

Chapter 6

Low SWR for the Right Reasons

(Adapted from *QST*, December 1974)

Sec 6.1 Introduction

Chapter 5 concludes with the statement that in transmitter operation, where conjugately-matched coupling is normally used to deliver power to a load through a transmission line, the match is in one direction only — forward. The transmitter (or generator) output is matched to the line, but looking back into the generator, the line is totally mismatched during the time the generator is actively supplying power to the line through the conjugate coupling of the pi-network tank. If the correct procedure is followed, the conjugate relationship may be demonstrated by making impedance measurements in either direction from any point on the line. These measurements will show an impedance $R + jX$ looking in one direction, and the equal but opposite-sign impedance $R - jX$ in the opposite direction. The net reactance of zero obtained from these two impedances proves that the system is resonant! However, these measurements cannot be performed while the generator is active; it must be *turned off* and replaced with a passive impedance equal to its optimum load impedance. When this is done the impedance terminating the generator end of the line will be seen as a dissipative load when measuring the impedance in the direction of the generator.

One of the prevalent misconceptions

is that when reflected power returns to the generator, it sees the plate resistance of the amplifier tube as a dissipative terminating resistance. The fact that dissipation occurs in the impedance that replaces the generator impedance during the measurements described above is partly responsible for the erroneous inference that power reflected toward the generator is dissipated in a similar manner in the internal impedance of the generator. However, when the generator is supplying power to the line, its internal impedance is never seen as a terminating load for power reflected from a mismatched load terminating the line. This is true because of the wave interference between the source wave, the load-reflected wave, and the canceling wave supplied by the coupling network that eliminates the reactance and achieves the total re-reflection of the load-reflected wave. Hence, the line is totally mismatched looking rearward in the direction of the generator. This entire phenomenon is described in detail in Chapter 4.

On the other hand, in laboratory work the signal generator is usually matched in *both* directions. Here the generator has a Z_c source impedance at the output terminals, usually 50 ohms, and it is also isolated from the line with a resistive pad or attenuator. The pad usually has an insertion loss of around 20 dB (sometimes more) with the same impedance as the generator output and line Z_c . Thus, the

generator sees a match looking into the pad. This is because the pad absorbs and dissipates both forward and reflected power like a lossy line, so that only about 1/100 of the source power reaches the load, and any power reflected from a mismatched load is also dissipated to 1/100 of its original value during its return to the generator. As a result, whether the load is a totally reflecting short- or open-circuit termination, the reflected power reaching the source is about 40 dB below, or only 1/10,000 of the power delivered by the generator. Consequently, the power delivered by the source of the generator remains constant, unaffected by the returning reflected power. Thus the pad appears to the generator as either an infinitely long line, or a line having a perfect $R = Z_c$ termination. As a result, reflected power returning from a mismatched termination on the line, the device under test, does not reach the generator with sufficient power to significantly affect its output power, or to modify the impedance which the generator source sees as its load. Consequently the forward voltage entering the transmission line at the output of the pad is held constant, independent of the load impedance presented by the device under test. (*Ref 19, p 48*). Thus, it is understandable that confusion between these two forms of matching can be responsible for misleading us into thinking that reflected power in the transmitter case is dissipated and lost on return to the source.

Sec 6.2 Reflected Power Versus “Lost” Power

The erroneous conception that reflected power is lost is widespread, having been nurtured on the air for a long time, and supported in print in so many published articles it would be impossible to count them. Two such articles, one by

Houghton (*Ref 102*) and the other an SWR-indicator review by Scherer (*Ref 103*), are especially pertinent, because they contain explicit statements supporting the erroneous concept, whereas statements in many other articles only support the error implicitly. Let us now make a further analysis of the reflection mechanics involved in generator matching. In this analysis, two important ingredients that have been overlooked for a long time will be revealed. In so doing, we will see why statements concerning lost power published in the two articles mentioned above are incorrect. We will also see why it was so easy for these ingredients to be overlooked early in the amateur use of coaxial transmission line, with the result that many amateurs have been misled into seeking low SWR for the wrong reason.

Let's consider a lossless transmission line having a perfectly matched load termination. The line is also matched to the generator or transmitter. Under these conditions, there is no reflected power in the line and therefore no reflection loss. Thus the generator delivers what is defined as the maximum available matched power, and the load absorbs all the power delivered. If the load termination is now changed, creating a mismatch between the line impedance Z_c and the terminating load, less power will be absorbed by the load. The amount of reduction in absorbed power resulting from the change in load impedance is the measure of the reflection loss. As the reflected power wave returns toward the generator, it causes a change in the line impedance from Z_c to a complex $Z = E/I$ all along the line. This change is as described in Chapters 3 and 4, and as shown for an SWR = 3.0 in Fig 3-2. When the reflected wave reaches the input terminals of the line, the generator is presented with a change in line-input impedance from the original

Z_c value to some new value determined by the complex E/I vector relationship appearing at the line-input terminals. This new impedance at the line input has exactly the same degree of mismatch to the line Z_c as the mismatched terminating load that generated the reflection. Thus, the line is also now mismatched to the generator in the same degree, and in this condition the generator will automatically make less power available to the line in the amount determined by the resulting mismatch.

The reduction of power delivered to the line is exactly the same amount as the power reflected at the load. In other words, the reflection loss at the load is referred back along the line to the generator. Thus, reflection loss is simply a *non-dissipative* loss representing only the unavailability of power to the load because the generator makes less power available as a result of the impedance mismatch at the line input.

It will now become evident that reflection loss represents only the *unavailability* of power to the load, as we see that the load absorbs all the power the generator makes available to the line. On reaching the generator terminals and causing the mismatch to the generator, the reflected power adds to the reduced source power by exactly the same amount of power as the decrease in power made available by the generator. Since forward power now equals the source power plus the reflected power, the forward power reaching the mismatched load remains the same as before the generator reduced its power available for delivery. Hence, the reflection loss equals the amount of decrease in power made available by the generator. But because the power reflected by the load is now a part of the forward power reaching the load, the forward power continues at the same level

as that originally delivered by the generator prior to decreasing its delivery. Thus the load continues to receive the original amount of power, and reflects the original amount of power, and therefore absorbs *all* of the decreased power delivered by the generator. If an impedance match is now provided anywhere along the line, even at the input terminals, the reflected power is prevented from traveling past the match point toward the generator, as explained in Chapter 4. Thus, the line impedance between the match point and the generator is now unaffected by the reflected wave, and remains at its Z_c value at the input. Consequently the generator no longer sees a mismatch and again delivers its *maximum available* power to the line. The impedance match has thus provided a negative reflection, commonly called “reflection gain,” which exactly equals and cancels the reflection loss. Consequently, all the power delivered by the generator is absorbed in the load in either case — with or without the reflection gain. The generator simply made less power available before the reflection gain restored the matched condition between the generator and the line (*Ref 19 p 37*). So we now ask the question, how does this situation relate to the Houghton nomograph (*Ref 102*), where reflected power is stated to be “lost power,” and to the “useful power” table from the *Knight SWR* indicator review (*Ref 103*)? It is this: This nomograph simply converts SWR back to reflected power, ρ^2 , which is what the SWR indicator actually measures but converts to SWR by means of its scale construction. As discussed in Chapter 3, 2 is the measure of reflection loss or power reflected, which equals the decrease in power made available from the transmitter, calculated directly from the mismatch between the characteristic impedance Z_C of the line and the terminating load im-

pedance Z_L . The reflected power is the square of the voltage (or current) reflection coefficient, ρ , from Eq 3-1 (also see Fig 3-2), but remember further that reflected power ρ^2 is a *non-dissipative* power, because it all eventually reaches the load and absorbed, as explained in Chapters 4 and 5.

