ANOTHER LOOK AT REFLECTIONS

By: M. Walter Maxwell, W2DU/W8HKK

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All references, as well as the bibliography for this series of articles appear at the end of this series.

Compiled with the help of Tom Hammond, NØSS 06/99

This article is the first in a series of QST "Extras" discussing the various facets of the operation of rf transmission lines - transmission-line theory, if you will. Close attention and study of the material, rather than a casual reading, will be rewarding for those interested in gaining a further knowledge of the subject. Some background in complex algebra will aid the reader in understanding the presentation, although the discussions hinge on simple algebraic expressions and manipulations of the Smith chart. In later parts of the series, the author, M. Walter Maxwell, uses a method of presentation which is much more vivid than the conventional way of simply showing vector combinations for a complex load.

Licensed as W8KHK in 1933, Mr. Maxwell's professional antenna experience includes participation from 1940 to 1944 in building antenna farms at FCC monitoring stations in Hawaii and at Allegan, Michigan. With a BS degree from Central Michigan University, he has been an engineer with the RCA Corporation since 1949, and a charter member of RCA's Astro-Electronics Division. Since 1960 he has been in charge of the antenna laboratory and test range at RCA's Space Center, Princeton, NJ. More than 30 earth-orbiting spacecraft have antenna systems designed solely by Mr. Maxwell, including Echo I and all Tiros-ESSA weather satellites. There are many others to which he contributed design assistance. He assisted in the design of Apollo's lunar-rover (moon buggy) earth-link antenna, set up its test-range facilities, and performed many of its pattern, gain, and impedance-matching measurements. He also engineered the prelaunch spacecraft-checkout (ground station) antenna systems at Cape Kennedy for the Tiros and Relay projects, and had complete engineering responsibility for the rf portion (receivers, transmitters, and antennas) of the five ground-station complexes used in Project SCORE, the orbiting Atlas which broadcast President Eisenhower's "Christmas Message from Space" in December, 1958.

Even though Mr. Maxwell's profession deals with far-out space communications, he is very much a down-to-earth amateur. Now licensed as W2DU, he still holds his original call, W8KHK, as well. In addition, he is the trustee of K2BSA, the station of the National Headquarters Radio Club, Boy Scouts of America.

Another Look at Reflections

Part 1 - Too Low a VSWR Can Kill You

BY M. WALTER MAXWELL,* W2DU/W8KHK

JUDGING BY WHAT we hear on the air, nearly everyone is looking for a VSWR of one-to-one. Question why, and the answer may be, "I'm not getting out on this frequency because my SWR is 2.5:1. There's too much power coming back and not enough getting into the antenna," or, "If I feed a line having that much SWR, the reflected power flowing back into the amplifier will burn it up," or still, "I don't want my feed line to radiate." Any of these answers shows misunderstanding of reflection mechanics, and are symptomatic of the current state of education on this subject. Rational and creative thinking toward antenna and feed-line design practice has been absent for a long time, having been replaced with an unscientific and thought-inhibiting attitude, as in the days before Copernicus persuaded the multitudes that the universe did not revolve around the earth. This situation originated with the introduction of coaxial transmission lines for amateur use around the time we got back on the air after World War II, and has gained momentum since SWR indicators appeared on the scene and since the loading capacitor of the

* Engineer. Chief of Space Center Antenna Laborator and Test Range, Astro-Electronics Division, RCA Corporation, Princeton, N.J.

pi-net tank replaced the swinging link as an output-coupling control. We are in this state because so much misleading information has been, and is still being published concerning behavior of antennas which are not self-resonant, feed-line performance in the presence of reflections when mismatched to the antenna, and especially the meaning and interpretation of the VSWR data.

Articles containing explicitly erroneous information and distorted concepts find their way into print, become gospel, and continue to be propagated with chain-letter effectiveness. These include such gems of intuitive logic as (1) always requiring a perfect match between the feed line and antenna; (2) evaluating antenna performance or radiating efficiency only on the basis of feed line SWR - the lower the better; (3) pruning a dipole to exact resonance at the operating (single) frequency and feeding with an exact multiple of a half-wavelength coax - no other length will do; (4) adjusting the height - perhaps just lowering the ends into an inverted V - to make the resistive component equal to the line impedance; or (5) subtracting percent reflected power from 100 to determine usable percentage of transmitter output power (nomographs have even been published for this erroneous method). As a result of these misdirected concepts, we have been conditioned to avoid any mismatch and reflection like the plague. One-to-one all the way! Sound exaggerated? Not if the readers' receivers are tuning the same amateur bands as the author's! In the current vernacular, one could say we have a severe SWR hangup! In many instances, from the viewpoint of good engineering, this hangup is inducing us to concentrate our impedance-matching efforts at the wrong end of the transmission line. (See reference 6.)[1]

It is ironic that we should be in this situation, because the amateur is generally quite practical when it comes to following theoretical considerations. In this case we have been following the perfect-match theory down the narrow path because many of the aforementioned articles have misled us to believe that all reflected power is lost, with never an inkling that, properly controlled, reflections can be turned to our advantage in obtaining increased bandwidth which we are presently throwing away.

That so much misinformation gained foothold is surprising in view of the correct teachings of the ARRL *Handbook (ref. 1)*, the ARRL *Antenna Book (ref. 2)*, the works of Grammer *(refs. 3, 4, and 5)*, Goodman *(ref. 7)*, McCoy *(refs. 8 through 13 and 41)*, Drumeller *(ref. 14)*, Smith *(ref. 15)*, and especially two articles addressed to a subject nearly identical to this one, by Grammer *(ref. 6)* and Beers *(ref 16)*. One objective of this article, therefore, is to identify some of the many erroneous notions concerning reflection principles with sufficient clarity to challenge the reader to question his own position on the subject. Once we correctly understand mismatch and reflections we can obtain improvement in operational antenna flexibility, similar to going VFO after being rockbound with a single crystal. And when we discover how little we gain by achieving a low SWR on the feed line we will avoid unnecessary and time-consuming antenna modifications, often involving hazardous climbing and precarious operations on a roof or tower, which can result in injuries or even death. Let's kill SWR misconceptions - not ham operators!

Open-Wire Versus Coax Feed Lines

The theory behind the transmission of power through a feed line with minimum loss by eliminating all reflections - terminating the line with a perfect match - is equally valid, of course, for open-wire and coaxial lines. But in the days of open wire, prior to our widespread use of coax, it was tempered with practical considerations. Open-wire line was, and still is, used with high VSWR to obtain tremendous antenna bandwidths with high efficiency. This is because all power reflected from the line/antenna mismatch which reaches the input source is conserved, not dissipated, and is returned to the antenna by the "antenna tuner" (transmatch) at the line input. But, although the loss from reflections and high SWR is not zero, this additional loss is negligible because of the low attenuation of open-wire lines. If the line were lossless (zero attenuation) no loss whatever would result because of reflections.

The error in our thinking, that standing waves on coaxial line must always be completely eliminated, originated quite naturally, because the permissible reflection and SWR limits *are lower* than in open wire. When using coax for truly single-frequency operation it makes sense to match the load and line to the degree economically feasible. But it makes no sense to match *at the load* in many amateur applications where we are chiefly interested in operating over a band of frequencies; single-frequency operators we are not, except as our misguided concern over increasing SWR restricts our departure from the antenna resonant frequency.

Many authors are responsible for perpetuating the unscientific and erroneous viewpoint that the coax-fed antenna must be operated at its self-resonant frequency, by continually overemphasizing the necessity for its being matched to the line within some arbitrary, low SWR value to preserve transmission efficiency, and by implying that efficiency equals 100 minus percent reflected power. The viewpoint is unscientific because it neglects the most important factor in the equation for determining efficiency - line attenuation. And it is also erroneous because efficiency does not relate to reflected power by simple subtraction. Setting an SWR limit alone for this purpose is meaningless, because the amount of reflected power actually lost is not dependent on SWR alone. The attenuation factor for the specific feed line must also be included because *the only reflected power lost is the amount dissipated in the line because of attenuation - the remainder returns to the load.* These authors have so wrongly conditioned us concerning what happens to the reflected power that many of us have overlooked the correct approach to the subject. It is clearly presented in both the ARRL *Handbook* and the *Antenna Book* that transmission efficiency is a *two*-variable function of both *mismatch* and *line attenuation*. With this knowledge and by using a graph of the function appearing in these ARRL books, presented here as Fig. 1, the amateur can determine how much efficiency he will lose for a given SWR with the attenuation factor of each specific feed line. He can then decide for himself what the realistic SWR limit should be.

Unimportance of Low SWR Values

In our efforts to obtain low feed-line SWRs of 1.1, 1.2, or even 1.5 to 1, we have gone far past the diminishing-returns point with respect to efficient power transfer, *even for single-frequency operation*, for the same reason one would not install a No. 4 or 6 wire in a house wiring run where No. 12 is sufficient. Reference to the basic transmission-line equations, which have always been readily available in engineering texts and handbooks *(refs. 1, 2, 17, 18, 19, 33),* will verify this analogy in addition to making it clearly apparent that authors who simply insist on low SWR or find 1.5 or 2 to 1 objectionably high have failed to comprehend the true relationship between reflected and dissipated power. From the viewpoint of amateur communications, it can be shown mathematically and easily verified in practice that the difference in power transferred through *any* coaxial line with an SWR of 2 to 1 is imperceptible compared to having a perfectly matched 1.0-to-1 termination, no matter what the length or attenuation of the line, and that many typical coaxial feed lines that we use in the hf bands with an SWR of 3 or 4, and often as high as 5 to 1, have an equally imperceptible difference. When feed-line attenuation is low, allowing such higher values of SWR permits operating over reasonably wide frequency excursions from the self-resonant frequency of the antenna with the imperceptible power loss just described, in spite of the popular impression to the contrary.

The relative unimportance of low SWR when feed-line attenuation is low is demonstrated rather vividly in the following two examples of spacecraft antenna applications. First, in the Tiros-ESSA-Itos-APT weather satellites, of which the entire multifrequency antenna-systems design was the work of the author, the dipole terminal impedance at the beacon-telemetry frequency (108 MHz in early models) was 150 - *j*100 ohms, for a VSWR of 4.4, reflected power 40 percent. Matching was performed *at the line input*, where it was fed by a 30 *milli*watt telemetry transmitter. (We can't afford much power loss here!) The feed-line and matching-network attenuation was 0.2 dB, and the additional loss from SWR on the feed line was 0.24 dB (5.4 percent), for a total loss of 0.44 dB (9.6 percent). On the prevalent but erroneous assumption that all reflected power (40 percent) is lost, only 18.1 *milli*watts would reach the antenna, and efficiency, determined on the same erroneous basis, would be only 60 percent. But 27.1 milliwatts were measured; of the 2.9 milliwatts lost in total attenuation, only 1.6 milliwatts of it was from the 4.4:1 VSWR. So the real efficiency would have been 95.5 percent if perfectly matched at the load, but reduces to 90.4 percent by allowing the

4.4 VSWR to remain on the feed line. Second, in the Navy Navigational Satellite (NAVSAT), used for precise position indications for ships at sea, the antenna terminal impedance at 150 MHz is 10.5 - *j*48 ohms, for a VSWR of 9.8, reflected power 66 percent. Also matched at the line *input*, flat-line attenuation is 0.25 dB, and the additional loss from SWR is 0.9 dB, for a total system loss of 1.15 dB, approximately 1/6 of an S unit. This is an insignificant amount of loss for this situation, even in a space environment where power is at a premium. Why did we match at the line input? Because critical interrelated electrical, mechanical and thermal design problems made it impractical to match at the load. Line-input matching provided a simple solution by permitting the matching elements to be moved to a noncritical location. This design freedom afforded tremendous saving in engineering effort with negligible compromise in rf efficiency, in spite of SWR levels many amateurs would consider unthinkable.

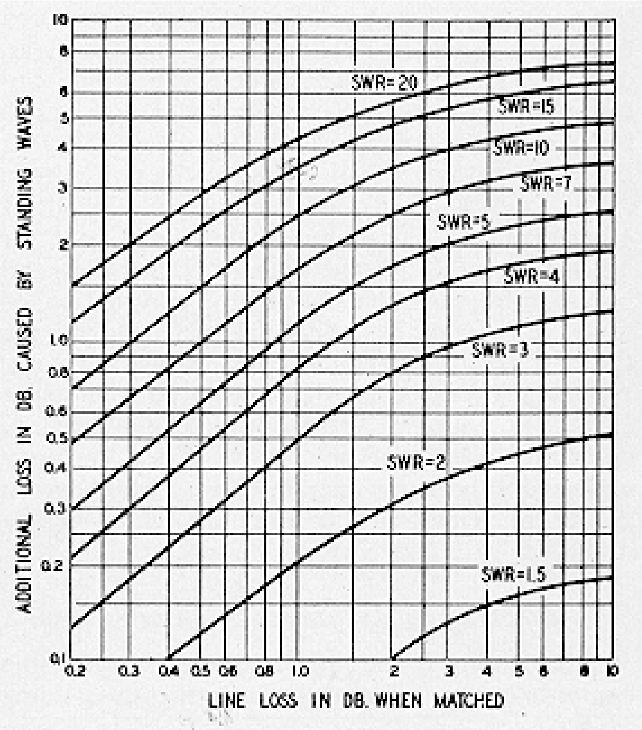


Fig. 1 — Increase in line loss because of standing waves (SWR value at the load). To determine the total loss in decibels in a line having an SWR greater than 1, first determine the loss for the particular type of line, length and frequency, on the assumption that the line is perfectly matched. Locate this point on the horizontal axis and move up to the curve corresponding to the actual SWR. The corresponding value on the vertical axis gives the additional loss in decibels caused by the standing waves.

Another factor which contributes to misunder-standing is the confusion between two distinct, line-usage conditions - one of constant incident *voltage*, and the other of constant input power *(for relative amplitudes, see ref. 19, Fig. 1.3, page 6, and Fig. 3.6, page 29).* Laboratory and experimental work often requires holding incident voltage constant with variation in loading. A constant-voltage source is usually obtained for this purpose by inserting a pad having 15 to 20 dB attenuation between the generator and the line to absorb the reflected power, preventing it from reaching the generator where it would alter the line coupling and cause the generator output voltage to vary. Because of the absorption of the pad, the generator sees a nearly perfect match for all load conditions and all reflected power is lost - but these are laboratory control conditions required to obtain valid test data.

