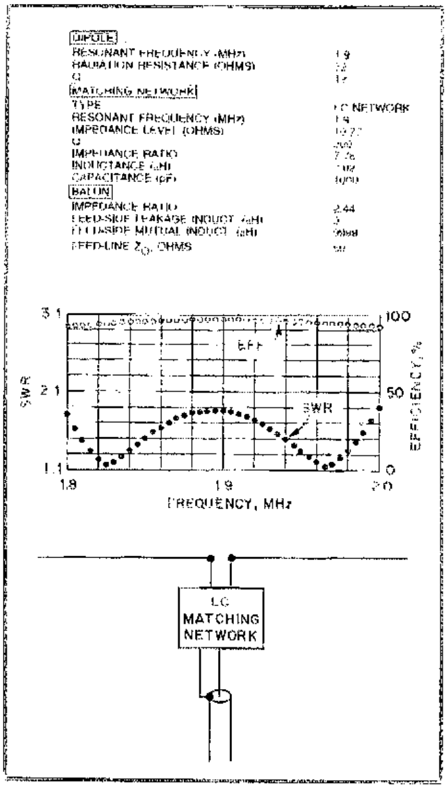


Fig 21—A 160-meter half-wave dipole with an LC Chebyshev matching network.

of the antenna near resonance be valid. See Appendix 1.

Conclusion

A number of dipole broadbanding schemes have been compared. A major point which has been brought out is the importance of evaluating broadbanding techniques not only by comparing SWR results, but also by comparing efficiencies. The superiority of LC matching (compared to transmission-line resonator matching), especially when efficiency is a consideration, has been covered. Design equations which enable one to calculate the optimum broadband matching parameters are presented in Appendix 2.

I hope that this article is viewed as another example of how well the Amateur Radio and home computing hobbies complement one another. I am indebted to my wife, Barbara, N1DIS, to John Kenny, W1RR, and to Joe Reisert, W1JR, for their interest and encouragement during the course of this project.

APPENDIX 1

A by-product of the investigation of dipole broadbanding methods is a flexible, user friendly program which runs on the Commodore 64 computer.[15] The flexibility is partially illustrated by the many examples presented in the article.

The program was written using SIMONS BASIC, which adds 114 commands to the set of BASIC commands built into the C-64.[16] This enhanced BASIC provides a number of commands which simplify the programming process. Of particular value in this application were:

1) Procedures. This feature allows one to call and execute procedures by name (instead of subroutines by line number).

2) Menu processing commands. Coding of menus is simplified by the ON KEY and PLACE commands.

3) Graphics. SIMONS BASIC has a rich set of high resolution graphics commands which enable one to mix graphics and alphanumeric data on the screen.

4) Screen dumps. Both low and high resolution screen dumps are obtainable with single commands.

Many other commands are available with SIMONS BASIC; for example, sprite and sound generation and powerful string-handling commands are provided, but were not used.

The user sees two kinds of presentations when using the program: menus and output data. Three menus are presented. The main menu, Table 2, shows the major tasks to be performed. Another displays the input parameters for the calculation. It is through interaction with this menu that the user can modify the antenna, matching network, balun and feed-line parameters. All of the data associated with the SWR/efficiency versus frequency graphs are replicas of this menu. The frequency menu, also Table 2, allows the user to change the frequency range and the number of points in the band where the analysis is performed.

Output data is provided in two forms: tabular and graphical. An example of the tabular output is given in Table 3, which shows the data associated with Fig 15. The graphs shown in this article were redrawn, but are similar to those available on a video monitor or on a dot matrix printer. The original graphs were produced on a Panasonic KX-P1090 printer with a Cardco +G serial to parallel interface unit.