The tabulated data in the SWR-indicator-review article correctly lists percentage of reflected power ρ^2 for corresponding values of SWR. However, the “useful-power” column is incorrectly labeled, and is therefore misleading because it is actually listing percentage values of $(1 - \rho^2)$, which is the portion of the maximum-available *matched* power the transmitter actually delivers, depending on the degree of mismatch it sees. In other words, the “useful power” column is simply specifying the amount of power the transmitter will deliver into the mismatch if first tuned to a line having a matched Z_C load, and is then switched to the mismatched load without the benefit of *retuning* or *rematching* to the new impedance at the line input. But we do not operate in this manner — we retune, thereby rematching the transmitter to the new load, and consequently establishing the reflection gain as

$$\left(\frac{1}{1 - \rho^2} \right) - 1$$

which completely cancels the reflection loss ρ^2 and the effect of the load mismatch. The transmitter now returns to

delivering *100%* of its matched available power to the line, whatever the SWR on the line may be! Thus, the two missing ingredients are

- 1) understanding the concept of reflection loss and reflection gain, and
- 2) the discovery that the reflected power is totally re-reflected at the transmitter output terminals, either with or *without* the reflection gain.

It is now evident that the information presented by Houghton and Scherer is not specifying “lost” power at all, but only the non-dissipative *reflection loss* — the amount of power made unavailable by the transmitter until an impedance match provides the reflection gain which cancels the reflection loss and permits the transmitter to again deliver its maximum-available matched power. And as stated on several previous occasions, the conjugate impedance match is automatically attained (sometimes unknowingly) either by proper tuning of the transmitter tank circuit to the complex line-input impedance E/I , or (knowingly) by use of a line-matching network if the pi-network tank of the transmitter lacks sufficient range to obtain the match by itself. How the tank performs the impedance match, and the effects of undercoupling, overcoupling, and possible reactive loading of the tank which can result in the absence of the impedance match are explained in detail in Chapter 7.

Sec 6.3 Reflection Gain

Refer now to Fig 6-1. This figure was

2 Houghton (Ref 102) states that reflected power is lost, accompanied by a nomograph for converting SWR into percent reflected power “to make it easier to determine just how much power is lost.” The reader is invited to read an excellent rebuttal to this nomograph presentation by Anderson (Ref 55).

3 Scherer Ref 103 erroneously shows data on page 90, where 100% minus the reflected power is given as “useful” power. The succeeding paragraph also states incorrectly that the SWR indicator must be placed at a multiple of $(/2$ from the load to indicate the true SWR. This statement is simply untrue! The example demonstrates that the SWR indicator was either not properly adjusted to the impedance of the line, or that it was unreliable, or that no balun was being used to feed a dipole with an unbalanced feed line. See Chapter 2 (statements 21 and 23), Chapter 21, and Ref 38, pp 25-26, and Ref 59.

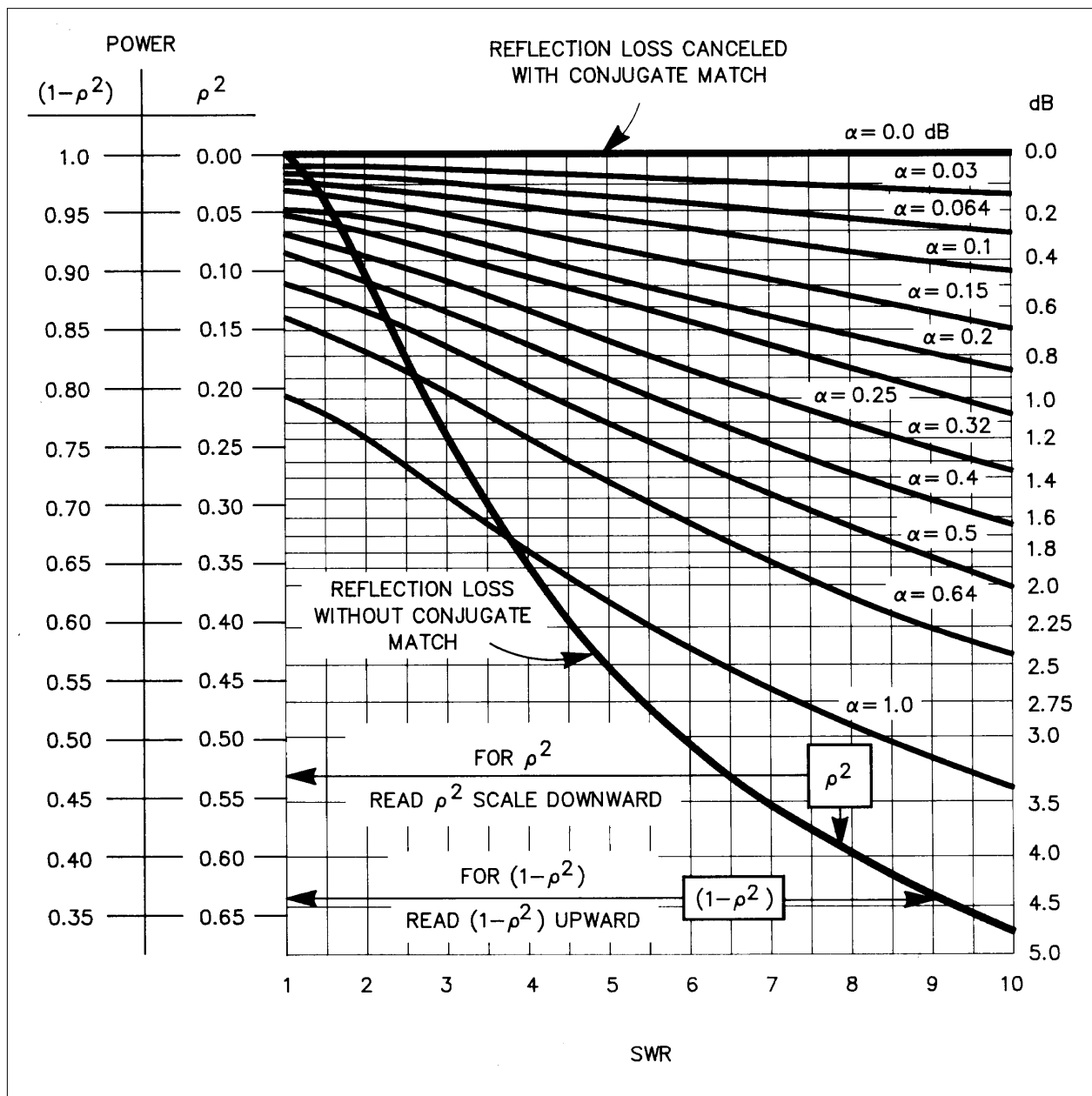
developed to illustrate the concept of *reflection-gain*, in order to emphasize the effect of misinterpreting reflection loss to be “lost” or dissipated power. The impact of this single misunderstanding of transmission-line principles has been disastrous, because it is the principal cause of the prevalent “low-VSWR mania” (low-vis-war-ma’nya). It is the reason why so many amateurs wrongly believe that “getting the SWR down” is the most important factor in “getting the power into the antenna.” They fail to realize that, whatever the SWR with a low-loss feed line, the reflection gain obtained inherently by the tuning and loading procedure, has canceled the effect of the load mismatch if the transmitter can be made to tune and load properly into the line, and that all the available transmitter power is already being taken by the antenna. And therefore, as explained in Chapter 5, they have also been unaware that no more power of any significance reaches the antenna by achieving the lower SWR. They have also been unaware that line attenuation is the key to whether the SWR level has any *practical* effect on efficiency at all.