When we amateurs make a change that alters line loading, which in turn alters the transmitter-to-line coupling because of returning reflected power, we can readjust the coupling, returning the line-input power *(not forward power)* to its previous value regardless of the reflected power value.[2] We amateurs can adjust coupling for changes in loading - in the laboratory this is not convenient. We amateurs use low-attenuation lines to conserve reflected power - laboratory setups insert attenuation to *dissipate* it. Confusion over these distinctions has helped perpetuate the erroneous "lost reflected power" concept.

As a result of these various misunderstandings, many amateurs never even wonder whether there are any benefits to be gained by not matching at the line-antenna junction. Many now even shun the use of open-wire lines (not the Old Timers), completely missing the joy of a QSY to the opposite end of the band with only a simple change in transmatch tuning, because the fear of reflections engendered by the exaggerated application of the theory to coax has crept into the thinking concerning *any* form of mismatched connection. Adding still further to the confusion is the old-wives' tale that the reflected power is dissipated in the transmitter, causing tube and tank-coil heating and all kinds of other damage. This myth developed out of ignorance of the true mechanics of reflections and became the easy, but fallacious, explanation of what seems to be abnormal behavior in the transmitter when feeding a line with reflections. What really happens at the transmitter is simply a change in coupling, which will be explained in detail in a section to follow. Then we may understand how to operate with absolutely no danger of damaging the amplifier while feeding into a line with high SWR.[3]

Engineering an Amateur Antenna System

Engineering is the process of making workable compromises in design goals where theories guiding different aspects of the design are in conflict, making it impossible to optimize all the goals. *Good* engineering is simply recognizing the correct choices in the compromises and relaxing the right goals, as in the spacecraft antenna design mentioned earlier. We amateurs spend many hours budding and pruning antenna systems. Wouldn't it be worthwhile spending some of that time learning how to engineer the design in order to make correct trade-off decisions among related factors instead of letting King VSWR dictate the design?

FIRST, we need to improve our knowledge of reflection mechanics and transmission-line propagation to understand…

1) why reflected power by itself is an unimportant factor in determining how efficiently power is being delivered to the antenna.

2) the effect of line attenuation (to discover why it is the KEY factor which will tell us when and how much to be concerned with reflected power and when to ignore it).

3) why *all* power fed into the line, minus the amount lost in line attenuation, is absorbed in the load *regardless of the mismatch at the antenna terminals*.

4) why reflection loss (mismatch loss) is canceled at the line input by reflection gain *(ref. 19, p. 36, and ref. 25, part 2, p. 33).*

5) why a low SWR reading by itself is no more a guarantee that power is being radiated efficiently than a high SWR reading guarantees it is being wasted.

6) why SWR is not the culprit in transmitter loading problems - why the real culprit is the change in line-input impedance resulting from the SWR, and why we have complete control over the impedance without necessarily being concerned with the SWR.

7) the importance of thinking in terms *of resistive and reactive* components of impedance instead of SWR alone, and why SWR by itself is ambiguous, especially from the viewpoint of the selection and adjustment of coupling and matching circuitry.

SECOND, we need to become aware that with moderate lengths of low-loss coax, such as we commonly use for feed lines, loss of power because of reflected power in the hf bands can be insignificant, no matter how high the SWR. For example, if the line SWR is 3, 4, or even 5 to 1 and the attenuation is low enough to ignore the reflected power, *reducing the SWR will yield no significant improvement in radiated power because all the power being fed into the line is already being absorbed in the load.* This point has especial significance for center-loaded mobile whips, because of the extremely low attenuation of the short feed line.

THIRD, we should become more familiar with the universally known, predictable behavior of off-resonance antenna-terminal *impedance and its correlation with* SWR. This knowledge provides a *scientific* basis for evaluating SWR-indicator readings in determining whether the behavior of our system is normal or abnormal, instead of blindly accepting low SWR as good, or rejecting high SWR as bad. The following two examples emphasize the importance of this point by showing how easily one may be misled by a low SWR reading:

1) A ground system having 100 properly installed radials has negligible loss resistance *(ref. 20).* Many a-m broadcast stations use 240 radials, while the FCC requires a minimum of 120. With such a ground system the terminal impedance of the average quarter-wave vertical is 36.5 + *j*22 ohms, and approximately 32 ohms when shortened to resonance. When fed with a 50-ohm line, the SWR at resonance will be close to 1.6, rising *predictably* on either side of resonance. But a 15-radial ground system will have approximately 16 ohms of ground-loss resistance with this antenna. If we remove a few radials at a time from the 100-radial dial system, the increasing ground resistance, added to the radiation resistance, increases the total line-terminating resistance. The terminating resistance comes closer and closer to 50 ohms, reducing the SWR. When enough radials have been removed for the loss resistance to reach 18 ohms,

the terminating resistance will be 50 ohms for a *perfect one-to-one match!* But while the SWR went down, so did the radiated power, because now the power is dividing between 32 ohms of radiation resistance and 18 ohms of ground resistance!

Ground systems having from two to four radials may have a loss resistance as high as 30 to 36 ohms, so now the resonant-frequency SWR will be around 1.4 or 1.5. But instead of rising from this value, as it should at frequencies away from resonance, the ground loss holds the off-resonant SWR to low values. The low SWR simply indicates that the line is well matched, but it offers no clue that approximately half the power is heating the ground.

2) Some amateurs who employ a one-to-one balun believe that "one-to-one" means it will provide a one-to-one match between the feed line and the antenna. This is a serious error because "one-to-one" only specifies the output-to-input impedance *ratio* - no matter what impedance terminates the output, the same value is seen at the input. Nevertheless, these amateurs are convinced the baluns are "matching," because the SWR sometimes goes down dramatically when the balun is inserted. Often with a balun the SWR is less than 2:1 over the entire 75-80 meter band, where somewhat over 5:1 is normal at the band ends.

Off-resonance SWR is reduced here because the ferrite core of the balun saturates while attempting to handle the reactive current, which now exceeds the maximum core-current level. Thus, the full excursion of the reactive component of antenna impedance is prevented from appearing at the balun input. All power above the saturation level is lost in heating the balun, while the low SWR is deceiving the unsuspecting amateur.

The true SWR will be unchanged by a 1:1 balun with a core capable of handling the current without saturating (if it has no significant leakage reactance).[4] However, the SWR indicator may not show the true SWR without the balun if antenna current on the outside of the coax is present at the SWR meter *(ref. 36).*

So it is important to know approximately what SWR to expect - if it is low, determine whether it should be. Don't assume that a low SWR indicates success, or guarantees a great system! Be especially suspicious if the SWR remains low or relatively constant over a moderate frequency range, unless specific broad- banding steps have been performed on the radiating system. This knowledge is elementary and routine for the antenna design engineer, but too little information in this area has been available for the amateur, considering the degree of his involvement with antennas. While antenna-terminal impedance behavior with frequency is shown in the *ARRL Antenna Book (ref. 2, Fig. 2-7)*, correlation of the impedance change with SWR will be covered in detail later, to enable us to predict normal SWR, within limits, with a nonresonant antenna terminating the feed line.

FOURTH, we need to reexamine the use of open-wire lines as tuned lines *(refs. 3, part 3, p. 20; 10; and 21, p. 23)*, to discover that the principles used there are exactly what we have been discussing. Remember, with tuned lines we completely ignore the mismatch at the antenna end, and compensate for the mismatch with the tuner at the input end, over the entire frequency range of the band. The SWR may run as high as 10, 15, or even 20 to 1, but the power reflected from the mismatch is re-reflected back to the antenna by the tuner. Tuning for maximum line current simply adjusts the phase of the reflected wave to rereflect down the line in phase with the forward wave, again reaching the antenna. Thus the reflection loss from the mismatch is canceled by the reflection gain of the tuner.

Many of us amateurs know from age-old practice that a 600-ohm line made of two No. 12 wires on six-inch spacing would work every time. We had little incentive to learn how they worked - why they transferred power *efficiently* with such high reflected power and SWR, or that adjusting the reflected-wave phase to rereflect in phase with the forward wave was just another way of viewing the reactance cancellation required to obtain maximum line and antenna current. Hence, our misunderstanding of the similarity between open-wire and coaxial-line operation with mismatched loads. The principle is the same in both, only the degree is different. In other words, for many applications, *coax can be used as a tuned line in precisely the same manner as open wire.* The spacecraft systems mentioned earlier are examples.

Thus, coax connected directly into the antenna may be operated with substantial mismatch. In this case, the SWR limits while operating away from the self-resonant frequency of the radiator are determined entirely by power lost because of line attenuation. Voltage breakdown and current heating should not be a problem at our legal power limit with RG-8 or -11/U, or with RG-58 or -59/U at lower powers. because voltage at an SWR maximum is only $\sqrt{SWR}$ times the matched value. The line-input impedance will no longer be 50 ohms, but depending on the magnitude of the mismatch and length of the cable, we may determine whether the output tank of the transmitter has sufficient impedance-matching range (surprisingly high in some rigs, none in others) to permit feeding the line directly *(ref. 4, part III),* or whether an intermediate matching device (transmatch, or other type of tuner *(refs. 9-12, incl., 22)* will be required to adjust for correct coupling between the line and the transmitter. (Balun and filter use will be discussed later.) The important point we are emphasizing is that, within the limits mentioned, *all required matching may be transferred back to the operating position instead of forcing the match to occur at the antenna feed point - without suffering any* SIGNIFICANT *loss in radiated power.* The use of this technique, which may come as a surprise to many, does not contradict any theory. It is actually an embodiment of the fundamental principle of network theory

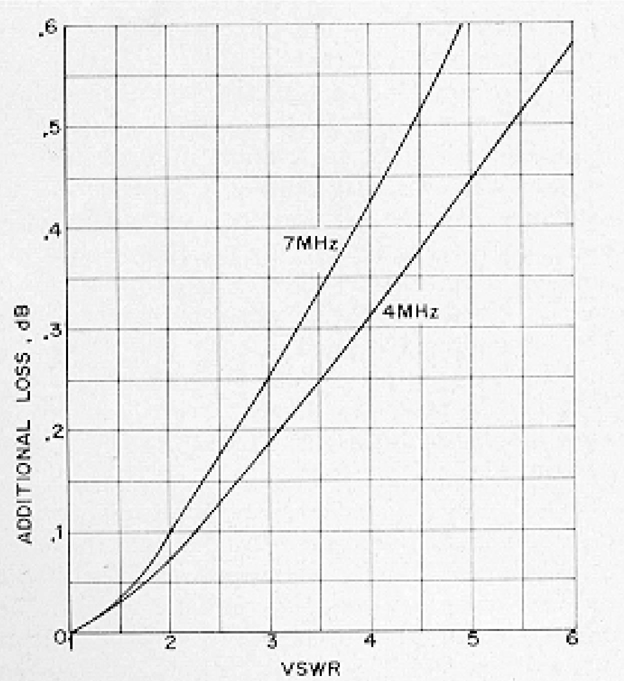


Fig. 2 — Effect of standing-wave ratio on line loss at 4 and 7 MHz. The ordinates give the *additional* loss in decibels over those for a perfectly matched 100-foot length of RG-8/U line for the SWR values shown on the horizontal scale.

called *conjugate matching, (refs. 17, p. 243; 19, p. 38; 35, p. 49)* which is the basis for all antenna tuner, or transmatch, operation with either open-wire or coaxial lines.

After learning of the benefits obtained with line-input, or conjugate matching in the two spacecraft examples described earlier, it is interesting to compare the results using this same input matching technique in typical 80- and 40-meter situations. Eighty meters is the widest amateur band in terms of percent of center frequency and thus suffers the greatest SWR increase with frequency excursion to the band ends. A dipole cut for resonance at 3.75 MHz will yield an SWR in a 50-ohm feed line somewhat above 5:1 at both 3.5 and 4.0 MHz. As shown in Fig. 2, in a 100-foot length of nonfoam RG-8/U, an SWR of 5:1 adds only 0.46 dB loss to the matched (i.e., flat line) loss of 0.32 dB at 4.0 MHz.. So out almost to the band ends, less than 1/12 of an S unit is lost because of the SWR, an imperceptible amount. This further verifies the principle and proves that full-band, coax-fed dipole operation on 80 meters also is practical. Even with the high SWR at the band ends, the loss cannot be distinguished from what it would have been had the SWR been a perfect one-to-one! At 40 meters, with the dipole resonated at 7.15 MHz, something is amiss if the SWR exceeds 2.5 at the band ends. And from Fig. 2 it may be seen that this SWR adds only 0.18 dB to the matched loss, which at 7 MHz is 0.44 dB for 100 feet of RG-8/U coax.

Nonreflective Load Versus Conjugate Matching

Now is perhaps a good time for the reader to contemplate the conflict between the no-reflection perfectly matched load theory and the conjugate match theory. It is amply evident from the standpoint of good engineering that as long as SWR does not exceed the value above which one cannot afford to compromise further power in exchange for improved operating flexibility, the convenience and increased bandwidth afforded by conjugate matching at the line input is obvious.

But it also presents a real challenge to learning more about complex impedance, because the line-input impedance now has resistive *and reactive* components, both of which vary with changes in line length and with frequency in the presence of reflections. Thus, we need to understand *complex* impedance in order to choose and adjust correct conjugate matching circuitry to couple the transmitter to the line, or to adjust the transmitter directly to the line if sufficient matching range is available. Practically all problems encountered while attempting to obtain proper coupling or loading to a line with reflections can be traced simply to not understanding the correlation of line length and relative phase of the incident and reflected waves with the resulting complex impedance seen at the input terminals of the line.

A detailed discussion of reflection mechanics and feed-line propagation will be presented in subsequent installments. Included will be a novel means for explaining impedance transformation along the line in direct relation to incident and reflected waves, which will simplify the understanding of what does and does not happen when a line length is changed, and how to select the correct length for given conditions. The relation of line attenuation to permissible SWR while using conjugate matching techniques, along with details on how to obtain proper coupling and loading of a transmitter to a line for which the input impedance has changed because of reflections, will also be presented.

Part 2 - Countdown for a Journey From Mythology to Reality

Part 1 of this series of articles appeared in *QST* for April, 1973. In that part we saw that obtaining a low SWR is relatively unimportant for an efficient transfer of power when line attenuation is low. Four steps to assist in understanding the operation of lines with reflection were suggested, and the concept of matching the complex impedance at the line input in the presence of reflections, called conjugate matching, was introduced. The paragraphs which follow present for consideration some of the basic principles involving efficient power transfer through any line terminated in a mismatch.