Table 2
Menus Presented in the BASIC Program, BROADBAND

Main Menu

```
        BROADBAND DIPOLE
   SWR & EFFICIENCY CALCULATION

1  EXAMINE OR CHANGE PARAMETERS
2  CALCULATE SWR & EFFICIENCY
3  PLOT SWR & EFFICIENCY
4  SELECT FREQUENCY RANGE
     BAND                          80 METERS
     LOWEST FREQUENCY              3.5
     HIGHEST FREQUENCY             4
     STEP SIZE                      .05
5  QUIT
```

Frequency Menu

```
         FREQUENCY RANGE

1  BAND                            80 METERS
2  LOWEST FREQUENCY                3.5
3  HIGHEST FREQUENCY               4
4  STEP SIZE (MHz)                  .05
5  NO CHANGE
```

Table 3
Tabulated Output of Author's Computer Program, BROADBAND

Frequency (MHz)	*SWR*	*Efficiency (%)*
3.500	5.88	100.
3.525	4.98	100.
3.550	4.19	100.
3.575	3.50	100.
3.600	2.91	100.
3.625	2.42	100.
3.650	2.00	100.
3.675	1.66	100.
3.700	1.39	100.
3.725	1.22	100.
3.750	1.24	100.
3.775	1.42	100.
3.800	1.68	100.
3.825	1.98	100.
3.850	2.33	100.
3.875	2.73	100.
3.900	3.18	100.
3.925	3.66	100.
3.950	4.20	100.
3.975	4.77	100.
4.000	5.39	100.

The sections that follow cover the basic assumptions and equations that were used in the analysis. Anyone desiring to extend the application of the program, to modify it, or simply to better understand the subject of antenna matching should find this material interesting.

The Dipole Antenna

In order to broaden the SWR bandwidth of a resonant antenna by means of a matching network, it is necessary to have a valid electrical equivalent circuit for that antenna. The half-wave dipole has been analyzed extensively by many researchers over a long period of time. An outstanding summary of the current state of understanding is given by Elliott.[17] Unfortunately, the available theoretical results fall short of our needs. The analyses usually deal with a straight antenna of specific physical dimensions in free space or over a perfectly conducting ground. Some analyses do treat the inverted-V dipole and some take into account a nonideal ground plane, but the real-world antennas we use have steel towers, trees, other antennas and an uncharacterized earth surface nearby.

Fortunately, the real-world dipole can be adequately modeled in the vicinity of resonance by choosing only three parameters: resonant frequency, radiation resistance at resonance, and Q. The equivalent circuit used in the program is shown in Fig 22. It consists of a frequency-dependent resistor in series with an open-circuited quarter-wave transmission line. This model is similar to the familiar RLC series-circuit model, but it is accurate over a wider frequency band. The assumed frequency dependence of the resistor in the model is also shown in Fig 22. It is assumed to vary linearly with frequency; the slope is consistent with data presented by Elliott for thin-wire antennas.[18]

The three model parameters (resonant frequency, radiation resistance and Q) are best determined by simply measuring the SWR of the dipole with no compensating network and using

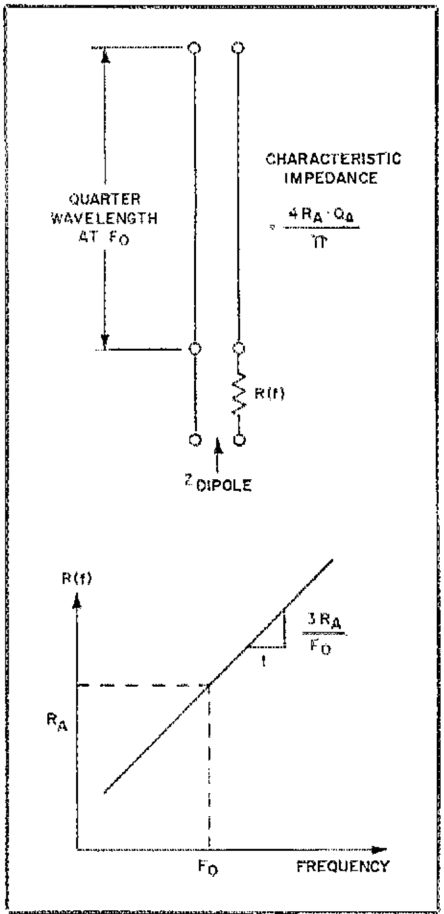


Fig 22—Dipole equivalent circuit. This transmission-line model is more accurate than the often-used RLC equivalent circuit. See Appendix 2 for definitions of terms.

this information to calculate the parameters. This procedure was used to find the dipole parameters of Fig 15. For long feed runs, it is important that the SWR measured in the shack be corrected.