In Fig 6-1, the heavy curve marked ρ^2 and $(1 - \rho^2)$ is based on lossless-line conditions. It is also an exact replotting of the Houghton “lost power” nomograph (*Ref 102*), and indicates both the reflected and so-called “useful power” columns from the Scherer’s SWR-kit review article (*Ref 103*). Reading downward from the top of the chart, the heavy curve represents reflected power ρ^2 versus SWR. Conversely, the power made available by the transmitter versus the mismatch it sees in terms of SWR, $(1 - \rho^2)$ is found by reading upward from the bottom of the chart. Thus, in reading upward, the curve represents the power being made available with the transmitter tuned for a perfect Z_c match, but actually looking into an

uncorrected mismatch. However, as explained earlier, when the reflection loss is canceled by the reflection gain obtained by re-establishing the impedance match by retuning the transmitter, a new curve, $\alpha = 0.0$ dB, which now represents the newly matched condition, follows *the heavy straight line horizontally across the top of the graph*. This indicates that 100% of the power is being made available, and is also absorbed by the load *regardless of the value of the SWR*. Suddenly the “lost” power is found! Since the curves we’ve just been discussing represent conditions on *lossless* lines, it is evident that neither Houghton nor Scherer was even considering line attenuation in their presentations.

As stated earlier, power can be “lost” in a transmission line only through line attenuation, alpha (α). If the attenuation is zero, lost power is also zero, as shown along the straight-line $\alpha = 0.0$ dB curve at the top of Fig 6-1. When the line has attenuation, power is lost, as shown by the various loss curves marked $\alpha = 0.03$ dB, etc. Since no allowance for the attenuation factor was made in either case in the material presented by Houghton and Scherer, we have still another reason why the terms “lost power” in one case and “useful power” in the other are incorrect and misleading (*Ref 55*).

In Chapter 15 I deal in detail with attenuation effects and show how to perform some of the pertinent calculations based on Eqs 6-1 and 6-2, which are used as the basis for the loss curves appearing in Fig 6-1. However, these loss curves are in a form practical for visualizing the correct relationship between the losses actually encountered in feed lines of various lengths, values of attenuation, and values of SWR. From these curves we can determine the total loss encountered on the line arising from attenuation for the



values of SWR indicated on the scale at the bottom of the chart. The curves represent the condition in which the transmitter is matched to the input of the line, and therefore signify that the effect of the load mismatch is canceled in each case. Each curve starts at the left, where the SWR is 1.0, thus indicating the actual attenuation encountered by a particular line when terminated in a perfectly matched load.

Continuing, the loss value along each curve is seen to increase logarithmically

as the SWR on the line rises with increasing mismatch due to mis-termination of the line. Thus, the difference between the loss incurred with a 1.0 SWR compared to that for any other given SWR value on the same line indicates the amount of additional loss that will be incurred for that SWR. Hence, the graph in Fig 6-1 presents further evidence that when the line attenuation is low, the additional loss from reflection is surprisingly small, even when the SWR is quite high.

Examine the SWR region between 1:1

Fig 6-1— Reflection loss versus SWR and matched-line loss of RF transmission lines. Total attenuation in a line operating with SWR may be determined from the dB scale at the right of the chart. The calibration scales at the left are discussed in the text. The α alpha curves in the body of the chart represent the matched-line loss for any given transmission line. For example, the following types and lengths of line would exhibit the α attenuation factors indicated at the frequency of 4 MHz: $\alpha = 0.03$ dB for 100 ft of no. 12 open-wire line; $\alpha = 0.064$ dB for 20 ft of RG-8; $\alpha = 0.1$ dB, for 100 ft of Amphenol twin-lead, no. 214-022; α alpha = 0.2 dB for 62.5 ft of RG-8; $\alpha = 0.32$ dB for 50 ft of RG-59, 100 ft of RG-8, or 200 ft of RG-17; $\alpha = 0.5$ dB for 87 ft of RG-59 or 175 ft of RG-8; $\alpha = 0.64$ dB for 100 ft of RG-59 or 200 ft of RG-8; $\alpha = 1.0$ dB for 119 ft of RG-58, 350 ft of RG-8, or 700 ft of RG-17. For mathematical derivations of Eqs 6-1 and 6-2 below, see Appendix 5 in Chapter 23. The curves are plots of the following expressions.

$$\text{INCIDENT, OR FORWARD POWER} \quad - \quad \text{POWER REFLECTED} \quad = \quad \text{POWER ABSORBED}^*$$

(multiply by source power delivered)

POWER AT MATCH POINT POWER AT LOAD
(after line attenuation)

$$\frac{1}{1 - \rho^2 \epsilon^{-4\alpha}} \quad - \quad \frac{\epsilon^{-2\alpha}}{1 - \rho^2 \epsilon^{-4\alpha}} \quad = \quad \frac{\rho^2 \epsilon^{-2\alpha}}{1 - \rho^2 \epsilon^{-4\alpha}} \quad = \quad \frac{(1 - \rho^2) \epsilon^{-2\alpha}}{1 - \rho^2 \epsilon^{-4\alpha}} \quad (\text{Eq 6-1})$$

With lossless line ($\alpha = 0$),

$$\frac{1}{1 - \rho^2} \quad - \quad \frac{1}{1 - \rho^2} \quad = \quad \frac{\rho^2}{1 - \rho^2} \quad = \quad 1 \quad (\text{Eq 6-2})$$

where

ρ = magnitude of voltage-reflection coefficient (see Chapter 3, Sec 3.1)