A conjugate match exists throughout the entire system when the internal resistance of the source is made equal to the resistive component of the line-input impedance (or vice versa) and all residual reactance components in the source and line-input impedances are canceled to zero. In this condition the system is resonant. All available power from the source enters the line, and reflections from any terminating mismatch or other line discontinuities are compensated by a complementary reflection obtained by introducing a nondissipative mismatch at the conjugate match point. This nondissipative mismatch is one which if placed in the system by itself, would produce the same magnitude of reflection, or SWR, as is produced by the mismatched line termination. The result is a precise and total rereflection of the arriving reflected wave. Andrew Afford makes a magnificent presentation of this concept *(ref. 39, pages 10-15).* Although it sounds very complicated, this entire set of conditions is automatically fulfilled simply by completing a correct tuning and loading procedure. It matters not whether a transmitter having sufficient matching range feeds the line directly, or whether an external transmatch is used where additional range is required. If the source generator is now replaced by a passive impedance equal to its internal impedance[5] the line can be opened at any point. And looking in either direction, one will see the conjugate of the impedance seen in the opposite direction - whatever R + *j*X value is seen in one direction, R - *j*X is seen in the other.

Contrary to our prevalent, deeply ingrained belief, it is therefore not true that when a transmitter delivers power into a line with reflections, a returning reflected wave always sees the internal generator impedance as a dissipative load and is converted to heat and lost. It can happen under certain conditions of pulse-type transmission; for instance, if the generator is turned off after delivering a single pulse into the line while retaining its internal impedance across the line, the returning pulse wave will be absorbed. But if a conjugate-matched generator is actively supplying, power when the reflected wave returns, the reflected wave encounters total reflection at the conjugate match point and is entirely conserved, because it never sees the generator resistance as a dissipative terminating load. This is because the source and reflected voltages and currents superpose, or add at the match point, just as if the reflected power had been supplied by a separate generator in series with the source. And since the source voltage is generally greater than the reflected, the sum of their voltages yields a net current flow which is always in the forward direction.[6] The reflected power adds to the source power, deriving reflection gain which compensates for the reflection loss suffered at the mismatched termination.

Line Losses

All reflected power reaching the source is returned to the load, as part of the forward or incident wave. The only reflected power lost is because of line attenuation, during its return to the source and once again during its return to the load. The higher the line attenuation, the less reflected power reaches the source to add to the forward power. Thus, the lower the line attenuation, the higher the allowable SWR for a given loss because of SWR. *No* reflected power is lost in a lossless line, *no matter how high the SWR*, because it *all* gets ultimately to the load. This is why open-wire line functions efficiently as a tuned line with any reasonable mismatch value -- its attenuation is almost negligible. Attenuation (being higher in coax) imposes lower limits on the mismatch and may require calculation of the loss penalty for a given SWR. *Both the attenuation and SWR must be quite high* to incur any substantial additional loss over and above the matched-line loss.[7]

Coax has higher rf losses than open wire at hf chiefly because of its lower impedance, causing higher current flow at lower voltage for the same power. This results in higher I^2R loss for the same effective conductor size. (Electric power distribution lines minimize I^2R loss by use of *high* voltage and *low* current). Skin effect increases the loss with rising frequency because of decreased effective conductor size, but only at VHF and higher does the dielectric loss become a substantial contributor to the attenuation factor. From this it is understandable why RG-8/U, especially the foam type with its larger center conductor *(ref. 23)*, will allow higher SWR (more bandwidth) than RG-58/U for the same additional loss penalty. And for any cable, the shorter it is the less loss is added for a given SWR.

A fifth step in improving our understanding of the reflected-power problem is to view the situation objectively, asking yourself, "Have I fallen prey to any of the erroneous teachings? Can I spot the wrong dope when I hear it discussed? Do I understand the principles well enough to convince others of the correct version if the opportunity arises?" Several pertinent short statements follow which may be used as self-test material. They high-light and summarize many reflection-related concepts known to be generally confused among the amateurs. All of the statements are **TRUE**. In the interest of brevity they are not intended to be completely self-explanatory, but sufficient material for obtaining a complete understanding of each point will appear in later installments, or is available in bibliography references included in part 1. Support for nearly every statement can be found in *The ARRL Antenna Book* alone.

True or False?

1) Reflected power does not represent lost power except for an increase in line attenuation over the matched-line attenuation. In a lossless line, *no power is lost because of reflection*. Only when the flat-line attenuation and SWR are both high is there significant power lost from reflection. On all hf bands with low-loss cable, reflected power loss is generally insignificant, though at VHF it becomes significant, and at UHF it is of extreme importance.

2) Reflected power does not flow back into the transmitter and cause dissipation and other damage. Damage blamed on reflections is really caused by improper output-coupling adjustment – not by SWR. Tube overheating is caused by either or both overcoupling and reactive (mistuned) loading. Tank-coil heating and arc-overs result from a rise in loaded *Q* caused by undercoupling. With some manipulation, proper output coupling (indicated by a normal resonant plate-current dip at the correct loading level) can be attained no matter how high the SWR. The transmitter doesn't "see" an SWR at all -- only an impedance resulting from the SWR. And the impedances are matchable without concern for the SWR. This is one of the most important points of confusion at issue.

3) Any effort to reduce an SWR of 2:1 on any coaxial line will be completely wasted from the standpoint of increasing power transfer significantly. *(See Fig. 1, part 1.)*

4) Low SWR is *not* proof of a good-quality antenna system or that it is working efficiently. On the contrary, lower than normal SWR exhibited over a frequency range by a straight dipole or a vertical over ground is a clue to trouble in the form of undesired loss resistance. Such resistance can be from poor connections, poor ground system, lossy cable, and so forth.

5) The radiator of an antenna system need not be of self-resonant length for maximum resonant current flow, the feed line need not be of any particular length, and a substantial mismatch at the line-antenna junction will not prevent the radiator from absorbing all real power available at the junction. *(refs. 3, part 3, p. 10; 24.)*

6) If a suitable transmatch cancels all the reactance developed by a nonresonant-length radiator and a random-length feed line which is mismatched at the antenna feedpoint, *the antenna system is resonant*, the mismatch effect is canceled, *maximum current flows in the radiator*, and all the real power available at the feed point is absorbed by the radiator.

7) The majority of tower radiators used in the standard a-m broadcast band (from 540 to 1600

kHz) are of heights which are *not* resonant lengths at the frequency of operation.

8) SWR on the line between the antenna and transmatch is determined *only* by the matching conditions at the load, and is not changed or "brought down" by the matching device. "Low SWR" obtained by using the device indicates only the mismatch remaining between the input impedance of the transmatch and impedance of the line from the transmitter.

9) Adjusting the transmatch for maximum line current creates a perfect mirror termination for the reflected wave, causing it to be totally rereflected on arrival at the input. The tuner provides the proper reactance to cancel the equal but opposite reactance resulting from the amplitude and phase difference between the source and reflected waves at the input. This causes the reflected wave to add *in phase* to the source wave to derive the incident power, which is the sum of the source and reflected power.

10) Total rereflection of the reflected power at the line input is the reason for its not being dissipated in the transmitter, and why it is conserved, rather than lost.

11) With a good "antenna tuner" or transmatch and a well-constructed open-wire feeder, a 130-foot center-fed dipole will not radiate significantly more power on 80 meters than one 80 feet long for the same power fed from the transmitter *(refs. 10; 21; 3, part 3, p. 20: 7, pp. 50 and 126).*

The SWR on the line is determined only by the matching conditions at the load.

12) A dipole cut to be self-resonant at 3.75 MHz and fed with either RG-8/U or RG-11/U coax will not radiate significantly more on 3.75 MHz than on 3.5 or 4.0 MHz with any feeder length up to 150 or 200 feet.

13) With the 3.75 MHz dipole the feed-line SWR will rise to around 5.0 at both 3.5 and 4.0 MHz, thus utilizing the coax as a tuned feeder, but with *insignificant* loss in radiated power *across the entire 80-meter band.*

14) With the use of a transmatch or a simple L-network at the line input, proper coupling between the transmitter and the tuned-coax feeder can be attained over the entire band with *any random coax length*.

15) From the standpoint of line loss because of SWR resulting from the change in quality, of the impedance match between the line and antenna, changing the height of the dipole above ground or lowering the ends of a horizontal dipole to make an inverted-V will have an *insignificant* effect on the amount of power reaching it from the transmitter.

16) As a tuned line at 4.0 MHz, RG-8/U will handle 700 watts CW *continuously*, within ratings, at an SWR of 5:1. With the duty cycle of SSB it is far below maximum ratings at 2 kW PEP. With a 100-foot length, the *total* attenuation (SWR = 5) is just 0.8 dB (0.46 dB because of SWR), which is insignificant in terms of received signal strength.

17) If a line length is critical in order to satisfy a particular matching condition, the same input impedance can be obtained with *any* length of line, shorter or longer, by adding a simple *L* network of only two components: either two capacitors, two inductors, or one of each, determined by the specific impedance change required of it. This statement is pertinent to coiled-up coax in mobiles. *(Refs. 19, pp. 118-128: 24; 30, p. 48; 3 1.)*

18) High SWR in a coaxial transmission line caused by a severe mismatch will *not* produce antenna currents on the line, nor cause the line to radiate *(refs. 32; 2, p. 101).*

19) High SWR in an open-wire line at hf caused by a severe mismatch will *not* produce

antenna currents on the line, nor cause the line to radiate, if the feed currents in each wire are balanced, and if the spacing is small at the wavelength of operation (also true at VHF if sharp bends are avoided.) *(ref 2, pp. 101, 106.)*

20) Both coax and open-wire feed lines may radiate *(ref. 32),* though not at a significant level, by re-radiating energy coupled into the line from the antenna because of asymmetrical positioning with respect to the antenna. The coupled energy results in antenna currents flowing on the *outside* of the outer coax conductor, or *in-phase* currents flowing in the wires of the open-wire line. But this condition has *no* relation to level of the line SWR in either case *(ref. 2, pp. 101, 106).*

21) SWR indicators need not be placed at the feed-line/antenna junction to obtain a more accurate measurement. Within its own accuracy limits, the indicator reads the SWR wherever it is located in the line. The SWR at any other point on the line may be determined by a simple calculation involving only the SWR at the point of measurement, the line attenuation per unit length (available in a later installment), and the distance from the measured point to the point where the SWR is desired.

22) SWR in a feed line *cannot* be adjusted or controlled in any practical manner by varying the line length *(ref. 7, p. 51).*

23) If SWR readings change significantly when moving the bridge a few feet one way or the other in the line, it probably indicates "antenna" current flowing on the *outside* of the coax, or else an unreliable instrument, or both, but it is *not* because the SWR is varying with line length. Some writers insist the bridge must be placed at a half-wave interval from the load to obtain a correct reading. *This is incorrect*. All readings are invalid if they change significantly along the line, even though they may repeat at half-wavelength intervals *(ref. 2, pp. 101, 106, and 132).*

24) Any reactance added to an already resonant (resistive) load of any value for the purpose of compensation to reduce the reflection on the line feeding the load will, instead, only increase or worsen the reflection. It is for this reason, though contrary to the teaching of several writers, that *lowest feed-line SWR occurs at the self-resonant frequency of the radiating element it feeds*, completely independent of feed-line length. Any measurements which contradict this indicate that either the measuring equipment or the technique (or both) are in error.

25) Of the several types of dipoles, such as the thin wire, folded, fan, sleeve, trap, or coaxial, none will radiate more field than another, providing each has insignificant ohmic losses and is fed the same amount of power *(ref. 3, part 3).*

26) If coax at least the size of RG-8/U is used in mobile installations (80 thru 10 meters), any matching required to load the transmitter may be done at the *cable input end* without significant power loss compared to matching at the antenna terminals, and with improvement in operating bandwidth.

27) With center-loaded mobile whips of equal size having no matching arrangement at the input terminals, best radiating efficiency is obtained on models having the lowest measured terminal resistance (highest resonant SWR, model for model). Models having lowest SWR are wasting power in the loading coil, because of either a low value of coil *Q* or excessive distributed coil capacitance, or both.

As was mentioned earlier, all of these statements are true. These examples have been centered around 80-meter operation because bandwidth and dipole length on this band present the maximum SWR problem. Of all the amateur bands, 80M has the largest bandwidth, 13.3 percent of center frequency, as compared to 4.2 percent on 40M, 2.5 percent on 20M, 2.1 percent on 15M, and 5.9 percent on 10M. Having the longest wavelength (excepting 160, of course), 80M poses the greatest problem with respect to physical construction of radiating systems on existing real estate. Some properties just won't permit an entire half wavelength on 80. So these examples should have a special interest for the fellow who wishes to work 80 meters but is forced to use a short antenna. Since the bandwidth and antenna-length problem are really one and the same, the 80-meter examples have maximum practical value. But regardless of which band we select, the principles are the same. Practices recommended at the 80-meter level are also valid on the higher hf bands. Interestingly enough, as we go to the higher bands, where the line losses increase, the percentage bandwidth of the amateur band decreases. This means inherently lower maximum SWR values will be obtained during frequency excursions from the design center to the band ends.

About This Series of Articles

The original idea for writing this paper was born while the writer was listening to and participating in many discussions concerning mismatch and reflections. It soon became apparent that if the twenty-seven true statements above were presented as a true-false test, many amateurs would mark them false. Those discussions also revealed that many experiments were performed in this area for which the results were predestined to futility, because the experimenter misunderstood the principles. In many cases, the experimenter was unaware of both the ultimate futility of his efforts and that the futility was a direct result of his misunderstanding. Then the appearance of the article by Drumeller *(ref. 14)* motivated the desire to do something constructive about the problem. Although the results of his experiment agree with the writer's teachings, Drumeller's preface implies that little is known on the subject of mismatch. This is not true, unless he was referring specifically to us amateurs.

The limited equipment available for amateurs to make precise and complete measurements at rf, as compared to DC, low-frequency AC, and audio, understandably limits the quality of our experiments. But this doesn't excuse us from needing to know some of the fundamental principles of transmission lines -- it actually makes knowledge of the fundamentals all the more necessary, so we can correctly diagnose and logically evaluate the limited measurement data we do obtain. Our experimental efforts can become more productive if we are able to visualize an accurate physical picture of the voltages, currents, and fields, and how they interact on the line.

So the writer initially planned to compile and publish an extensive bibliography of *accurate* references on the subject, references which are readily available to the amateur. But at one of the monthly Colts Neck meetings, a good friend and fellow 3999er, the late John Marsh, W3ZF, convinced the writer that a mere list of references would receive little attention. W3ZF then proceeded to fire real enthusiasm for writing an extensive paper which would not only complement and unify the references, but would also underscore the problem and stimulate interest in studying the subject. He felt strongly that this approach would make a more worthwhile contribution to amateur radio.