Coaxial Resonator Matching

The transmission-line equations used in this analysis were derived from those in the *Radio Engineers' Handbook.*[19] It turns out that resonators realized from coaxial transmission lines have a very low Q. Hence, it is better to use the more complete formulation which involves complex hyperbolic tangent and cotangent functions to adequately model them. A simpler equivalent parallel RLC tuned-circuit model of the transmission line resonators could have been used, but some accuracy is lost away from resonance.

The program contains the parameters for a wide variety of coaxial cables, including RG-8, RG-58A, ½-in and ¾-in Hardline and RG-141 (Teflon® dielectric).

LC Matching

Since the impedance level required to obtain any significant broadbanding of the dipole is very low, a simple parallel LC model will not lead to practical component values. Thus, taps on the inductor are used, as shown in Fig 12, to obtain transformer action. The program is designed so that the impedance transformation ratio, the inductance, the capacitance, the impedance level of the matching network and its resonant frequency and Q may be prescribed. The effect on other parameters when one element value is changed is automatically determined.

Balun Transformer

An important ingredient in obtaining maximum broadbanding is deliberate mismatching at band center. Thus, a transformer is required. Since the feed line is usually (but not necessarily) coaxial cable, a balanced-to-unbalanced transformation is also required and can be performed by the same physical transformer. It turns out that in this application the nonidealness of real transformers must be taken into account in order to obtain a close match between calculated and measured results.

Fig 23 shows the equivalent circuit used in the program. Much insight may be gained by assuming that the leakage inductance is zero and that the mutual inductance is high enough to have no effect. However, the calculations for real antennas described in the article make use of measurements made on the balun transformers which were used.

It is usually the leakage inductance which has the strongest unintentional influence on broadbanding. Fortunately, the effect of nonzero leakage inductance can be compensated by detuning other antenna parameters. The program permits one to perform such compensation.

In the case of the LC matching network of Fig 12, the functions of matching, impedance transformation and balanced-to-unbalanced conversion are combined in one network. However, these functions are separated in the model on the computer. No loss of accuracy results, but it is important to set the balun mutual inductance to a high value (9999 microhenrys will do), since that element is already modeled by the LC matching network. The leakage inductance of the LC matching network is small enough that it was assumed to be zero for the simulation.

SWR and Efficiency Calculations

SWR is calculated by the computer program at the point in the antenna system where the feed line connects to the balun transformer.[20] The only dissipative elements in the model of the antenna system are the radiation resistance in the dipole model and the ohmic losses in the matching network. In the program, these elements are converted to shunt conductances at the antenna terminals and the efficiency is calculated as follows:

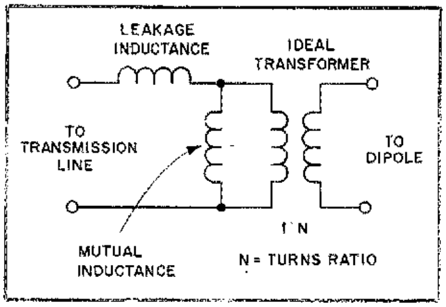


Fig 23—Transformer equivalent circuit.

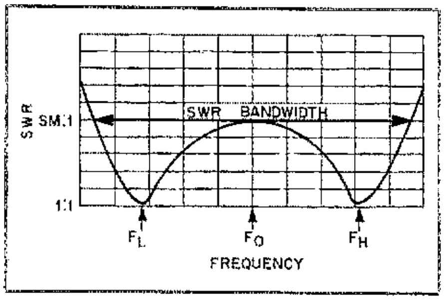


Fig 24—Chebyshev SWR characteristic.