α = line attenuation in nepers = dB + 8.6859

ϵ = 2.71828, the base of natural logarithms

$$* \text{ POWER ABSORBED} = \frac{(1 - \rho^2) \text{ less one-way line attenuation}}{1 - (\rho^2 \text{ less two-way line attenuation})} \times \text{source power}$$

and 2:1. Do you see enough difference in power level on any of these curves to justify any effort in reducing a 2:1 SWR to *any* lower value whatever? Do you still think you'll get out better by squeezing that SWR of 1.8 down to 1.2? A review of Chapter 1 is now appropriate to re-em-

phasize how the use of these concepts can broaden our design flexibility. It is also reassuring to check the efficiency values of both the TIROS and NAVSAT spacecraft feed-line examples from Chapter 1 in the Fig 6-1 graph. In the NAVSAT example, comparison of the 65% reflected

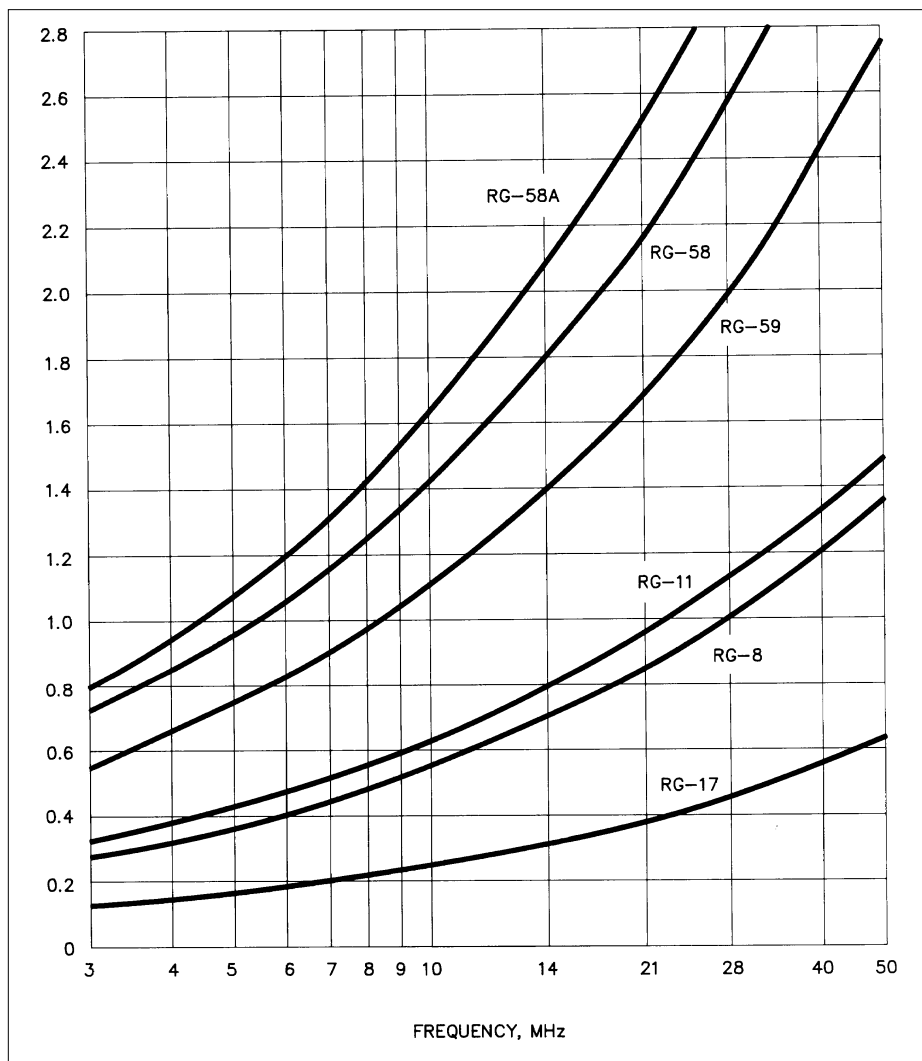


Fig 6-2 — Attenuation in decibels per hundred feet for various coaxial cables.

power with only 1.04 dB actual line loss is especially revealing.

Fig 6-2 provides additional line-attenuation data, permitting us to extend the use of the Fig 6-1 curves to other frequencies and transmission lines. Fig 6-2 may also be supplemented with further data available in *The ARRL Handbook* and *The ARRL Antenna Book (Refs 1, 2 and 71)*.

Sec 6.4 Radiation Resistance

In Chapter 2, Statement 26 says, in effect, that with mobile whip antennas in the 80- through 10-meter bands, no significant amount of power is saved by using matching circuitry between the feed

line and the antenna terminals. And Statement 27 goes on to say that in the absence of such matching circuitry, more power is radiated from center-loaded mobile antennas that have a high feed-line SWR at resonance than those that have low SWR. The concepts involved in those two statements are also widely misunderstood, so this is an appropriate time to clarify both of these statements. These concepts also fit the category of “low SWR for the wrong reasons.”

It is well known that the radiation resistance of the short whip-type mobile antenna is very low. And of all the HF amateur bands, the radiation resistance is the lowest at 80 meters, because the

electrical length of the radiating portion is the shortest on this band. Depending on the exact length of the antenna and other factors, the radiation resistance of the center-loaded antenna is approximately 1 ohm at 80 meters, as shown by Belrose (*Ref 60*). The capacitive-reactance component in the terminal impedance of this short antenna ranges from $-j3000$ to $-j3500$ ohms in typical 80-meter models, as shown by Belrose and also confirmed by my own measurements using a General Radio 1606-A RF impedance bridge. This capacitive reactance is canceled by the equal $+jX$ inductive reactance of the loading coil.

However, it is not well known that there are two other resistances which become important for consideration in antennas of this type. These resistances, from loading-coil loss and ground loss, add to the radiation resistance to comprise the total resistive component of the impedance appearing at the antenna terminals. Thus, by ignoring these two resistances, it is erroneously thought by many amateurs that the 1-ohm radiation resistance alone comprises the entire antenna-terminal impedance, and also that it requires a matching device at the antenna to match what is thought would be about a 50:1 mismatch if fed directly with a 50-ohm feed line. Actually, the loss resistance in the loading coil and any ground-loss resistance both add to the radiation resistance, causing the terminal resistance of the antenna to be much higher than is commonly realized, although still lower than the 50-ohm feed lines that are normally used. Thus, the actual mismatch value is much lower than is usually realized.

While there are some who recognize that loading-coil loss appears as part of the total terminal impedance, only a few are aware that ground-resistance loss also

exists, because, except for Belrose, most writers neglect to mention it or consider it in their system analysis. For example, see Swafford (*Ref 116*).