John's suggestion was followed, and as the work progressed it became apparent that additional references from professional engineering sources would be valuable in providing access to greater depth of study for the more advanced amateur, and by affording reliable sources for verification of points which have been clouded by controversy among amateurs.

During study of the numerous references for preparation of the manuscript, an interesting and valuable aspect emerged: the *ARRL Antenna Book* already unifies the other references, plus more, because in one handy volume it contains sufficient, well-presented material on every point of importance to the amateur's needs concerning antennas and transmission lines, including the fundamental principles! For the amateur who has no access to the other references, but could manage to acquire just one book, the *Antenna Book* is it! Every amateur who feeds rf into an antenna should have a copy -- and read it! One would be surprised to learn how many antenna engineers have a copy in their own personal libraries.

With the exception of the *ARRL Antenna Book*, the details of the reflection activity in a length of line seem to be somewhat obscure in the amateur literature, so the beginning of the next part of this series presents a simplified treatment of the interaction of the fields, currents, and voltages in a line. Following that, a new vector-type presentation using the Smith chart is given.[8] The presentation enhances the physical picture of standing-wave development, input impedance, and other activity along the line, and forms a basis for an arithmetical proof that the explanation presented is correct. The two-generator concept *(ref. 17, p. 133)* is used as the basis for determining line-input impedance, because this concept permits direct comparison of the incident -- and reflected-wave vector presentation with simple series-equivalent circuits. This reduces the line reflection problem to one of simple Ohm's Law calculations, and at the same time achieves a built-in proof of validity. This concept is simple to grasp, easy to remember, and will be recognized as a powerful tool for assisting the amateur in analyzing any feed-line situation from the viewpoint of matching the transmitter to the *line input* under any SWR condition. It uses simple arithmetic, but it is not short and it will require some study -- there isn't any short cut to understanding reflections on transmission lines.

For those who are unfamiliar with the Smith chart it is easy to learn *(ref. 25, part 1).* The chart will be found to be a valuable tool. It has separate sets of circles for resistance and reactance. To plot a single complex impedance R + *j*X one simply finds the point where the appropriate R and *j*X circles intersect. Knowledge of the Smith chart will be found helpful in understanding and devising matching circuitry; it is actually fun to use for this purpose. Additional references of interest are included in the bibliography.

The author wishes to express appreciation to his many friends who offered suggestions and criticisms, especially to Bob Allen, W8IO; Ken MacLean, W2KKM, the author's engineering supervisor for many years; and colleague Bert Sheffield, W2ANA. Their help has been invaluable. Acknowledgment with thanks is also given to Mr. Phillip H. Smith, ex-1ANB, for his kind permission to use the Smith chart in the Studies which follow.

Part 3 - Going Around in Circles to Get to the Point; Basic Reflection Mechanics

Basic Reflection Mechanics

It is generally well understood that the size or magnitude of a reflection arising from a mismatched line termination is determined by the degree of the mismatch, or how much of the incident-wave power is unabsorbed by the load, expressed as a voltage or current ratio relative to the size of the incident wave.[9] The reflection coefficient, $\overline{\rho}$ (rho),[10] is determined quantitatively from the line and load impedances by the expression

$$\overline{\rho} = \frac{Z_L - Z_C}{Z_L + Z_C} \qquad \text{(Eq. 1)}$$

where Z_L is the complex load impedance, R +*j*X, and Z_C is the characteristic impedance of the transmission line. This shows immediately that ρ = 0 (no reflection) when $Z_L = Z_C$. However, the load must be purely resistive (R + *j*0) for zero reflection, because we are considering only lossless and low-loss lines with a characteristic impedance this is purely resistive. Perhaps somewhat less appreciated than reflection magnitude is reflection *phase*, which is determined by the character of the mismatch, and expressed as an angle relative to the phase of the incident wave. The magnitude ratio, ρ and the relative phase, θ (theta.), together comprise the complex reflection coefficient $\overline{\rho} = \rho\angle\theta$ which tells us all we need to know about the reflection in order to use it in understanding transmission-line propagation and matching techniques. How we use it will be explained in a later section.

An open-circuit (infinite impedance), a short-circuit (zero impedance), or a purely reactive load on a transmission line is incapable of absorbing *any* power from an incident wave and will therefore cause total reflection of both volt and current incident waves. The reflection coefficient magnitude, ρ, at such a load is therefore unity (1.0) for both voltage and current. The step-by-step by which the reflections arise on low-loss lines, both coaxial and open wire, is as follows. The incident electromagnetic wave sees the characteristic impedance, Z_C, of the line as a resistive load as it leaves the generator in its forward travel down the line. Half of the energy is stored in the magnetic field, because of incident current, and the other half is stored in the electric field, because of incident voltage. The voltage and current travel in phase with each other because of the resistive Z_C. On reaching an open circuit the magnetic field collapses, because the current goes to zero. The changing magnetic field produces an electric field equal, in energy to the original magnetic field. The now electric field adds *in phase* to the existing electric field, resulting in a corresponding increase of voltage. at the open circuit to twice the incident-wave voltage. (At this instant a standing wave is developing, because now there is a current minimum and voltage maximum at the open-circuit terminals, where an instant before, current and voltage were constant, all along the line.) The increased voltage now starts the reflected voltage wave traveling in the opposite direction, as if it had been launched by a separate generator at the open-circuit point.[11] Since no energy was absorbed by the open-circuit load, the returning wave will be of the same magnitude as the original incident wave. As the electric field starts its rearward journey it sets up a new magnetic field in opposite phase to the original, and once more the energy will divide equally between the two fields. The new magnetic field causes the current to build up again to the same magnitude as before, *but in the opposite polarity*, to be relaunched into the line as the reflected current wave *(refs. 17. p. 139; 19. p. 4; 35, p. 21; 43).* The total voltage (or current) at the load at any instant is the sum of the voltages (or

[EDITOR'S NOTE: Here the author is not referring to the standing wave ratio which might be measured with an ordinary SWR indicator. Instead he is referring to traveling waves. The difference is discussed again in later text.]

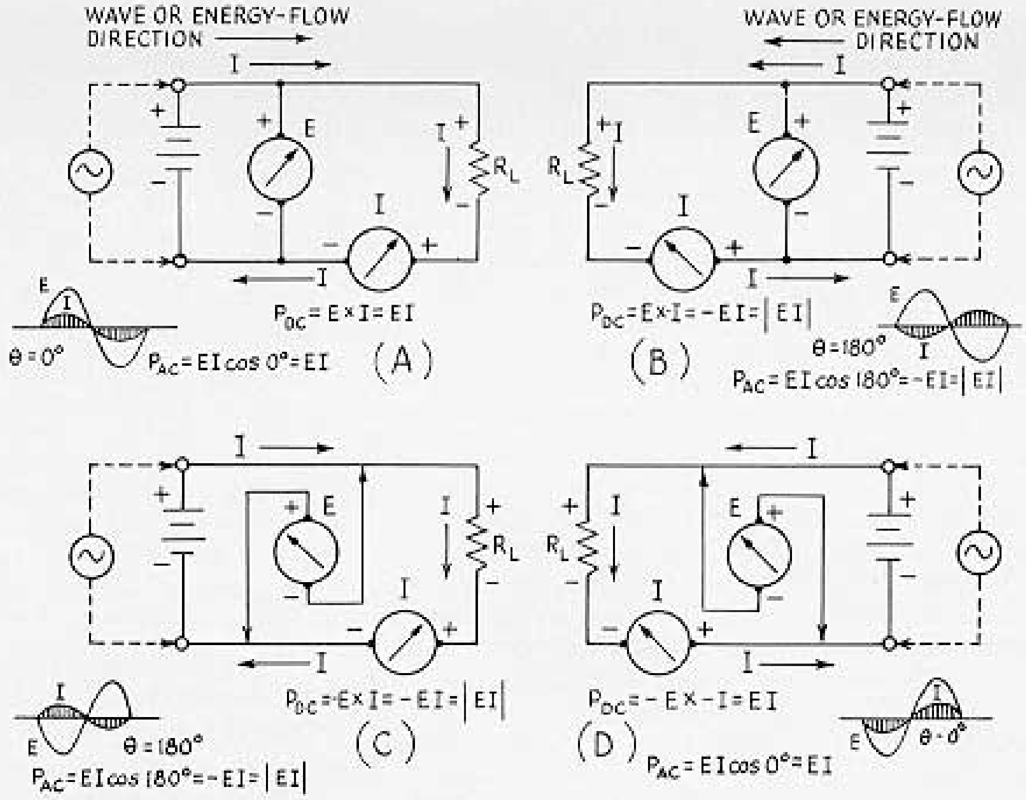


Fig. 3 — Illustrating relationships between current and various combinations of voltage reference and energy-flow direction.

currents) of the incident and reflected waves. Since the two currents add to zero at the open-circuit load, the generation of the reversed polarity reflected current wave is verified. The in-phase reflected voltage wave is similarity because the sum of the two voltages at the load is double the incident voltage. The phase angles, θ, of the reflection coefficients at the open-circuit load are therefore 0 degrees for voltage and 180 degrees for current.

When the load impedance is a short circuit the reflection-generation process is similar to the open-circuit case, except that the electric- and magnetic-field actions and the polarities of the reflected-wave components are reversed. This is expected when we recall that while current goes to zero in an open circuit, voltage must be zero in a short circuit. For the voltage to be zero the incident and reflected voltage waves must cancel one another at the load, thus verifying the reversed polarity of the reflected wave. The corresponding currents add to double the incident value, as the voltages did when the load was an open circuit. The phase angles, θ, of the reflection coefficients at the short-circuit load are therefore 180 degrees for voltage and 0 degrees for current. When the load impedance is a pure capacitance it is equivalent to an additional length of open-circuit line, while a purely inductive load is equivalent to an additional length of short-circuited line.

When the load impedance contains resistance the reflection will be generated in the same manner as with an open- or short-circuit load, but it will be less than total, the amount depending on how much power is absorbed in the resistance. The reflected wave is again generated by the changing electric and magnetic fields at the mismatch point, caused by the change in voltage and current when the incident wave encounters a change in load conditions. Hence the reflection coefficient, ρ, is dependent on the difference between incident-wave voltage on the line and the voltage measured across the load.

No reflection arises when the load is a pure resistance equal to the line Z_C, because all the incident energy is absorbed and there is no voltage or current variation when going from the line to the load. Thus there is no electric and magnetic field change, no new voltage or current generated, hence no reflected wave.

As the reflected wave propagates back up the line as a separate and distinct electromagnetic traveling wave, it encounters only the same low-loss line with resistive Z_c[12] encountered by the incident wave in its forward travel to the load. Hence, the magnitudes of both the reflected voltage and current remain constant as the wave plows rearward, having the same values as when leaving the reflection generator. (There is a *gradual* diminishing effect because of attenuation; this will be discussed later.) They are, completely unaffected by the standing waves being developed as the reflected and incident waves, slide past one another. The incident voltage and current waves are similarly unaffected, continuing in their forward travel with constant magnitude until reaching the load.[13] Also, as in the incident wave, both the reflected voltage and current pass through zero simultaneously, and reach their maximum values one-quarter cycle later, because the line Z_c is resistive. Are not the reflected voltage and current then also in phase with each other, like the incident voltage and current? Perhaps, but let's not overlook polarity -- the voltage maximum *could* be positive when the current maximum is negative, in which case they would be 180 degrees out of phase with each other. But does the phase really matter here? Indeed it does - there is probably no other relationship more important to the principles of wave mechanism on a transmission line! Even though (the incident and reflected waves travel separately in opposite directions, they are inescapably related to each other through the common line and load characteristics, and their respective voltages and currents add *vectorially* at every point along the line as the two waves slide past each other. Hence the polarity, or phase relationship between the voltage and current in both sets of waves determines the character of the resultant standing waves, line-input impedance, and any other effect resulting from the vector combination. Many aspects of transmission-line phenomena which seem difficult to follow can be resolved rather easily if we understand how the phase, or polarity relationships evolve. So how do we determine the polarity and how do we establish a reference?

Wave-Travel Analysis

Consider a single wave traveling in a two-conductor line. By following conventional current flow we can select the appropriate conductor as the voltage-polarity reference for a given direction of wave travel which will cause the voltage and current maxima to occur with the same polarity. This polarity relationship may be reversed simply, either by selecting the opposite conductor for the voltage reference or by reversing the direction of wave travel. Obtaining the opposite polarity or phase relationship by reversing the conductors is a simple enough concept, but obtaining it by reversing the wave-travel direction has been a point of confusion for many people.

To reduce the confusion factor, a set of simple current-flow diagrams showing both ac and dc treatment is presented in Fig. 3. Use of dc with center-zero meters as indicators makes the explanation of polarity easy. Conventional needle movement is to the left for negative polarity and to the right for positive. Once polarity is clear, the battery may be replaced with the ac generator, and phase will also become clear using the wave forms as indicators. The setups in A and B of Fig. 3 are the same as in C and D respectively, except that the voltmeter terminals have been reversed. Observe that the wave or energy-flow direction and voltage-polarity reference selected in A cause line voltage and current flow to be in the same polarity. Now notice that reversing either the wave direction (as in B) or the voltage reference (as in C) both result in *opposite* voltage and current polarities as stated previously. Observe that reversing both direction and voltmeter reference polarity (as in D) again results in line voltage and current flow with the same relative polarity, though reversed from A. It may be helpful at this point to perceive that changing the wave or energy-flow direction is simply equivalent to reversing the terminal connections of the current meter, because the source changes sides. This is the key to understanding the polarity-reversal problem, because for a given voltage-polarity reference the current-flow direction has to reverse when the wave-flow direction reverses.

On the basis of the conditions stated above, if a generator is now placed at each end of a single two-conductor line, a reference selected to make the voltage and current in phase on the line for one generator will result in 180-degree out-of-phase voltage and current for the other generator. This is the situation which exists with the mismatched rf transmission line -- a source generator at one end and the reflection generator at the other. By selecting the conventional reference to make incident voltage and current in phase with each other (or $\theta = 0°$) it follows that reflected voltage and current must he 180 degrees out of phase with each other (ref. 35, p. 23).