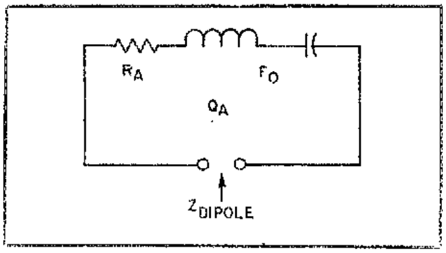


Fig 25—Simplified dipole equivalent circuit.

$$\text{Efficiency (\%)} = \frac{100 \times GD}{GD + GN}$$

where
GD = Dipole conductance and
GN = Network conductance

APPENDIX 2

The computer program described in Appendix 1 analyzes the broadband dipole. It may be used in a "cut and try" procedure to determine the matching network parameters. However, this is a tedious process which may not yield optimum results.

Fortunately, the optimum matching-network parameters (coaxial stub lengths or LC resonator element values and transformer ratios) may be calculated. Fig 24 shows the desired Chebyshev SWR characteristic.

For the purpose of computing the matching network parameters, it is sufficient to use the simple RLC dipole model shown in Fig 25. The differences between this model and the one given in Fig 22 are that a series LC circuit is used to represent the reactive part of the antenna impedance and that the radiation resistance is assumed to be independent of frequency. These simplifications make the mathematics tractable and provide adequate accuracy. The equations given in this appendix are the results of the analysis.

Definition of Terms

Input Data
FO = dipole resonant frequency (MHz)
QA = dipole Q
RA = dipole radiation resistance (ohms)
SM = maximum SWR over band (achieved at band center and band edges)
ZT = transmission line characteristic impedance (ohms)

For Coaxial Stub Resonators
ZS = stub characteristic impedance (ohms)
VF = velocity factor
K, M = attenuation constants

For LC Matching Network
QN = LC resonator Q

Unknown Parameters
BWD = SWR bandwidth of uncompensated dipole (MHz)
BW = SWR bandwidth of compensated dipole (MHz)
BWIF = bandwidth improvement factor
EFC = efficiency at band center (%)
EFE = efficiency at band edges (%)
ZN = matching network impedance level (ohms)
NZ = transformer impedance ratio
FL = lower perfect match frequency (MHz)
FH = upper perfect match frequency (MHz)

For Coaxial Stub Resonators
A = attenuation/100 feet (dB)
QN = resonator Q
L1 = length of first stub (feet)
L2 = length of second stub (feet)

For LC Matching Network
LN = inductance (microhenrys)
CN = capacitance (picofarads)

Dipole Bandwidth

For reference, it is always useful to know the 2:1 SWR bandwidth before compensation:

$$BWD = \frac{FO}{QA\sqrt{2}}$$

Bandwidth after compensation:

$$BW = \frac{FO}{QN \cdot QA} \times \left\{ \frac{(QN + QA)[(2QN + QA) \cdot SM - QA](SM - 1)}{SM} \right\}^{1/2}$$

$$BWIF = \frac{BW}{BWD}$$

Efficiency:

$$EFC = 100\left[1 - \frac{QA}{SM \bullet (QN + QA)}\right]$$

$$EFE = \frac{100\ SM \bullet QN^2}{(QN + QA)[SM \bullet (QN + QA) - QA]}$$

Matching Network Impedance Level:

If the matching-network impedance level and the dipole resonant frequency are known, the matching-network parameters are readily determined.

$$ZN = \frac{RA \bullet SM}{QA} + \frac{RA(SM - 1)}{QN}$$

Transformer Impedance Ratio:

$$NZ = \frac{RA}{ZT}\left(SM - \frac{QA}{QN + QA}\right)$$

Perfect Match Frequencies:

$$FL = (FO^2 + FM^2)^{1/2} - FM$$

where

$$FM = \frac{FO}{2QA}\left[\left(SM - 1\right)\left(1 + \frac{QA}{QN}\right)\right]^{1/2}$$

$$FH = \frac{FO^2}{FL}$$

Coaxial Stub Lengths

In order to determine the lengths of the coaxial stubs,[21] it is first necessary to compute the loss and the Q of the coaxial resonator:

$$A = K \bullet FO^M$$

$$QN = \frac{2.78\ FO}{A \bullet VF}$$

Half-Wave Resonator with Shorted Ends (Fig 4E):