The *Mobile Handbook* (*Ref 117*) not only fails to recognize the existence of ground resistance, but misconstrues what is actually the combined ground and radiation resistances as radiation resistance alone. For example, in subtracting 6 ohms of terminal resistance, the 8-ohm difference is simply taken to be radiation resistance, with no mention of any ground resistance. Hence, the ground resistance, which cannot be ignored, is unknowingly and improperly being included as a portion of the radiation resistance instead of as a loss resistance in the expression for efficiency. This oversight might have been avoided if an analysis had been made of the increase in radiation resistance actually obtained by raising the loading coil from the base to the center of the whip, because the 6 ohms of coil resistance plus a 1-ohm radiation resistance subtracted from the 14 ohms of measured terminal resistance leaves 7 ohms, which requires an explanation of where those remaining 7 ohms of resistance came from. Obviously, it is the ground resistance that was ignored.

From further study of the *Mobile Handbook* text, however, it is evident that an assumption of a greater amount of radiation from the coil than what is actually possible may have been the reason why such a high value of radiation resistance was considered plausible. But whatever the reason, we have been given the unrealistically high radiation resistance of 8 ohms and an impossible efficiency value of $8/14 = 58\%$. These values are fundamentally impossible to obtain in the center-loaded mobile antenna, and the true values have been obscured. In other words, a large portion of the power con-

sidered in the *Mobile Handbook* as being radiated is actually dissipated as heat in the ground. Belrose shows a proper analysis (Ref 60), which is supported by my own measurements.

It is practically impossible to obtain a mismatch of sufficient magnitude that requires any matching circuitry between the feed line and a properly resonated, conventional center-loaded mobile antenna for the purpose of conserving any significant amount of power. This statement is true, the many remotely controlled luggage-compartment tuning and matching arrangements notwithstanding. Now we will see why.

I have made measurements on loading coils and antennas using a Boonton 260-A Q meter and a General Radio 1606-A RF impedance bridge. Loading-coil loss resistances range from about 8 ohms for the better commercially available coils to as high as 31 ohms measured in poorer coils, depending on the Q and the self-resonant frequency of the coil. Ground-loss resistances encountered with conventional low-band mobile antenna installations range from about 5 ohms for low, wet ground to around 12 ohms for high, dry ground. Average ground yields about 7 ohms.

The ground-loss resistance in the mobile setup is less than that found in an antenna of a full $\lambda/4$ in physical height with no radials, because the radius of the circle where the minimum space-displacement currents return to the ground is shorter with the shorter antenna. Therefore, the return currents travel a shorter distance through lossy earth (see Fig 5-1). The current-flow pattern in the mobile system is also described by Belrose.

Consequently, from all of this we can see that the total terminal resistance of the mobile antenna is nowhere near the 1-ohm radiation resistance alone (which

would produce an SWR of 50:1), but it also includes the ground-loss resistance and the coil-loss resistance, all three appearing in series. The absolute minimum resistance is $1 + 5 + 8 = 14$ ohms, for an SWR of 3.5:1 at resonance when using the low-loss loading coils over good ground. The resistance can be as high as $1 + 12 + 31 = 44$ ohms for an SWR of 1.1:1 when the higher ground loss and higher coil loss occur simultaneously. Hence, the actual resistance appearing at the antenna terminals lies in the range from 14 to 44 ohms, with corresponding SWR values of 3.5:1 and 1.1:1.

Thus, as strange as it may seem, the *higher* the minimum SWR attainable *at resonance*, the greater the power will be radiated for the same amount of power delivered to the feed line by the transmitter. It does not seem so strange, however, when we consider that the low radiation resistance of around 1 ohm is the only portion of the total terminal resistance that contributes to radiation. And this 1 ohm is constant, fixed by the radiator length. So by making the loss resistance lower through the use of a higher Q loading coil, less power is dissipated as heat in the coil, leaving more to be radiated. Conversely, if the lower Q coil is used simply to achieve a lower SWR, less power is radiated because more power is now being spent in heating the coil.

Many amateurs unknowingly select loading coils of low Q because “they produce a lower SWR than coils having a higher Q.” Some lower Q coils, such as the Hustler manufactured by Newtronics, are advertised as producing a lower SWR and greater bandwidth. They do, indeed, produce a lower SWR and greater bandwidth because of their higher loss resistance, which results in a substantial loss of power. However, when using the higher Q, lower loss loading coil, even though its