It is of interest at this point to be concerned with the nature of the power in the incident and reflected waves. Some writers contend erroneously that the voltage-current phase relationship in the reflected wave is 90 degrees. If this was true, then the wave would contain only reactive volt-amperes, but no real power. The evidence above disproves this contention since we, have seen that the voltage-current relationship in the reflected wave is 180 degrees and not 90. And certainly we will agree that if real power is conveyed in A of Fig. 3, it is also real power in B, or C, even with reversed current-meter or voltmeter terminals. We will agree also that real power, *P* equals *EI cos θ*, in which cosine θ is the power factor. It matters not whether the phase angle is 0 or 180 degrees, for cos 0° = 1 and cos 180° = -1. This simply connotes the polarity difference discussed above. When conductor spacing is restricted to the near field, the fundamental principles governing transmission-line propagation are the same as those which govern all general ac-circuit relations, including electric power transmission. From these principles we know that real power flows for every value of θ in all four quadrants, except at 90 and 270 degrees where the cosine is zero, yielding zero power factor. Wherever the phase is other than 0, 90, 180, or 270 degrees, both real power and reactive, volt-amperes are present. But at 0 or 180 degrees, *only* real power exists because the absolute value of power factor is 1.0 in either case. This clearly proves that reflected power and incident power are both *real* power, and that no fictitious power, or reactive volt-amperes, exists in either one, because the current and voltage in the reflected wave are always mutually 180 degrees out of phase and the voltage and current in the incident wave are always mutually in phase.

The conflict concerning real vs. reactive power in reflected waves arises in part from confusion between traveling and standing waves, because of insufficient familiarity with both types. To broaden the familiarity, we have logically concentrated first on the *traveling* incident and reflected waves, because, from the physical view point, standing waves are derived from the resultant interaction between the two traveling waves. And thus, sufficient pertinent knowledge concerning traveling waves is essential before one can correctly understand the formation of standing waves and other correlated phenomena occurring on the transmission line which will become apparent as we proceed.

Now that we have a reasonably enlightened background concerning the ingredients; of standing waves, let us explore the details of their development, after which it will be appropriate to return to the power conflict for a few brief comments to clear away any remaining confusion.

Vector Graph Explanation and Standing-Wave Development

The newly launched reflected voltage and current waves, in their rearward travel toward the generator, combine with their respective incident waves at every point on the line. The continuously changing relative phase differences along the line cause alternate cancellation and reinforcement of the voltage and current distribution on the line. This results in the formation of the well-known standing wave and a change in the input terminal impedance from the initial line Z_C value.

A physical picture of this complex relationship greatly enhances the understanding of the phenomenon. Accordingly, Fig. 4 graphically illustrates the progressive phase relations with accurately scaled vector plots of the incident and reflected wave for visual comparison at every 22.5-degree (sixteenth-wavelength) point on the fine. The reference is from the termination point back toward the generator. These vector plots, being superimposed circularly around the Smith chart, present certain symmetrical phase-angle relationships with respect to line length which an not obvious with previous displays.[14] *(Refs. 2. p. 12; 17, p 146: 18. p. 110.)* This visual aid, in addition to providing a new dimension in presenting the formation of the standing wave, vividly emphasizes the development of capacitive and inductive components of complex impedance, the understanding of quarter-wave impedance inverting action, the impedance-repeating phase-inverting action of the half-wave line, and the reciprocal relationship between impedance and admittance.

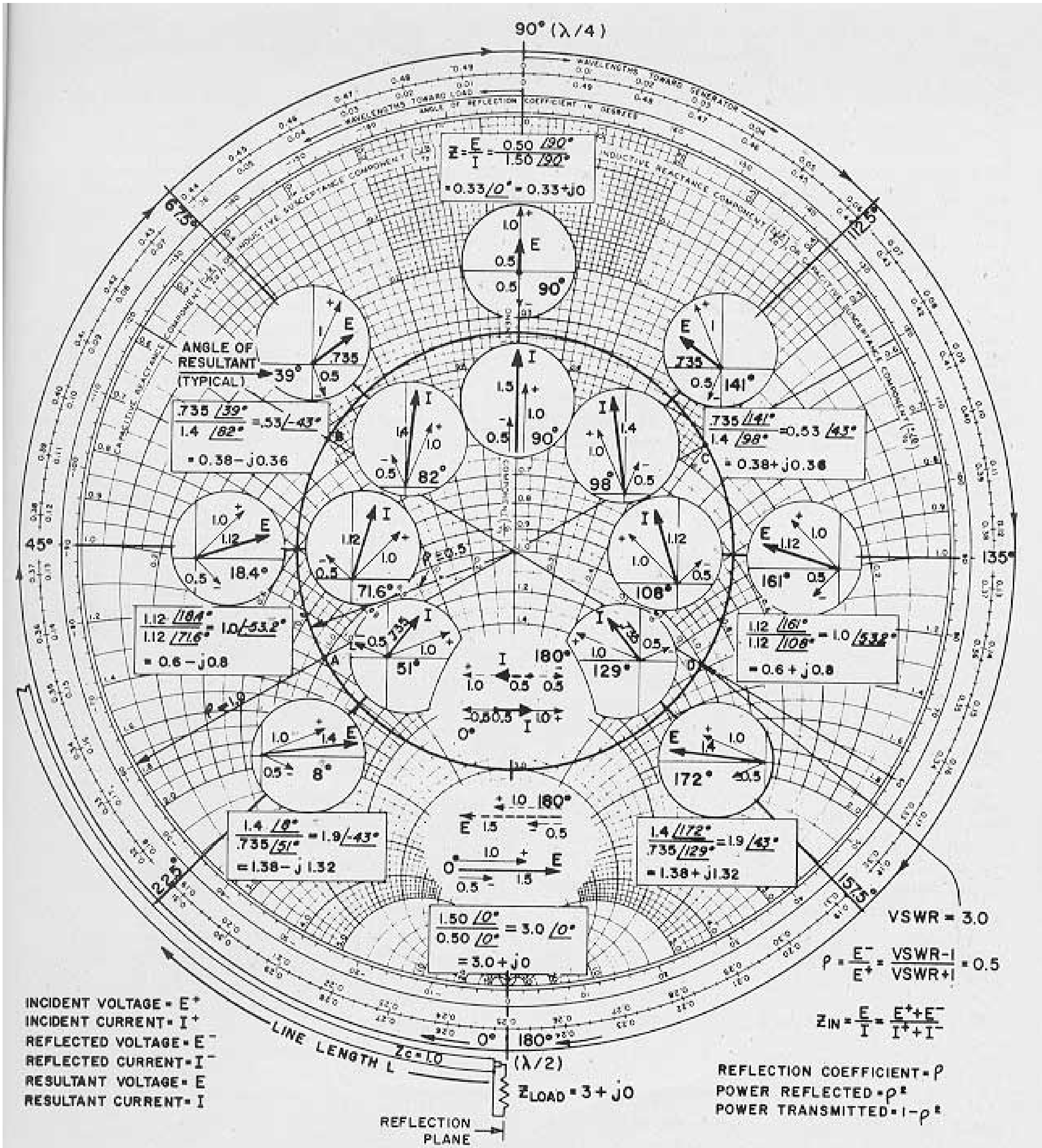


Fig. 4 — The W2DU Reflection Mechanics Vector Graph.

For this illustration we have terminated the line with a pure resistance equal to three times the line impedance Z_C. Hence a circle representing an SWR of 3.0 is shown. Line length, *L*, measured from 0° at the termination point (or reflection plane) toward the generator, is represented by clockwise rotation around the chart.

Current vectors are displayed inside the SWR circle, marked "+" for incident current (magnitude 1.0), “-“ for reflected current (magnitude 0.5), and *I* for the resultant current. Directly opposite and outside the circle are the corresponding voltage *E vectors*. The number in degrees shown with each vector set indicates the angle of the resultant. The vector lengths are proportional to the amplitudes of their respective voltage and current waves. By setting the incident-vector length equal to 1.0 the reflected vector length, by definition, equals the magnitude of the reflection coefficient, ρ. The reflected-vector lengths are thus 0.5 because ρ = 0.5 for an SWR of 3.0, from the relationship

$$\rho = \frac{\text{SWR} - 1}{\text{SWR} + 1} \qquad \text{(Eq. 2)}$$

While the vector lengths are proportional to each other, the lengths are not scaled to any chart dimension. These vector plots at each angular position on the graph contain the necessary amplitude and phase information to define both the standing wave and the impedance at the point on the transmission line represented by the point where the SWR circle intersects the radial line at each *L* = 22.5° interval.

The SWR = 3.0 (or ρ = 0.5) circle is based on the chart scales, and may be observed to have a radius of one half the chart radius. The perimeter of the chart has a radius of 1.0, representing total reflection, i.e., an infinite SWR, The ρ-circle radius is thus directly proportional to the reflection coefficient ρ, so an SWR or ρ circle may be constructed for any value of reflection by making the radius equal to the reflection value, ρ.

In order to make valid phase comparisons between points on the line, wave motion has been frozen at an arbitrary point in time, so that all vectors are shown in their true positions, *relative to each other*. It makes no difference when the motion is stopped, but the symmetry of the presentation is enhanced if we stop the motion when the incident vectors at the reflection plane are pointing in the zero direction in the standard polar coordinate system,

Observe that at any point on the line the incident voltage and current are always in phase, while in contrast, the reflected voltage and current are always 180 degrees out of phase, thus illustrating the conclusion of the discussion on this point in the previous section on basic reflection mechanics. At the reflection plane (*L* = 0°) it may be seen that all components are in phase except the reflected current, which is 180 degrees from the others, this is as it should be when the terminal impedance ties between Z_C and an open circuit.

As we travel clockwise from the reflection plane toward the generator, each incident and each reflected vector rotates the same number of degrees as the change in position along the line. But observe carefully, for this is very important: The incident-wave (+) vectors rotate counterclockwise (phase leading), while the reflected-wave (-) vectors rotate clockwise (phase lagging). For example, at 45 degrees from the termination the incident voltage vector is at +45 degrees, while the reflected voltage vector is at -45 degrees, for a total phase difference of 90 degrees. Thus, for every degree of motion along the line the relative phase angle between the incident and reflected voltage changes two degrees.[15] This can be readily understood when we consider that in the distance from the reflection plane to our observation point the reflected wave has traveled twice as far as the incident. From the observation point the incident wave travels only in the reflection plane, while the portion of the incident that is reflected travels an equal additional distance in returning to the observation point.

Now let's see what happens at *L* = 90°, or λ/4 of travel from the load. At the load, where *L* = 0°, the incident and reflected voltages are exactly in phase with each other, giving the reinforcement mentioned earlier (resultant = 1.5). But at 90 degrees toward the generator the two voltage vectors have each rotated 90 degrees in opposite directions and are now 180 degrees out of phase and opposing (resultant = 0.5). Going on to *L*= 180°, or λ/2 from the load, we see that each voltage vector has now rotated 180 degrees, but in opposite directions of rotation. Hence the vectors have rotated 360 degrees relatively, and once again are exactly in phase with each other and reinforcing.[16]

At the points between $L = 0°$ and $L = 90°$, the resultant voltage vector, *E*, is seen to diminish gradually from the maximum of 1.5 to a minimum of 0.5 and then increase back to the 1.5 maximum at $L = 180°$. In Fig. 5 these resultant magnitude values at each point have been plotted on the more familiar rectangular coordinate graph. The smooth curves connecting the plotted values indeed yield the familiar standing-wave pattern. The curves verify the relationship

$$SWR = \frac{1+\rho}{1-\rho} = 3.0 \quad \text{(Eq. 3)}$$

by first adding to and then subtracting ρ - 0.5 from the incident voltage 1.0.

Line lengths greater than a half wave are accommodated merely by continuing on around the circle again (Fig. 4), repeating the some values encountered 180 degrees earlier, thus establishing the periodicity of the standing wave. Only lossless line is being considered here; correction factors for attenuation, which change the circle into a spiral, will be presented later. The basis for the phase- or polarity-reversing characteristic of the 180-degree or λ/2, line may be observed on the vector graph by noting that the specific phase of the voltage vectors at the 180-degree point on the line is 180 degrees from their phase orientation at the reflection plane ($L = 0°$).

The significance of the constant 180-degree phase difference between the reflected voltage and current will emerge if we now compare the phase and magnitude of the current vectors with the voltage vectors previously studied. We see that at the reflection plane ($L = 0°$), where the incident and reflected voltage waves are in phase and adding to create a voltage maximum, the corresponding currents are out of phase and opposing to create a current minimum. And at $L = 90°$, where the voltage waves are out of phase and opposing for a minimum, the currents are in phase for a maximum. Which graphically illustrates why the maxima and minima of the voltage standing wave are ALWAYS separated by 90 degrees from the corresponding maxima and minima of the current standing wave. This phenomenon is caused directly by the 180-degree phase difference between the

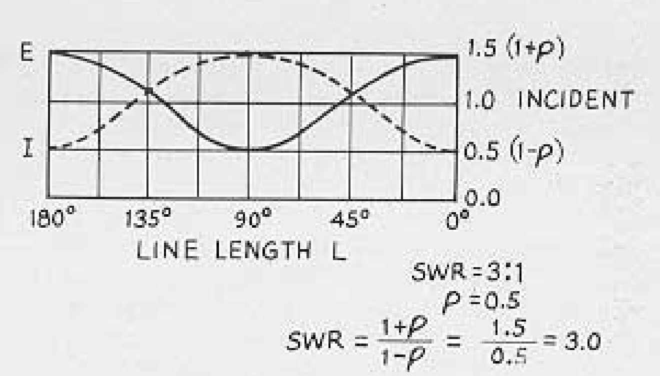


Fig. 5 — Illustrating how the voltage and current standing waves relate to the magnitude of the reflection coefficient, ρ.

reflected voltage and current, as strikingly shown by the vector display.

A visual comparison of the angular positions and magnitudes of both the voltage- and current-vector resultants on the vector graph (Fig. 4) may be made with the corresponding positions along the plot of Fig. 5. This comparison will enhance the understanding of this important concept. It is important because as we proceed it will be seen to be the basis for the impedance-transforming properties of the transmission line, including the quarter- and half-wave sections, which are only two specific conditions of the general case.