$$L1 = \frac{492\ VF}{\pi\ FO} \sin^{-1}\left(\frac{\pi\ ZN}{2\ ZS}\right)^{1/2}$$

$$L2 = \frac{492\ VF}{FO} - L1$$

Half-Wave Resonator with Open Ends (Fig 4F):

$$L1 = \frac{492\ VF}{\pi\ FO} \cos^{-1}\left(\frac{\pi\ ZN}{2\ ZS}\right)^{1/2}$$

$$L2 = \frac{492\ VF}{FO} - L1$$

Quarter-Wave Resonator with One Shorted End and One Open End (Fig 4G):
For the shorted end,

$$L2 = \frac{492\ VF}{\pi\ FO} \sin^{-1}\left(\frac{\pi\ ZN}{4\ ZS}\right)^{1/2}$$

For the open end

$$L1 = \frac{246\ VF}{FO} - L2$$

Table 4

Broadband Dipole Design Spreadsheet

The data shown here are for the 80-meter dipole of Fig 13. Changing any of the input data values results in recalculation of all data values in the entire sheet.

BROADBAND DIPOLE DESIGN AI1H

	Dipole Center Frequency =	3750 kHz	
	Dipole Q =	12	
INPUT	Dipole Radiation Resistance =	72 ohms	INPUT
DATA	Maximum SWR over Band =	1.7:1	DATA
	Trans Line Char Impedance =	50 ohms	
	LC Matching Network Q =	200	

Dipole Bandwidth (for SWR = 2:1) = 220.97 kHz

Matching Network Type	Band-width (kHz)	Efficiency (percent)		Matching Network Imp Lev (ohms)	Transf Imp Ratio	Freqs Where SWR = 1:1 (kHz)	
		Band Center	Band Edge				
SWR =	1.7:1						
RG-174U	731.73	61.89	20.03	17.93	1.52	3536	3977
RG-58AU	495.58	77.96	50.16	12.72	1.91	3588	3919
RG-141U	482.28	79.55	53.49	12.44	1.95	3592	3915
RG-8U	432.82	86.69	69.06	11.43	2.12	3604	3902
.50 in hl	404.51	91.93	80.99	10.87	2.25	3612	3893
.75 in hl	393.95	94.17	86.19	10.66	2.31	3615	3890
LC	383.03	96.67	92.07	10.45	2.37	3618	3887

Matching Network Type	Stub Char Imp	Stub Length (feet)					
		Half Wave				Quarter Wave	
		Shorted Ends		Open Ends		Shorted End	Open End
RG-174U	50	23.36	63.10	19.87	66.59	15.40	27.83
RG-58AU	52	18.40	68.06	24.83	61.63	12.49	30.75
RG-141U	50	19.60	71.58	25.99	65.19	13.29	32.30
RG-8U	52	17.28	69.18	25.95	60.51	11.79	31.44
.50 in hl	75	20.77	110.43	44.83	86.37	14.37	51.23
.75 in hl	75	20.55	110.65	45.05	86.15	14.23	51.37

LC—Inductance = .44359562 microhenrys
Capacitance = 4060.5835 picofarads

LC Matching Network Parameters:

The inductance and capacitance of the parallel-tuned LC matching network are given by:

$$LN = \frac{ZN}{2\pi\ FO}$$

$$CN = \frac{10^6}{2\pi\ FO \bullet ZN}$$

Bandwidth Versus Efficiency Trade-Off:

The equations used for Fig 14 are based on simplifications of the BWIF and EFE equations given earlier, for the case SM = 2:

$$EFE = \frac{200}{\left(\frac{QA}{QN} + 1\right)\left(\frac{QA}{QN} + 2\right)}$$

$$BWIF = \left[\left(\frac{QA}{QN} + 1\right)\left(\frac{QA}{QN} + 4\right)\right]^{1/2}$$

Design Spreadsheets

The above analysis can be conveniently put to work with the aid of a spreadsheet program. Table 4 shows a specific result which contains, among other things, the dimensions of the coaxial resonator stubs and transformer impedance ratio used in Fig 11. By simply changing the maximum SWR, a new spreadsheet can be created which yields the LC matching network impedance level and transformer impedance ratio used in Fig 13. Or the frequency and maximum SWR can be changed to obtain the parameters for the 160-meter antenna of Fig 21. The specific spreadsheet program used on the Commodore 64 is PRACTICALC II.[22]

Notice that the SWR plots of Figs 11, 13 and 21 do not exactly match the ideal Chebyshev characteristic. The small differences arise because the BROADBAND analysis program uses more accurate models of the transmission line resonators and the dipole than those used in the calculations of this appendix. However, the differences are small enough to have no significance in practice.