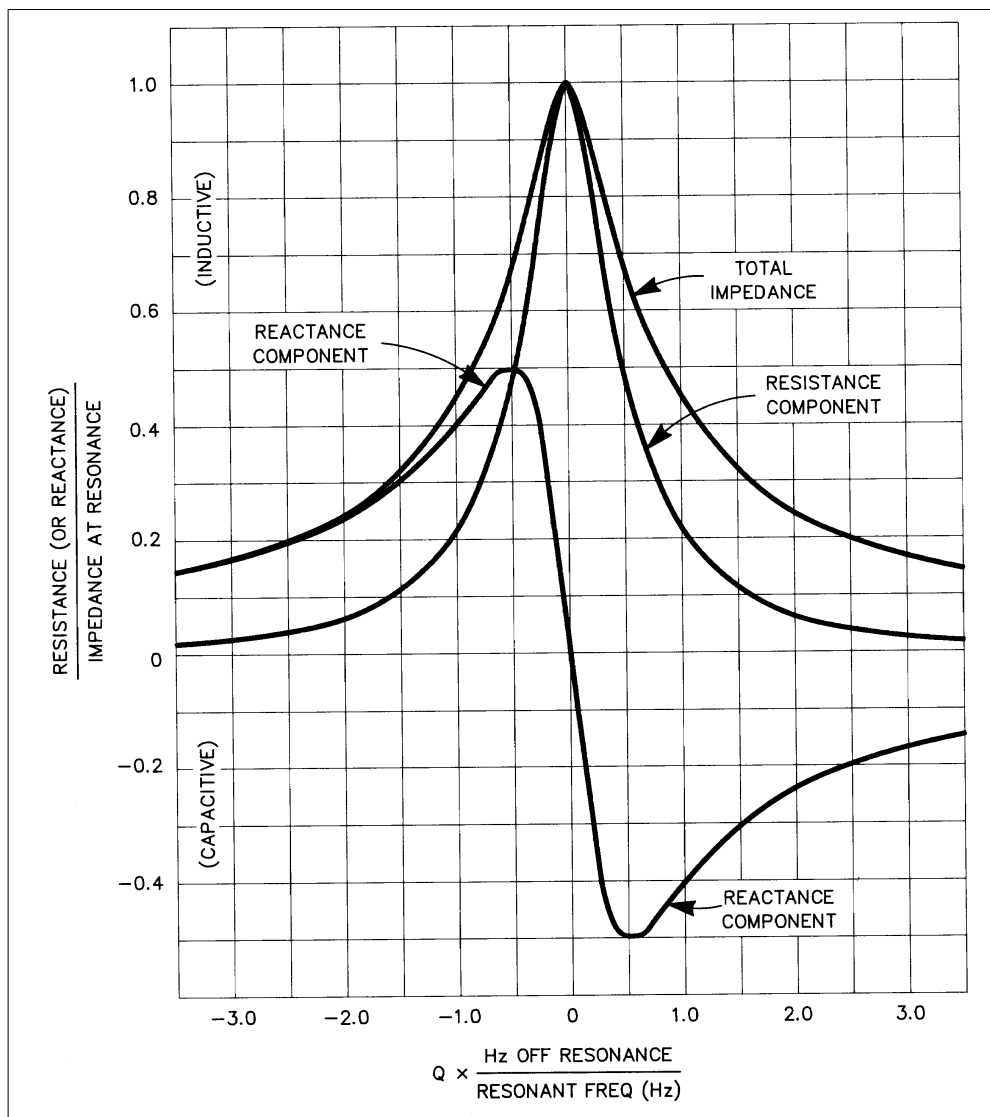


Fig 6-3 — Universal curves giving the resistance and reactance components of the impedance of a high-Q parallel-resonant circuit as a function of frequency. After Terman (Ref 126, p 145)

lower loss resistance results in a larger load mismatch and higher feed-line SWR, the resulting increase in radiated power with the higher Q coil is still proportional to the decrease in total resistance. Any additional loss from the higher SWR is so small that it can be neglected, because line attenuation in the short feed lines used in mobiles is extremely low. Remember, line attenuation is the only cause of power loss in the feed line, regardless of the SWR level.

I don't know whether it is through unawareness or by intent in the design

that Hustler loading coils have close-wound turns (no space between turns). Close-wound turns result in higher distributed capacitance, which contributes to the high loss resistance that provides the low SWR and greater bandwidth of these coils. Let me digress here for a moment to examine the nature of coil resistance so that we may understand why some loading coils have high Q and low loss, while others have low Q and high loss, depending on their design.

The AC resistance loss in inductors is comprised of two separate resistance

components, resistance from skin effect, related to wire size, and resistance from distributed capacitance, related to the spacing between turns. For a coil of given length, diameter and number of turns, there is an optimum wire size that yields minimum coil resistance and maximum Q. On one hand, the largest wire that will fit the given length (no space between turns) yields minimum resistance from skin effect. On the other hand, when the turns are close spaced, the resistance from distributed capacitance is maximum. Spacing the turns apart (by using smaller wire for the same number of turns over the same length) reduces the resistance from distributed capacitance. Hence, the optimum wire size is a trade-off between a large wire that minimizes the resistance from skin effect and a small wire that minimizes the resistance from distributed capacitance because of the increased spacing between the turns.

The decrease in skin-effect resistance from increasing the wire size is obvious. However, the decrease in resistance from lower distributed capacitance with the smaller wire size is not obvious, so I'll now explain why this is true. It is well known that there is distributed capacitance between turns of an inductor. The closer the spacing between the turns, the greater the distributed capacitance, with maximum capacitance appearing when there is no spacing (except for insulation). The distributed capacitance appears inherently in parallel with the inductance of the coil. Consequently, there is a frequency (above the operating frequency of the coil) where the reactance of the distributed capacitance is equal to the inductive reactance of the coil, and the coil becomes self resonant.

At the self-resonant frequency of the coil, the impedance appearing at its terminals is a very high, pure resistance

(which we don't want in a loading coil), and the natural inductive reactance disappears. At this frequency, the coil no longer supplies any inductive reactance to cancel the capacitive reactance of the antenna. Instead, the coil simply introduces its high resistance in series with the antenna terminals, and the dipole portion extending above the coil effectively disappears.

The curves of Fig 6-3 show the relative impedance, resistance and reactance components of the impedance versus frequency in parallel-resonant circuits, such as we have in the self-resonant inductor. For a given frequency, the resistance curve shows the amount of coil resistance relative to the maximum resistance that appears at the self-resonant frequency. We know that in a resonant circuit and with a given inductance, the smaller the capacitance, the higher the resonant frequency. Hence, in the loading coil, the lower the distributed capacitance (from increased turn spacing), the higher the self-resonant frequency and the lower the loss resistance for a given operating frequency. Certainly the design goal for obtaining minimum loss resistance in an inductor is to make its self-resonant frequency as high as possible relative to the operating frequency by optimizing the turn spacing. Because turn spacing (for a fixed winding length and number of turns) has no substantial effect on the amount of inductance in the coil, there is an optimum spacing versus wire diameter that yields the lowest resistance, and hence the highest Q.

It is interesting to note that of the commercially available 80-meter mobile loading coils I have measured, the Hustler coils having the loss resistance of 31 ohms at 4.0 MHz were self-resonant at 6.0 MHz. The better Webster KW-80 coils having 8 ohms of resistance at 4.0 MHz

were self-resonant at 14.0 MHz. The windings in the Hustler coils are close spaced, while the spacing between turns in the Webster coils is approximately equal to the wire diameter. Need I say more?

Before leaving the subject of mobile loading-coil loss relative to distributed capacitance, here is some important information and a warning concerning the use of top-hat capacitance loading. When using an adjustable loading coil, the increased capacitance to ground afforded by top loading can reduce the coil loss, and thus improve the efficiency of the loading coil substantially. This is because the extra capacitance in the top loading reduces the amount of coil inductance required to obtain resonance, and hence reduces coil loss. However, to obtain the maximum improvement in efficiency, the top hat must be placed at the top of the radiator, not part way down. If the hat is large enough to be effective, any portion of the radiator protruding above the hat is ineffective.

And now the warning. I have seen many top-loading devices placed directly above the loading-coil inductor, with the entire top portion of the radiator extending above the top hat. This placement is incorrect, an absolute no-no! This incorrect placement of the hat causes an undesirable increase in the hat-to-coil capacitance. That capacitance adds to the already undesirable distributed capacitance of the coil, further reducing the Q and increasing the coil resistance as described earlier — just the opposite of the hat's intended purpose. Consequently, the coil performance is degraded by the hat, rather than enhanced. This incorrect placement of the hat does reduce the resonant SWR, as many have discovered, but this reduction in SWR is for the same wrong reason as explained earlier—it results from the increased resistance of the

coil. For further information on capacitive loading, I refer you to a section I wrote that appears in *The ARRL Handbook* on “Top-Loading Capacitance.” It appears in Chapter 33 of all editions from 1986 through 1994 (*Ref 1*).

In Fig 6-1, the loss curve = 0.064 dB represents the loss characteristics of a typical mobile feed line — 20 feet of RG-8 at 4.0 MHz. The curve shows a matched-line attenuation of 0.064 dB, plus an additional loss of 0.056 dB resulting from the 3.5:1 SWR, for a total loss of 0.12 dB. When this condition is terminal resistance and a transmitter power of 100 watts, the difference in power radiated between matching at the antenna terminals and leaving the 3.5:1 SWR on the line and matching at the line input amounts to less than 0.1 watt!