As we discuss the concept of impedance in the next section and solve matching problems later on, conjuring up a mental image of the action occurring on the line will at times be more helpful in understanding the difficult points than through logical reasoning alone. It is especially important to have a clear image of the formation of the reflected wave, because the reflected wave must be considered as a separate traveling wave identical to the incident wave, except for the direction and, usually, the magnitude. This point is important, because it helps us keep in mind that the *reflected* voltage and current waves travel the line 180 degrees out of phase with each other, and thus transfer actual real power during their travel. It is essential, to the process of reinforcement and cancellation of voltage and current along the line in the formation of the standing wave that real power be conveyed in the reflected wave, as if it had been developed by many tiny little generators at every point all along the line. It will also become clear in the section to follow why the impedance along a line changes in the presence of reflections only because real power is flowing, in *both* directions. *(refs. 2, p. 70;35, p. 24; 42).*

We are stressing this point because, as mentioned previously, some writers have presented the erroneous viewpoint that the reflected wave conveys no real power, with the argument that the reflected voltage and current are 90 degrees out of phase with each other and are therefore wattless. The reflected wave would indeed convey zero power if its voltage and current *were* 90 degrees out of phase, but the, argument is incorrect because they actually travel 180 degrees out of phase, as we have previously shown. At least two writers, W9IK[17] and W5GO,[18] would have us believe otherwise. It is easy to reach the wrong conclusion, however, because of the lack of a clear image of the reflection process and the somewhat complicated wave mechanics on the line. It may be that W9IK has confused the *reflected* wave with the standing wave, which both writers may have confused *reflected* voltage $\overline{E}$ and current $\overline{I}$ with line voltage *E* and current *I*.[19] The true nature of the reflected wave as a separate electromagnetic traveling wave must be appreciated. It is interesting to note that W9IK states, ". . . re-reflection of power at the input end (of the line) is impossible to accept since the necessary conditions of impedance mismatch are not present," and yet his Fig. 2 shows a, means for obtaining a (conjugate) match. This is also true in his Fig, 1, since he implies that his final amplifier is loaded and tuned properly. Now the very essence of the conjugate match is its totally reflecting mismatch for waves traveling toward the generator, while presenting a perfect match for waves traveling toward the load. Yet misunderstanding of this important basic concept is widespread among the amateurs. This concept is basic to the operation of passive frequency-selective filters. and the mechanism for developing the conjugate match simply constitutes a filter of this type. The mechanism behind its operation involves wave interference and reflections, which will be described in detail shortly, again using the vector graph as a visual aid.

It is understandable that the 90-degree difference of position on the line which exists between the voltage and current maxima (or minima) of the *standing* waves (described a few paragraphs earlier and shown in Fig. 5) could have been mistaken as the phase difference between the *reflected* voltage and current. And the similarity between *line* voltage and current behavior with the voltage-current relationship in ordinary ac circuitry also makes it easy to understand why *line* voltage *E* and current *I* are being confused with the *reflected* voltage $\overline{E}$ and current $\overline{I}$. This is because, in addition to the in-phase line voltage and current components, which convey only the *net* power flow, line voltage and current do contain reactive components which are 90 degrees out of phase with each other when reflections are present. Obviously, these reactive components convey no real power. Unfortunately, some people who are well versed in ac circuitry using lumped constants, but who are less familiar with transmission-line operation, sometimes make the error of assuming that the two circuit types are identical in electrical performance, and so we should be wary of making unwarranted comparisons which can bring about disastrous consequences of the type we are attempting to straighten out here.

Another specious argument set forth is that reflected power cannot be real power because it cannot perform work. We will prove this argument erroneous by showing how power in the reflected wave can do work. Now a simple rf voltmeter and ammeter in the line will indicate only the *resultant* voltage and current of the *combined* Incident and reflected waves (*E* and *I* on the vector graph), the product of which, when weighted by the cosine of the phase angle between them, yields only the resultant, or *net power* flowing toward the load. But there are many devices in common every-day use which selectively extract either the reflected or the incident wave (or both) from the line separately (independently of the standing wave), and thus permits separate measurement and analysis of the power associated with the waves traveling in either direction. One such device is the directional coupler. Another is the circulator. This is a three-port directional device in which the reflected wave from a mismatched load on the second port is completely diverted away from the input feedline and emerges from the third port. The reflected wave cannot get back onto the feed line to interact with the incident wave to develop a standing wave, and thus does not change the line-input impedance at the source feeding port one. However, current flow through a resistor placed on the reflected-power-output port (port three) develops heat (I^2R) equal to the amount of the power reflected from the mismatch at port two. A four-port hybrid coupler can be connected to perform in the same manner as the circulator. Directional rf devices most familiar to amateurs are the simple reflectometer SWR indicator and the directional wattmeter *(ref. 18, p. 180; 38; 40; 42).* The meter in either one is actuated by rf power -- *absorbed* from one of the traveling waves on the line -- either the forward or reflected, as selected. If the reflected wave were wattless reactive power, no power would be available to actuate the meter movement in the SWR indicator, or to produce heat from the current flow in the resistor on the third port of the circulator. Furthermore, for power to become wattless on being reflected would violate the most general and fundamental of all physical laws, namely the law of conservation of energy *(ref. 35. p. 25),* On the basis of this law, if all of the energy flowing in the line toward the load cannot be absorbed in or dissipated by the load, that portion which is not absorbed must appear somewhere. It cannot just disappear or cease to exist as if by magic. The reflected power recovered as heat in the circulator is a typical proof of this fact.

Here is another way of expressing net power flow through the line, which enables us to break power down into its incident and reflected power components. The expression is obtained from the power formulas, e.g.,

$$P = EI \qquad \text{(Eq. 4)}$$

or

$$P = \frac{E_{max} \times E_{min}}{Z_c} \qquad \text{(Eq. 5)}$$

On rf lines, $\text{power} = P = E_{max} \times I_{min}$ (Eq. 6)

Or

$$P = \frac{E_{max} \text{ x } E_{min}}{Z_c}$$

Now E_{max} = (produced by $E^+ + E^-$) occurs as shown in Fig. 4, where the incident and reflected voltages are in phase at L = 0 and 180°, and E_{min} (produced by $E^+ - E^-$) occurs where they are 180 degrees out of phase at L = 90°, Later we shall see that both the resultant voltage E and current I are nonreactive at these points on the line, while being reactive everywhere else along the line between the points. Because these points are nonreactive, the products of their voltages divided by the line impedance, Z_C, yields the net power flow exactly. But recalling that E^+ and E^- are always nonreactive, we can replace the term E_{max} by $E^+ + E^-$ and E_{min} by $E^+ - E^-$, and thus:

$$P = \frac{E_{max} \text{ x } E_{min}}{Z_c}$$

$$= \frac{\left(\left|E^+\right| + \left|E^-\right|\right) \times \left(\left|E^+\right| - \left|E^-\right|\right)}{Z_c} \qquad \text{(Eq. 7)}$$

Multiplying out the numerator terms gives the desired incident and reflected components:

$$P = \frac{\left|E^+\right|^2}{Z_c} - \frac{\left|E^-\right|^2}{Z_c} = \text{net power flow} \qquad \text{(Eq. 9)}$$

The first term on the right of the P expresses the power associated with the incident wave and the second term the reflected power. This simple separation of power into two separate components, each associated with one of the traveling waves, can be done on a lossless or low-loss line, where the Z_C is resistive. If the line has appreciable loss the interaction of the two waves gives rise to a third component of power which we can disregard, as lines normally used by amateurs are usually in the low-loss category. *(See refs. 18, p. 150, and 37, p. 129.)*

This separability of the forward and reflected powers forms the physical basis for the operation of reflectometers and directional wattmeters, *(ref. 38)* in which either the forward or reflected component is sensed by taking advantage of the 180-degree out-of-phase relationship of the reflected components of voltage and current while the forward voltage and current components are in phase with each other. In these operations a sample of the *voltage across the line* is added to a sample of a voltage derived from the *current in the line*. When the amplitudes of the samples are adjusted to the correct relationship (determined by line impedance Z_C), the two reflected components cancel, so that the sum represents the forward component alone. By reversing the phase of the current sample 180 degrees, the forward components cancel and the resulting sum represents the reflected components alone. A meter connected to indicate the voltage sums can now be calibrated in *power*, because the square of the voltage is proportional to power. Thus, the beautiful aspect of the directional wattmeter is that. when properly calibrated, it indicates the *true* power in the transmission line with the line terminated in *any* load impedance; the load can be either a match or a mismatch, and it can be reactive or nonreactive. The meter does this because the forward power value is always equal to the sum of the line-input power plus the reflected power; thus it indicates the true power which is actually incident on the load. In the reflected-power position the meter indicates the amount of the incident power which was not absorbed by the load, but which adds to the line-input power from the transmitter at the line input, or at whatever point in the line the conjugate match is performed. *(Ref. 18, p. 191.)* The difference between the forward and reflected readings, is, thus, the *net* power flow in the line at whatever point the wattmeter is inserted. In a lossless line the net power flow indicates the line-input power, which is the absorbed power exactly; the two are identical *anywhere on the line*. In a line with attenuation the meter indicates the line-Input power if it is placed at the line input, or it reads the absorbed power if it is placed immediately ahead of the load. The difference between these readings is related to the line attenuation. Of course there may be, for practical reasons, errors in the actual results from SWR measurements -- diode nonlinearity at various power levels, for one example *(ref. 40).*

Except for a somewhat different viewpoint concerning the nature of reflected power, the work of W6EL, formerly K6CYG. closely parallels the basic theme of this series of articles and reaches the same conclusions (ref. 44). Reference to Shallon's work was omitted from Part 1 of this series because of author error, but his work deserves thorough review at this point.

A detailed explanation of why the points mentioned above take place and of the wave mechanics of conjugate matching will appear in Part IV of this series, in a subsequent issue of QST.

Part 4 - A View into the Conjugate Mirror

IN PART 3 some basic concepts were presented concerning reflection generation, wave propagation along the line, and the development of standing waves. Then it was shown mathematically how net power is separated into its incident and reflected components (on lossless and low-loss lines), after which it followed logically to explain how the power separation is physically realized by directional devices, such as a directional wattmeter. While learning about wattmeter operation and how to interpret the indications, we saw that the incident or forward power in the line between, the matching point and the load is greater than the power supplied by the source generator when the line is terminated in a mismatched load. In this part we will explore this situation in detail, because it is of considerable importance to the amateur since it relates directly to the operational flexibility of his antenna system. Appreciation of the fundamentals involved in this seemingly anomalous situation will free him from the prevalent notion that he is restricted to operating with little or no mismatch at the antenna/transmission-line terminals.

The explanation of directional wattmeter operation in Part 3 should help in understanding why the incident power appearing on the line between the matching point (such as a line matching network) and the mismatched load can be higher than what the transmitter can supply. This is a normal condition which *must exist* in order for a mismatched load to absorb *all the power delivered* by the source,[20] while at the same time reflecting a percentage of the total power it receives. To do this, the load must receive more incident power than what is supplied by the transmitter. The basis for understanding this rather subtle concept lies in the wave mechanics behind the principles of conjugate matching introduced in Part 1 and defined in Part 2. The wave aspect of this subject has been presented in the literature (known to the author) only by Slater *(ref. 35)* and Alford *(ref. 39)*. Perhaps this restricted exposure may account for some of the confusion in this area among engineers and amateurs alike. For example, the many "cook book" recipes and graphical directions for stub-matching a mismatched line tell "how" to do it, but offer little insight toward visualizing the wave mechanics through which the match is accomplished. This insight, however, goes to the heart of the transmitter-to-line coupling problem. It clarifies how the reflected wave becomes rereflected at the matching point. If the matching did not produce this effect, the reflected wave would travel back to the generator, and would thus reduce the amount of the power made available by the source generator *(ref. 19, p. 37).*

Reflection Mechanics of Stub Matching

An introduction to the reflection mechanics involved in conjugate matching concerns concepts of line-input impedance and angle of reflection coefficient, which will be explored in detail in a later section. (The reflection coefficient angle was introduced briefly in Part 3, para. 1, and footnote 15.) As stated with the definition of the conjugate match in Part 2, the matching is accomplished by inserting a nondissipative mismatch at the match point; this produces a complementary reflection with which to compensate, and cancel, the wave reflected from the mismatched load. It was also stated that conjugate matching conditions can be satisfied by a correct adjustment of either the final tank tuning circuit *(ref. 4, Part 3),* or of a line-matching network if one is used. Because stub matching uses the identical principles and is easier to visualize, we will use its technique to demonstrate the wave mechanics. In stub matching, the stub provides what seems like an anomaly -- a nondissipative discontinuity, or mismatch. While we usually think of the stub as providing a *match,* rather than a mismatch, we will discover that the conjugate match results from the mutual cancellation of two complementary reflected waves generated by two complementary mismatches. One wave is that reflected from the terminating load mismatch, and the other is a new reflected wave generated by the stub mismatch, equal to the load-reflected wave in both magnitude and phase, but of *opposite phase sign*. Wave interference between these two complementary waves at the stub point causes a cancellation of energy flow, or a null in the generator direction resulting from the *difference* between the two reflected waves, and an energy maximum in the load direction comprising the sum of the two reflected waves and the source wave. The effect of the wave-interference cancellation presents a virtual one-way open-circuit[21] to waves traveling toward the generator, which blocks both the load-reflected wave and the stub-reflected wave at the stub point from any further rearward travel. These waves are totally reflected toward the load, being in phase with the incident wave. Thus, there is complete cancellation of the effect of discontinuities (such as the stub) and the reflections of waves traveling toward the load.

This wave-interference mechanism which accomplishes the matching will become more evident as, we investigate the reflection coefficients of the two mismatches with the aid of examples using the Vector Graph (Fig. 4, Part 3). There we show a load of 3 + *j*0, which gives a three-to-one standing wave ratio along the whole line.[22] The VSWR of 3 is shown by the dark concentric circle, which is the locus of impedances anywhere on the line for a load of 3 + *j*0. This circle intersects the resistance circle marked 1.0 at the two points, A and D. Therefore at these points along the line the resistive component of the impedance equals 1.0 times Z_C, which is the desired point for attaching a matching stub. Points A and D are also intersected by a reactance circle. At point A the reactance is negative, 1.15 times Z_C, and at point D the reactance is positive, also 1.15 times Z_C. The conditions for obtaining reflected-wave cancellation by wave interference are the well-known stub-matching requirements as follows: (1) a stub is placed where the line resistance component equals the line characteristic impedance Z_C (such as at points A and D), and (2) the stub reactance is made equal in magnitude and opposite in sign to the line reactance (resulting from the phase relationship between the incident and reflected waves at the stub point) so that the reactances cancel to zero, *(refs. 2, p. 116; 19, p. 97)*. This sounds almost like the conjugate-match definition itself, doesn't it? The correct point for inserting a stub in series with the line nearest the load is at point A or D on the VSWR circle; any half-wave interval from these points further from the load may also be used since the impedances are repeated every half wave on the diagram (and on the line). These examples show series stubs to permit *impedance* treatment for clarity. (While parallel or shunt stubs are used more prevalently, analysis using the shunt form would require *admittance* treatment.)

We will now see how reflections add at a matching point to produce the matching effect. We'll also see why a directional wattmeter will give a true reading of incident power between the matching point and the load which is greater than the power supplied by the transmitter, when the line is terminated with a mismatched load. A little later we'll also see how these principles apply to practical feed-line matching networks.