Notes

[1] Jerry Hall, "The Search for a Simple, Broadband 80-Meter Dipole," *QST*, Apr 1983, pp 22-27.

[2] Jerry Hall, "Maxcom Antenna Matcher and Dipole Cable Kit," Product Review, *QST*, Nov 1984, pp 53-54.

[3] Richard D. Snyder, "The Snyder Antenna," *RF Design*, Sep/Oct 1984, pp 49-51.

[4] Richard D. Snyder, "Broadband Antennae Employing Coaxial Transmission Line Sections," United States Patent no. 4,479,130, issued Oct 23, 1984.

[5] William Conwell, "Broadband Antennas Employing Coaxial Transmission Line Sections," *QEX*, Apr 1985, pp 8-9.

[6] See note 1.

[7] See note 1.

[8] See note 1.

[9] See notes 3-5.

[10] Richard C. Johnson and Henry Jasik, *Antenna Engineering Handbook*, 2nd ed (New York: McGraw-Hill, 1984), pp 43-27 to 43-31.

[11] R. M. Fano, "Theoretical Limitations on the Broadband Matching of Arbitrary Impedances," *Journal of the Franklin Institute*, Jan 1950, pp 57-83, and Feb 1950, pp 139-155.

[12] Alan Bloom, "Once More with the 80-Meter Broadband Dipole," Technical Correspondence, *QST*, Jun 1985, p 42.

[13] Frank J. Witt, "Top-Loaded Delta Loop Antenna," *Ham Radio*, Dec 1978, p 60.

[14] *The ARRL Handbook* (Newington, CT: The American Radio Relay League, 1985 or 1986 eds) Chap 16, Fig 23.

[15] The BASIC analysis program (called BROADBAND) described in Appendix 1 is available on disk from the author. Also on the disk is the PRACTICALC II spreadsheet file for generating customized BROADBAND DIPOLE DESIGN charts of the form shown in Table 4. Send an SASE to the author for details. A photocopy of the 11-page BASIC program listing is available from ARRL HQ for a $5 copy fee.

[16] *SIMONS BASIC* (Commodore Business Machines, Inc, 1200 Wilson Dr, West Chester, PA 19380).

[17] Robert S. Elliott, *Antenna Theory and Design* (Englewood Cliffs, NJ: Prentice Hall, 1981), pp 277-321.

[18] See note 17.

[19] *Reference Data for Radio Engineers*, 5th ed. (Indianapolis: Howard W. Sams & Co, subsidiary of ITT, 1968), Chap 22, "Transmission Lines," pp 22-1 to 22-42.

[20] See note 19.

[21] See note 19.

[22] PRACTICALC II (PractiCorp International, The Silk Mill, 44 Oak St, Newton Upper Falls, MA 02164).

Frank Witt holds a BSEE and an MSEE from Johns Hopkins University, and has been a licensed amateur since 1948. He has held the calls W3NMU, K2TOP, W1DTV and, presently, AI1H. His ham radio interests have involved a variety of homemade projects, including an RF clipper, a 6502 microprocessor-based controller for modernizing his Heath 2036A 2-meter transceiver and a variety of antenna projects. His top-loaded delta-loop design (referenced in note 13) has received high marks as a vertically polarized DX antenna for 80 and 160 meters.

Frank's wife, Barbara, and three of their sons, Mike, Chris and Jerry, are licensed amateurs, with calls N1DIS, N1BMI, N1BDT and N1BEB, respectively. A fourth son, Tom, plans to get his license soon. Frank can be found on 75 meters (around 3.820 MHz) most Thursday nights, 9:30 Eastern Time, talking to W3OWN and VP9HK.