It is also of interest to note that with the lower loss resistance ratio (1 to 14 ohms) while using the high-Q coil, the radiating efficiency is 7.14%, or 11.46 dB below the transmitter power delivered (excluding line loss). With the 1 to 44-ohm resistance ratio (with the higher loss coil and poor below the transmitter power. We thus have a 5.07 dB loss in efficiency in return for lowering the SWR from 3.5 down to 1:1 by using a “low-SWR” loading coil. Contrast these actual values of efficiency with the 58% recited in the *Mobile Handbook* (*Ref 117*).

So, contrary to statements found in numerous articles which have been insisting that we believe otherwise, no significant improvement in efficiency can be obtained on the lower frequency bands by performing the matching function between the feed line and the mobile antenna terminals when a low-loss loading coil is already in use. The matching can be performed equally well at the input of the feed line, either by the transmitter output tank itself, or by a separate match-

ing network if the transmitter tank lacks sufficient range, with the feed line connected directly to the antenna terminals! See Refs 4, 24 and 61. Thus, as emphasized in the opening paragraphs of Chapter 5, the important point here, again, is that a flexible, open choice is available in our system design.

Whether the matching which is required to transfer maximum the load end of the feed line is a choice which should be determined according to personal preference of the operator. It should be based on convenience and accessibility to adjustment, and not on an arbitrary, low SWR dictated for the wrong reason by a decree of an SWR King who doesn't understand his Subject! But wherever the matching is performed during operation, the antenna that will radiate the strongest signal is the one with the loading coil that is capable of producing the *highest SWR at resonance* with no matching at the antenna terminals, for the reasons explained above.

Now I'll again explain why loading coils of low Q have a high resistance that reduces the SWR, but also reduces the radiated power. A principal contributor to low Q in loading-coil inductors is the careless (or intentional?) lack of attention to coil design, particularly in relation to the spacing between turns. As is well known, there is distributed capacitance between every turn. The total distributed capacitance resonates with the coil inductance at some frequency, and at this frequency the coil becomes a self-resonant L-C tank circuit. As is also well known, the series AC resistance in a resonant tank circuit is very high. Insufficient spacing between turns results in a large distributed capacitance, which lowers the self-resonant frequency. Hence, the effective RF resistance of the coil will be high unless the operating frequency is substantially lower than

the self-resonating frequency. On 80-meter coils, where I measured 31 ohms of series resistance at 4.0 MHz, the resonant frequency of the coils was around 6 MHz. On the higher Q coils that measured 8 ohms, the resonant frequency was around 14 MHz — a considerable difference!

Finally, here is a suggestion that may be helpful in tuning a mobile antenna. The use of a dip oscillator or a noise bridge in an attempt to determine the resonant frequency of the coil and radiator combination can rarely give the correct answer when you measure at the input terminals of the feed line. The reason is that when using these measuring devices in such a way, they are measuring the resonant frequency of the entire system, including the length of the feed line. As the length of the feed line is changed, the resonant frequency of the antenna system will also change. However, an SWR indicator connected directly at the input of the feed line provides an accurate indication of antenna resonance at the frequency of lowest SWR, providing the instrument is reliable and accurately calibrated for the impedance of the line (Ref 59). This is because the minimum SWR for a given terminating load of the type we're discussing occurs at the resonant frequency of the load, regardless of the length of the feed line.

Sec 6.5 Low SWR for the Right Reasons

It is of interest to know that in TV broadcasting, where a long feed line is required to reach the antenna on a high tower, low SWR is an absolute necessity. However, the requirement here is primarily to avoid multiple, displaced, ghost images from appearing in the received picture images which would result from reflections on the feed line caused by a mismatch. Similarly, low SWR on the feed

line is a necessity in FM stereo broadcasting to avoid cross contamination between the two audio modulation channels. However, in Amateur Radio we do not have the problems encountered in TV and FM broadcasting. To summarize the discussions of SWR and reflections as they pertain to Amateur Radio operations, we have seen that *we do not need* a low SWR on the antenna feed line

1) to prevent reflected power from dissipating in the transmitter, because none dissipates in the properly coupled transmitter anyway, whatever the SWR;

2) to prevent feed-line radiation or TVI, because a mismatched load on the feed line doesn't cause feed-line radiation or TVI;

3) to attain proper coupling to the transmitter, because we can couple to or match the impedance at the input terminals of the line, whatever the SWR.

From examining Figs 6-1 and 6-2, it is evident that we do not need an SWR lower than 2:1 on any feed line to avoid any significant loss in efficiency, or with SWR values considerably higher than 2:1 when using feed lines having low attenuation. It seems that there aren't many reasons for needing a low SWR on amateur feed lines at HF, if an antenna tuner is used when required to obtain a satisfactory impedance match for the transmitter. (In fact, now would be an appropriate time to review Statements 11 through 17 in Chapter 2.) So let's see how we can briefly develop realistic SWR limits in relation to the attenuation values found in practical feed lines. Here are a few time-tested rules of thumb to use as guidelines.

1) When operation is near the dipole-resonant frequency, either 50- or 75-ohm feed line may be used equally well. Depending on the height above ground, the antenna terminal resistance at resonance

will fall somewhere between 40 and 80 ohms, so the resulting mismatch with either line impedance is so small as to be inconsequential, despite arguments to the contrary from those who are still afflicted with low-viswarmania. However, to obtain accurate readings with an SWR indicator, the impedance of the indicator must be compatible with the impedance of the line on which it's being used.

2) An impedance-matching device placed anywhere on the feed line between a mismatched load and the transmitter compensates for the mismatch at the load, with the resulting effect that a match now exists everywhere on the line (*Ref 17, p 243*). In other words, if a mismatched load impedance, $Z_L = R + jX$, is conjugately matched anywhere on the line, the reflection generated by the complementary mismatch at the matching point causes the impedance looking into the termination end of the line to change from Z_C to $Z = R - jX$. (*See Chapter 4.*)

3) Now let's take advantage of the increased useable bandwidth immediately available to us simply, using our knowledge that nothing magical or miraculous happens in "bringing the SWR down" to 1.0. When we use coaxial line to center-feed a dipole, the operation is usually for one amateur band only. One exception is that satisfactory operation may be obtained on 15 meters with a 40-meter dipole. However, we now have the freedom *to operate anywhere within the entire band*, letting the SWR climb to whatever value it should as the antenna terminal impedance changes with frequency (but still staying within limits that are defined presently). To minimize the increase in mismatch and SWR resulting from the frequency excursion to either end of the band, the dipole should be cut to resonate near the center of the band. On the 75-80 meter band, where the percentage of fre-

frequency excursion is the greatest, the mismatch at the ends of the band will be somewhat less severe with a 75-ohm feed line than with a 50-ohm line. Of the smaller sized coax, RG-59 is preferred over RG-58, because the combination of the somewhat lower maximum SWR and lower matched-line attenuation with the RG-59 permits either a greater frequency excursion away from the self-resonant frequency of the dipole, or a longer line for the same amount of loss. Of the larger size cable, either RG-8 or RG-11 gives nearly equal results, because the matched-line attenuation of the RG-11 is a little greater than with RG-8, thus offsetting the gain resulting from its lower maximum SWR. However, the lower attenuation of the larger cables permits either a greater frequency excursion, or a longer line for the same loss than with the smaller cables, irrespective of their relative power-handling capabilities.