At point A, which is 30 degrees from the load (at *L* = 30°), the unmatched voltage reflection coefficient is $\bar{\rho}_E = 0.5 \angle -60°$. This means that the phase of the reflected voltage wave lags the incident wave by 60 degrees at point A. The line impedance *E/I* at this point is 1 - *j*1.15. A match can be effected by connecting an inductive reactance, such as a stub or a lumped inductance of 0 + *j*1.15 in series with the line at point A. Now, the reactance-cancellation effect of the positive reactance stub on the equally negative reactance of the line is generally understood, but several points are not always clear regarding the effect on the component waves: What characteristics of the stub cause it to counteract the reflections from the load; also, why does the stub cause the incident power to rise between the matching point and load'?

In answer to these questions, let's determine first the reflection coefficient produced by the stub if it were inserted in a perfectly matched line. In this condition we may analyze the stub generated reflection in the absence of any other disturbance or reflection on the line. If we station ourselves just on the load side of the stub point with the stub attached and look, into the line toward the matched termination, we will see a pure resistance equal to the line characteristic impedance, Z_C. We know from matched-line theory that if we remove the line portion extending from the stub to the load and insert the terminating resistance R = Z_C directly in series with the stub across the open-ended line, we may look into the line toward the stub from the generator and see the same conditions of reflection as were present before the line section was removed. Thus the series circuit comprising the

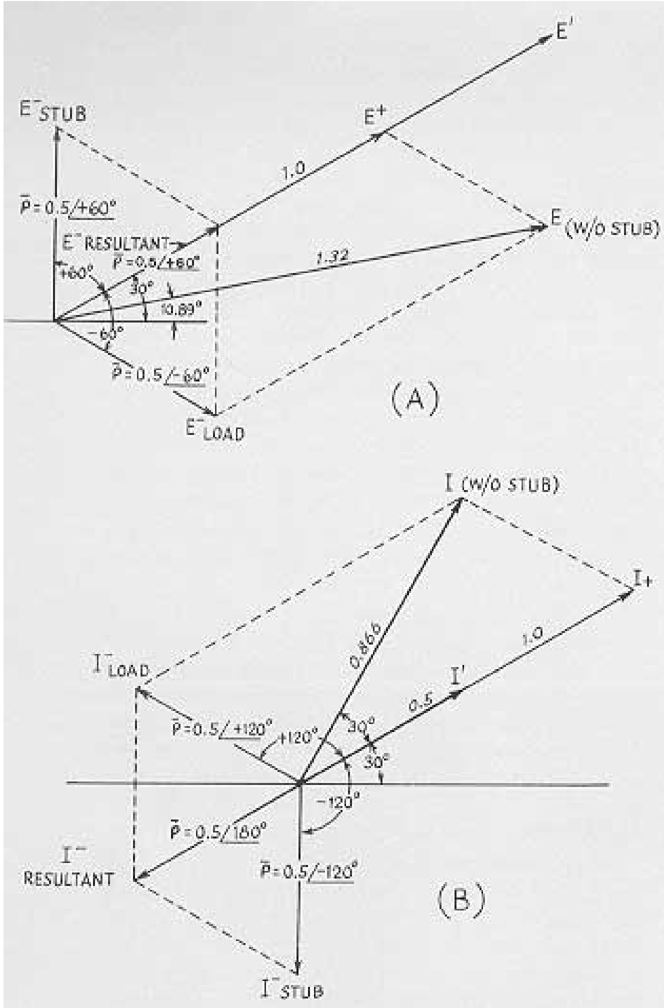


Fig. 6 — At A, various voltage-vector relationships, and at B, various current-vector relationships at *L* = 30°.

matching resistor and the stub performs as a 1 + *j*1.15-ohm mismatched load terminating the line. Precisely this same reflection will be produced no matter where the stub is inserted in a matched line. The Vector Graph shows this impedance of 1 + *j*1.15 to appear at point D, for which the voltage reflection coefficient is $\bar{\rho}_E$ = 0.5 ∠ + 60°. Note that this is the same magnitude and phase but of *opposite phase sign* to the reflection appearing at point A resulting from the load mismatch (3 + *j*0).

Thus the stub mismatch produces the same magnitude of reflection (and the same SWR) as was produced by the load mismatch, but the stub-reflected voltage wave *leads* the incident wave by 60 degrees, while the load-mismatch wave *lags* by 60 degrees. If the stub is now attached at the matching point (corresponding to *L* = 30° on the Vector Graph) with the 3 + *j*0 load terminating the line, both the stub- and load-mismatch reflections will be produced simultaneously. As a result of their opposite-sign phase relationship, the *leading* stub-reflected wave and the *lagging* load-reflected wave cancel each other at the match point. The, voltage reflection coefficients of load and stub thus add vectorially to zero degrees ($\bar{\rho}_E$ = 0.5 ∠0°), which tells us that the resultant of the two reflections is exactly in phase with the incident voltage wave at the match point as shown in Fig. 6A. (The amplitude resulting from this trigonometric addition will be considered later, but knowledge of these angular relationships should add to the appreciation of the mechanics of line-reactance cancellation by the stub.)

Now that we know what is happening with the voltage waves we also want to investigate the current waves to learn about the impedance relationship at the matching point. As defined earlier, reflected current is 180 degrees out of phase with reflected voltage, so the current coefficient is found on the Vector Graph 180 degrees away or diametrically opposite the corresponding voltage-coefficient point. Thus we find the current reflection coefficient for the load mismatch at point C with $\bar{\rho}_I$ = 0.5 ∠+120° Similarly, the stub-mismatch current coefficient found at point *B* is $\bar{\rho}_I$ = 0.5 ∠-120°. Note that the current coefficients of load and stub are also of equal magnitude and phase, but of opposite phase sign. But while the voltage angles add to 0 degrees, the current coefficients add vectorially to $\bar{\rho}_I$ = 0.5∠180°. The resultant current reflected wave is then 180 degrees out of phase with the incident current as shown in Fig. 6B. So we have incident and reflected voltages in phase, and incident and reflected currents out of phase -- and the wave arriving at the match point from the generator sees a perfect match.

These facts portray a significant message. In the opening paragraphs of Part 3 the wave mechanics involved in a line terminated by an open circuit were described. There we learned that the reflection coefficient angle of the voltage is 0 degrees and of the current is 180 degrees. With an open-circuit condition, the unabsorbed voltage wave which is incident on the termination is reflected with no change in phase, while the unabsorbed current wave is reflected with a phase change of 180 degrees, a complete reversal of polarity. The reflected-wave phase relationships at the match point which we established above (by the voltage and current resultants of the stub- and load-mismatch waves) indicate precisely the same conditions that prevail in a line terminated in an open circuit, so far as the reflected waves are concerned (but not the incident wave.) Therefore, the effect of the two reflected waves arriving at the match point is to establish an open circuit to the waves generated by the two mismatches. Thus, these waves become totally reflected at the match point and undergo open-circuit phase-change relationships as described above. The resultant reflected voltage wave thus does not change phase during this reflection; remember that it was already in phase with the incident wave prior to its reflection. The resultant current wave changes phase by 180 degrees on reflection, and because it was 180 degrees out of phase with the incident current wave just prior to its reflection, the present 180-degree reversal now places it also in phase with the incident current wave. Now that both voltage and current rereflected waves are *in phase* with their corresponding incident waves, addition of the voltages and currents occurs at the reflection point. Thus the conclusion: *The power contained in the reflected waves adds to the incident power*.

Before we proceed any further, let us consider that the above conclusion was based on the assumption that an open-circuit condition exists at the match point for the reflected waves traveling toward the generator. The assumption was based on the similarity between the reflection coefficients which we established at the match point by wave interference and those known to exist at an open-circuit termination. We can verify this assumption by an alternate method, based, for example, on the open-circuit magnetic-field theory from Part 3.

Let us first observe the net value of *all* currents flowing at the match point at the instant the two reflected currents form their resultant $\theta = 180°$; at this instant we will see an initial sudden drop in resultant line current *I* because of wave cancellation as the reflected-wave resultant becomes aligned exactly out of phase with the incident current. This drop is shown graphically in Fig. 6B, where the original resultant current *I* (as with no matching stub present) suddenly drops to the new *instantaneous* resultant value *I'* from the effect of the stub discontinuity. Now recalling briefly from the open-circuit field theory presented in Part 3, para. 2, we know that when current drops, the magnetic field also drops. The changing magnetic field produces an electric field equal to the energy reduction in the magnetic field. The new electric field adds in phase to the existing electric field, producing an increase of voltage at the match point. This increase in voltage now starts a wave traveling in the opposite direction, which is actually now in the same direction as the incident wave, thus adding to it. The increased electric field (now an enlarged *incident* electric field), as it moves toward the load, produces a new magnetic field equal in magnitude but of polarity opposite to that of the original field. This new magnetic field now causes current to build up again to the same magnitude as the original reflected current, but of opposite polarity and direction. Thus the new current wave is now also traveling in the same direction with the same polarity as the incident current wave, adding to it and enlarging it just as the rereflected voltage wave added to and enlarged the incident voltage wave.

By following these field-current-voltage reactions through their natural sequence of events, it can be seen that we have obtained the same conclusions as those previously obtained, thus justifying the assumption that the resultant reflection coefficients at the match point have defined an open circuit to the reflected waves. The existence of the reflectance at the matching point is therefore verified, with the result that both the reflected voltage and current have indeed been rereflected and the power associated with them has thus been effectively added to the power contained in the incident wave. Thus when the line is terminated in a mismatch, causing reflected power to exist on the line, the *sum* of the source and rereflected powers (which is traveling only toward the load) must be greater than the power delivered by the generator alone. And since we have shown how the stub acts to counteract reflections from the load, our original questions concerning the stub characteristics have been answered.

With a 3:1 SWR, where $\rho = 0.5$, 100 watts supplied by the transmitter will yield 133.3 watts of incident and 33.3 watts of reflected power. Neglecting losses, 100 watts will also be absorbed in the load. From Fig. 4, reflected power $\rho^2 = 0.25$, or 25 percent of the incident power, leaving 75 percent absorbed by the load ($1 - \rho^2 = 0.75$). Incident power is $1 / (1 - \rho^2)$ times the supplied power, so 1/0.75 = 1.333, and 1.333 times 100 watts equals 133.3 watts.

In a typical realistic case where the flat-line attenuation is 0.50 dB (corresponding to 175 feet of RG-8/U at 4 MHz, 85 feet at 14 MHz, or 85 feet of RG-59/U at 4 MHz), if the load were perfectly matched to the line (1.0 SWR) the 100 watts delivered would be attenuated to 89.13 watts during travel to the load. But with a 3:1 mismatched load the additional one-way line attenuation (because of the SWR) is 0.288 dB. The incident power at the conjugate-match point would then be 124.78 watts (0.288 dB below 133.33 watts), and 111.21 watts of power (0.5 dB below 124.78 watts) reach the load; 27.80 watts (25 percent) are reflected, leaving 83.41 watts to be absorbed. Of the 27.80 watts reflected, 24.78 watts arrive back at the input to join the 100 watts of source power to develop the 124.78 watts of incident power. The 5.72 watts difference between the power absorbed in the matched and the 3:1 mismatched load (0.288 dB) is insignificant. Information on calculating these values will be presented later. These values are typical of data obtained during actual routine measurements in a professional laboratory. They provide additional evidence that reflected power is real and not fictitious; if it were fictitious power, no more than 66.85 watts (75 percent of 89.13 watts) would be available to the 3:1 mismatched load. But the 83.41 watts actually absorbed is 93.58 percent of the amount absorbed in the matched load, the loss of 6.42 percent being completely accounted for in line attenuation alone.

Matching Networks and Reflection Mechanics

We now wish to delve further into the wave-interference principles demonstrated using the stub technique, to learn how the principles also apply to both resonant quarter-wave series matching-transformer operation and the typical amateur "antenna tuner" (line-matching network) or Transmatch. In order to visualize the inherent generality of these principles we need to develop some additional concepts concerning stub matching and embark on a somewhat different line of reasoning. As may be surmised from the example presented above, the fundamental principle behind the elimination of reflections is to have each reflected wave canceled at the point where the elimination of the reflection is desired by interference from another wave of equal magnitude and phase but opposite phase sign *(ref. 35, p. 58)*. A transmission line of the appropriate length which has one end effectively open circuited and the other end short circuited possesses the reflection-producing characteristics required to develop canceling waves of the correct phase in relation to the wave to be canceled.

Canceling waves can be developed by using other line arrangements, but for the purpose of demonstrating the principle, we will use the arrangement just stated, which, as shown in Fig. 7, shows how a stub performs the matching function in practice. Fig. 7A is the conventional representation of a typical series-stub circuit (using the values of our previous SWR = 3 example), in which section F is called the *feed line*, section S is the stub, and section T is an impedance-transforming section which we will call the *transformer*. We'll now discuss these in greater detail. To clarify the approach, Fig. 7A is redrawn in Fig. 7B, with the stub and transformer shown as one continuous straight-line section. This straight-line section will presently come to life as the heart of the wave-interference-producing mechanism found in all stub-matching operations. This is because its physical length will be adjusted arbitrarily so that waves reflected at each end will return to the feed point with equal magnitude and phase, but with opposite phase sign.

We have seen earlier that a voltage wave is reflected with zero phase change from an open circuit (or from any resistive termination greater than a matched load), and is reflected with 180 degrees of phase change from a short circuit (or from any resistive termination less than a matched load). With a current wave the inverse is true. So from the viewpoint of reflection behavior, one end of the straight-line section will be considered as being open circuited and the other end short circuited; which end will be open and which end short circuited will depend on the character of the load. In our example in Fig. 7B the load end behaves like an open circuit as far as wave reflection is concerned, while the other end (the stub) is short circuited. The action occurring in this line section in the process of developing the interfering wave-canceling relationship is as follows. Either a voltage or current wave is assumed to enter the line section at the feed-line entry point. The energy divides, one portion of the wave traveling toward one end of the section, and the other wave portion traveling toward the opposite end. After each wave portion encounters *one* reflection the returning waves will each have the same absolute value of phase but opposite sign, or polarity, on return to the point of entry.

The opposite phase polarity between the two reflected waves (arriving from opposite directions) results because reflection at one end is accompanied by a 180-degree phase reversal, while reflection from the other end is not. As stated above, the phase reversal of one wave but not the other is caused by the opposite conditions of reflection at the two ends of the line, one end open- and the other end short-circuited. Note in Fig. 7 that in each case, the phase of the reflected waves (of both voltage and current) is of opposite polarity on opposite sides of the feed line as they return from the stub and load directions. The wave entry point, where the feed line is attached, is the matching point, and divides the line section into its two complementary portions: the stub portion, S, and the *impedance-transformer* portion, T. Electrically, each portion is the complement of the other, because the waves reflected from the end of each portion returning to the match point are complementary in phase relationship and equal in magnitude. Herein lies the basis for the term *complementary mismatches* as used earlier, because each portion presents a complementary mismatch to the feed line.