8877 Linear Amplifier

(continued from page 26)

the new numeric scale on the meter. Be careful not to bend the pointer when reassembling the meter.

Mounting the Front Panel

Remove the masking tape from the panel, and carefully label the front panel with dry-transfer lettering before mounting any components. Mount the meters, potentiometer and switches in place, and complete the wiring. Wire nylon connectors to the power switches, ALC control, multimeter switch and meters to mate with the connectors in the RF deck. This will result in a totally removable panel.

When mounting the front panel to the cabinet, give careful attention to aligning the vacuum-capacitor and band-switch shafts. Loosen the component mounting screws and mate the shafts, then retighten the screws. Continue the alignment process until the controls operate smoothly.

Rear-Panel Assembly

Up to now, nothing has been done to the rear panel. Mount the panel and decide where the components should mount to obtain short connections. Lay the rear panel out in the same way as the front panel, and cut the necessary holes.

Mount the blower off the rear panel, located to allow good circulation of air up to the front of the under-chassis area and back to the tube socket. Positioning of the blower is not critical as long as air is not directly blown across the tube socket, which could cause backpressure.

Mount all components to the rear panel, and mount and wire the panel to the RF deck. The rear panel is not easily removable like the front panel.

Testing the Amplifier and Power Supply

Testing is the *big* moment and the climax of several months' work. First hook up the power-supply control cable and the ground cable. Leave the high-voltage cable off. Test the control circuits to ensure that the power supply can be turned on from the RF deck and the 3-minute time delay works. Make sure that the tube filaments are on. This isn't as easy as with a glass tube, which allows you to see the filaments glow. Let the tube run for about 10 minutes, then turn the power off. Immediately remove the tube and feel if the base is hot.

Test the amplifier first on 20 or 40 meters, since these bands use the midrange of the TUNE and LOAD capacitors. With high voltage applied to the tube, first key the amplifier with no input drive power. The resting idle current should be between 100 and 200 mA on the plate-current meter.

Now apply a little drive power to the tube. The plate current should rise. Move the TUNE and LOAD controls until power output is indicated on a wattmeter. Increase the drive while adjusting the TUNE and LOAD controls to achieve about 20 mA grid current and 650 to 700 mA of plate current, to realize 1500 W output. If the amplifier works properly on this band, proceed to the other bands and repeat the tests.

Efficiency should be at least 60% on all bands, except perhaps 10 and 12 meters, where efficiency may drop to 55%. If efficiency is poor, try moving the coil taps, but recognize that moving the taps also changes the Q of the coil. Decreasing the inductance and increasing the capacitance in the tank circuit will increase the Q. The amplifier may provide more power output over a larger frequency range with a lower tank-circuit Q, but harmonic suppression will also decrease and the amplifier could start generating interference or not meet FCC standards.

I must warn you one last time. *This device could kill you in one instant if you get tied into the high-voltage circuit. Put a good ground on both power supply and RF deck, and treat the equipment with proper respect!*

Conclusion

I have spent many enjoyable hours operating with this amplifier/power supply combination, with nothing but good signal reports. It was an exhausting task, but now that it's finished, I'm glad I did it. Try building one and you'll see what I mean.

Strays

CALL FOR ARTICLES

Space—the final frontier. How many times have you heard this phrase? With the launch of JAS-1, you'll be hearing even more about satellites. What does OSCAR communication mean to you? Or rain scatter on 10 GHz, or earth-moon-earth communications?

Modes A, B and L are accessible to amateurs through the OSCAR and Russian satellites. Future plans include a Phase IIIC, UoSAT 3, and a French "bird." Will you be ready to greet these newcomers? How will you do it?

Share your space communication techniques with others through the pages of *QST*. Submit a detailed summary of your station setup, operating practices and secrets that make those hard-to-get contacts easy. Send your information to Paul K. Pagel, N1FB, Senior Assistant Technical Editor, ARRL, 225 Main St, Newington, CT 06111. An OSCAR enthusiast once said, "It's lonely up there!"