4) The smallest reduction in power that can just barely be detected as a change in level at the receiving station with the AGC disabled is 1 dB. So, to find the SWR that reduces the radiated power by 1 dB, we first use Fig 6-2 to find the attenuation per hundred feet of the correct feed-line type at the desired operating frequency. Then apply the correction for the length of the actual feed line to be used to determine the attenuation, α , of the line. Now go to Fig 6-1 and find the α -loss curve which most nearly corresponds to the value of the feed-line attenuation. Starting where that loss curve crosses the SWR = 1.0 line, follow the curve to the right until 1.0 dB additional loss is indicated on the dB scale on the right side of the graph. Read the SWR at this point from the scale on the bottom. This is the SWR which will reduce the radiated power by the “just barely noticeable” amount at the receiving station, compared

with the signal that would be received if the line had been perfectly matched at the load. More exact data is presented later, but Table 6-1 shows SWR values to be expected for a dipole at the ends of the bands when the dipole is cut for resonance at the center of the band. Applying these data to Fig 6-1 readily shows that it requires feed lines of lengths substantially longer than the average to lose enough additional power from SWR ever to be noticed at the receiving station. In other words, a full 1 dB of additional loss will seldom be encountered, and therefore, no “pileup punch” will be sacrificed in obtaining the increased operating-bandwidth flexibility.

5) At an SWR of around 4:1, the additional loss because of SWR just equals the perfectly matched line attenuation for any line. Thus, in effect, a 4:1 SWR multiplies the matched line loss by a factor of two. As an example, this statement means that the power lost in 350 feet of RG-8, or in 174 feet of RG-59, at 4.0 MHz with an SWR of 4:1, will have a “just barely noticeable” difference compared to a line having no attenuation loss whatever! This is because these lines each have a matched-line attenuation of 1.0 dB. However, also note that this “just barely noticeable” difference could not be noticed when the signal is well up on the S meter, because on many receivers it takes a *6 dB change in signal level for a change of one S unit*.

6) The SWR on the feed line may be monitored to determine that the SWR is within the limit based on the line attenuation, by placing the SWR indicator between the line-matching network and the feed-line input terminals. But remember, *the SWR remains on the line even after the matching network has been properly adjusted*. The match between the transmitter and the matching network may be

Table 6-1 — Approximate SWR Values at Bands Edges for Wire Dipoles Cut to Resonate at Band Center

<i>Frequency, MHz</i>	<i>Max. SWR Value</i>
3.5 to 4.0	5:1 or 6:1 (50-ohm line)
3.5 to 4.0	4:1 or 4.5:1 (75-ohm line)
7.0 to 7.3	2.5:1
14.0 to 14.35	2.0:1
21.0 to 21.45	2.0:1
28.0 to 30.0	3.0:1

monitored with the SWR indicator placed between the transmitter and the network. The network is properly adjusted when the forward power is maximum and the power reflected from the network is zero. If the forward power readings are the same as those obtained with a dummy load, and the reflected power reading is zero in both cases, the input impedance of the network is the same as the impedance of the dummy load. If the SWR indicator shows some power being reflected from the matching network and the transmitter still loads properly, obtaining further reduction of the reflected power is probably unimportant. This indication of reflected power is not showing an “SWR,” but only the degree of impedance mismatch remaining between the transmitter and the input of the network. If insufficient TVI rejection is obtained with the line-matching network alone, a conventional TVI filter may be used between the transmitter and the matching network with the same degree of effectiveness as when used in a line that is matched at the load.

As I have now shown, any required matching can be performed at the input to the feed line, instead of at the terminating load such as an antenna. Therefore, no SWR bandwidth limit for amateur use (such as the commonly used, low arbitrary value of 2:1) is realistic unless it is based on the attenuation of the specified feed-line installation and the amount of total attenuation allowed. The arbitrary

2:1 SWR limit came into existence because the matching range of most amateur transmitters was thought to be limited to 2:1 by design, with economics being considered more important than operational flexibility. But simple line-matching networks as described in the bibliography references can extend the inherent matching range of the transmitters to accommodate values of load impedance far beyond the limits defined by a 2:1 SWR. If it were not for the cost and space factors, line-matching networks could be built into modern transmitters, giving us amateurs back the matching range we were accustomed to having with the old swinging-link method of coupling. We weren’t as conscious of SWR before the pi-net coupling replaced the swinging link. Matching at the line input in those days involved basically a simple adjustment of the adjustable link position to achieve the proper degree of coupling to match the resistive component of the load, and retuning the plate-tank capacitor to cancel the reflected reactance. Using this technique, we often loaded our transmitters into lines with high SWR values without even knowing about the SWR. But with the appearance of the SWR indicators after the departure of the link, we “discovered” SWR and then learned how to misinterpret the meaning of the SWR readings.

In conclusion, if the feed-line loss is within your acceptable limits at a given SWR level, determined from consulting Figs 6-1 and 6-2, and if the transmitter

can be adjusted to load and tune properly (either with or without an additional line-matching network), operate, and don't worry about the SWR, because you are now using realistic SWR for the *right reasons!*

When the series of articles, "Another Look at Reflections," originally appeared in *QST*, reader response was excellent. But some readers told the author, "Your story is interesting, but you'll never convince me that I won't get out better with a perfect 1.0 SWR." Now I remind any reader who still entertains any skepticism of these entire proceedings concerning SWR that the information presented here is not simply a recitation of my own ideas or opinions, but has been taken directly from the professional scientific and engineering literature (note the extensive bibliography in Chapter 24), and paraphrased specifically for the radio amateur with great care not to change the meaning. Moreover, in striking contrast to the many differing opinions heard on the subject during amateur discussions, there *are*

no such differing opinions among the professional sources, because among the professionals (including textbook authors) the principles involved are completely understood and are based on true, proven scientific facts which are not subject to divergent opinions such as we find in politics and religion.

Apparently many have forgotten that this story was told for the amateur in *QST* no less than twice prior to the initial appearance of this series, by two well-known experts in this subject area. They are the late George Grammer, W1DF, former engineer and Technical Editor of *QST*, and Dr. Yardley Beers, W0JF, formerly a professor of physics, Chief of the Radio Standards Physics Division, National Bureau of Standards (now the *NTIA*), and Senior Scientist, Quantum Electronics Division of the National Bureau of Standards. Their illuminating contributions, listed as Refs 3 through 6, 16, and 22 of the bibliography should be reviewed, even if it means a trip to the library — the trip will be very rewarding.