We will see a little later that this complementary mismatch concept is of great importance to matching in general, because the complementary relationship holds no matter where the feed-line entry point is positioned on the stub-transformer line section. The importance prevails because the canceling wave and the reflected wave to be canceled will be of the same magnitude and phase but opposite phase sign at whatever point on the quarter-wave line section the feed line is attached. This is true with two provisions: (1) the characteristic impedance Z_C of the feed line F must be the same as the resistive component of the transformed impedance appearing on the transformer line at the feed point, and (2) the length of the stub portion S must be adjusted to produce a reactance equal and opposite in polarity to the line reactance appearing at the feed point. The length may be found from the expression

$$S_L = \arctan\left|\frac{jX}{Z_C}\right| \quad \text{(Eq. 10)}$$

where S_L is the stub length in electrical degrees
jX is the line reactance (obtained from the Vector Graph)
Z_C is the characteristic impedance of the stub section. (Z_C = 1.0 here, because we are using normalized impedances. See footnote 22.)

The transformer section T transforms the load to varying values along the line. Hence, for a proper match, the magnitude of the F feed line impedance, Z_C, depends on the location of the feed point and vice versa. This concept is not generally appreciated, and it is certainly not readily apparent from the usual stub-length and position-indicating graphs appearing in many publications.

Stub Matching Versus Network Matching

We are now getting closer to seeing how stub-matching principles extend to line-matching network operation. If we look further into the reflection characteristics of what have generally been considered to be *different* techniques of matching, a fascinating revelation of the similarity between all of these various techniques will emerge; stub, hairpin, λ/4-series transformer, Transmatch, *L* network, and so on, are all in this category. And there is a logical reason for this similarity; these techniques all have one essential ingredient in common -- reflections! Reflection and matching are applied in transformers that rely on reflections from the end terminals, where a change in impedance level exists. As we discussed in Part 3, any abrupt change in impedance level appears as a discontinuity to the smooth flow of the electromagnetic wave, and results in producing a reflection. The transformer accomplishes the task of matching its input and output impedances by controlling the phase and magnitude of the waves produced by reflection at its boundaries, or end terminals, so that all the reflections produced at either end are canceled by those arriving from the other end *(ref. 35, p. 58).* This is what was meant in the reference to "controlled reflections" in Part 1, para. 3. A corollary to the seeming anomaly of the stub producing a *mismatch* instead of a match, is

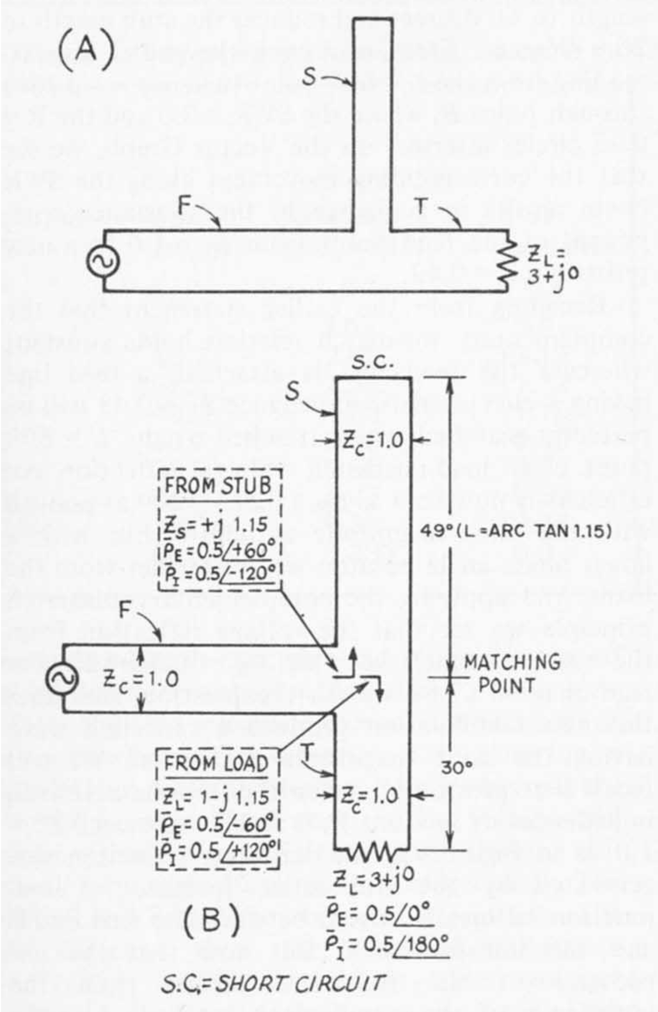
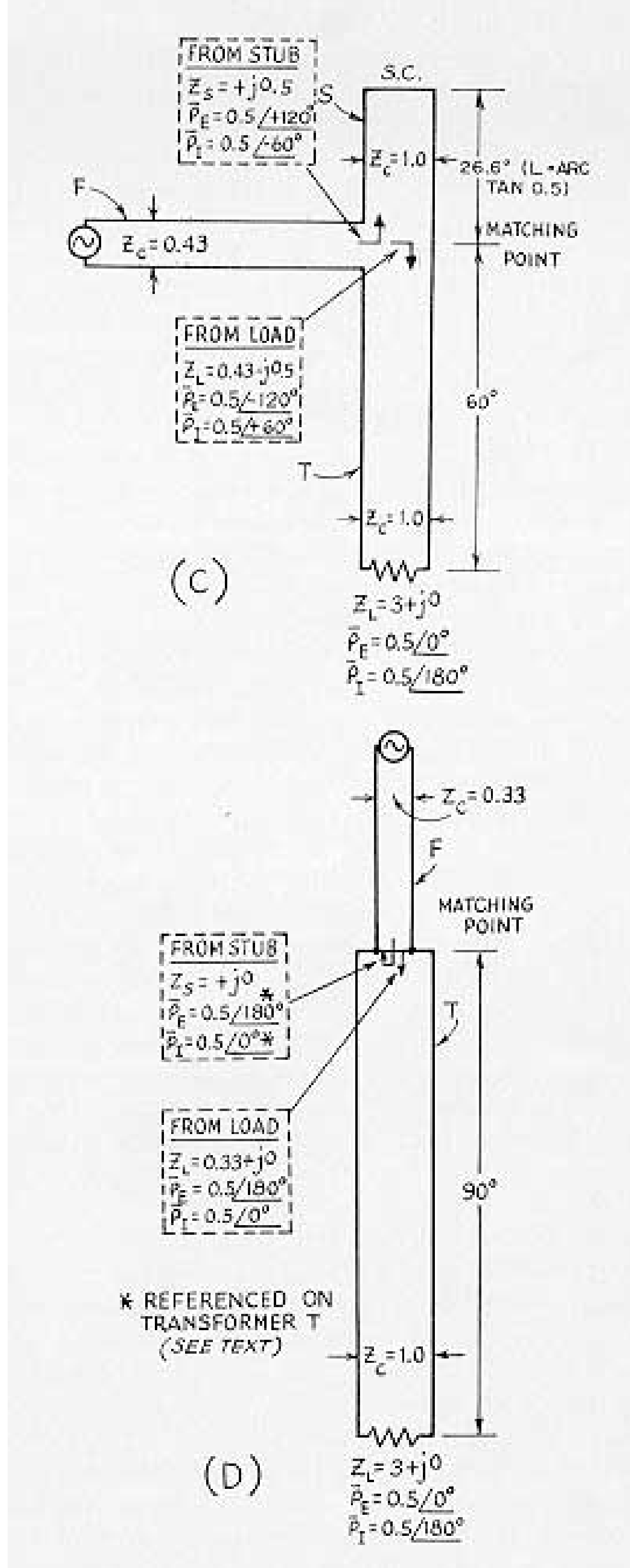


Fig. 7 — At A, the concept of stub matching with a series-connected stub. At B, the same matching arrangement redrawn. At C, a similar matching arrangement but with different stub and transformer line lengths. At D, a quarter-wave line-transformer section evolves as the stub length goes to zero.

that we match to avoid reflections, but we can't match without them when different impedance levels are involved.

From Eq. 1, Part 3, we know that the load-reflection magnitude is determined by the ratio between the load impedance Z_L and the line impedance Z_C of the transformer. So in furthering the understanding of the role played by reflections in the process of impedance matching, it is interesting to make two additional observations on the Vector Graph. First, the reflection magnitude, or SWR, determines the position on the transformer section T where the resistance component of line impedance *E/I* equals the line Z_C. This position, we recall, is the matching point, and fixes the length of the transformer section T. In making this observation remember that the diameter of the SWR circle is proportional to the VSWR. By



tracing along the R = 1.0 circle we can see that as the diameter of the SWR circle changes, the point where the R = 1.0 circle and the SWR circle intersect moves accordingly. A radial line drawn through this intersection point, and extending to the line-length scale *L*, will thus indicate the angular distance (T) from the load to the matching point for a given SWR. We recognize this observation as simply the conventional method of using the Smith Chart for determining the stub position when the transformer and feed-line impedance Z_C are equal. But second, using a radial indicating line in a similar fashion while tracing along the SWR = 3.0 circle as it intersects the various other resistance circles, we see that for a given SWR the resistance component of the line impedance *E/I* changes with position along the transformer. These two observations together reveal a flexibility available in the approach to a matching design that provides a step toward visualizing the fundamental similarity of the different matching techniques. This flexibility includes the following three conditions, which will be explained in more detail:

1) There is no restriction on the characteristic impedance Z_C of the transformer section T that requires it to be of the same value as the F feed-line impedance Z_C -- it can range from low (coax) to high (open-wire line).

2) The *length* of the transformer section having a given Z_C can be found which will transform the resistance component of the impedance to the value of a matching F feed-line impedance Z_C which differs from the transformer Z_C. However, the transformer section can have a length which is not limited to the distance from the load to the first point at which the resistance component equals the feed-line Z_C. The transformer can extend from the load to either of the two points where the resistance component is seen to equal the feed-line Z_C on the SWR circle, or any electrical length extending beyond these points by an integral multiple of a half wavelength. (We will see later how the use of the Transmatch or an *L* network assists in obtaining the required electrical length, and thus *removes all restrictions from any specified physical length of transformer, ie., from the load all the way to the operating position.)*

3) The action of the stub portion can be performed by any nondissipative reactance of the proper value, whether by a lumped-constant component, or by a separate line section of any reasonable Z_C value which has the proper length to present the required value of reactance. The *electrical* length of the stub is always directly related to its reactance. Now we will see in terms of wave or reflection mechanics how matching obtained by the various techniques recited previously is described by these three parameters: transformer impedance, transformer length, and stub reactive elements.

In our earlier example using the stub technique the magnitude of the reflections appearing at each end of the *transformer* section was the same (0.5, or an SWR of 3:1 for the load at one end and the stub at the other). In other words, the magnitudes of each complementary mismatch were identical. For the present, we will retain the characteristic impedance Z_C = 1.0 for the entire stub-transformer line section, but based on conditions 1 and 2 above, we may change the feed-line impedance as conditions dictate. Consider now the effect of increasing the length of the transformer section and shortening the stub section in accordance with equation 10. For example, while referring to Fig. 7C and the Vector Graph, let us move the feed-line entry point farther away from the load, from $L = 30°$ to $L = 60°$. This increases the transformer length to 60 degrees and reduces the stub length to 26.6 degrees. From observing the radial line extending from the $L = 60°$ point (where $\theta = -120°$) through point *B*, where the SWR = 3.0 and the R = 0.43 circles intersect on the Vector Graph, we see that the corresponding movement along the SWR circle results in a change in the resistance component at the feed point from R = 1.0 to a new resistance R = 0.43.

Recalling from the earlier statement that the complementary mismatch relation holds constant wherever the feed line is attached, a feed line having a characteristic impedance Z_C = 0.43 will be perfectly matched when attached at the $L = 60°$ point. The load-mismatch voltage reflection coefficient is now read as $\bar{\rho}_E = 0.5 \angle -120°$ at point B with the same magnitude as before, but with a larger phase angle because we are farther from the load. And applying the complementary-mismatch principle we see that the voltage reflection from the stub mismatch becomes $\bar{\rho}_E = 0.5 \angle +120°$, as read at point *C*. So we ask the question, how does this new combination produce a canceling wave having the same magnitude as before? We will recall that previously, when the line-characteristic impedances of sections F, S and T were each Z_C = 1.0 as in Figure 7B, the canceling reflection was generated by the stub alone, because no line-junction mismatch existed between the feed line F and the transformer T. But now that the impedance of the feed line differs from the impedance of the transformer, we have an additional discontinuity at the feed point, which also generates a reflection. And the *shorter* (series) stub portion now generates a reflection which is smaller than when all the line sections had a Z_C = 1.0, the stub-reflection magnitude being reduced by the amount of the reflection presently being generated by the feed-line to transformer mismatch. Thus by the complementary mismatch principle the resulting canceling wave still retains the correct magnitude and phase to cancel the load-mismatch reflected wave at this new feed point on the transformer. This canceling wave is evidently generated by the *combined* discontinuities of both the differing line impedances at the junction, and of the stub with the corrected length. We have thus matched a feed line of Z_C = 0.43 to a load of $Z_L = 3 + j0$ through a transformer of Z_C = 1.0.

Using this same line of reasoning we may, conversely, shorten the transformer section and change the stub section according to the tangent relation in Eq. 10 to obtain a match for feed lines of higher impedance. We need merely to position the feed-line entry point where the resistance component of impedance in the transformer T has been transformed to the value of the feed-line impedance Z_C that we wish to use and then adjust the stub length to cancel the line reactance. As explained above, the resistance circle which is intersected by the SWR circle for a given transformer length indicates the feed-point resistance component. This is also the Z_C value of the feed line which will be perfectly matched when attached at the feed point. The data presented in Table 1, taken from points along the SWR = 3.0 circle on the Vector Graph. show a few selected transformer-length examples and pinpoint some of the pertinent information for clarity. Notice especially that the resistance component decreases as the transformer length increases. It is interesting to discover that when the feed point goes beyond the L = 45° position, the θ angles of the voltage- and current-reflection coefficients pass through 90 degrees from opposite directions, respectively. The result of this is that their respective resultants shift 180 degrees. Thus the resultant reflection-coefficient angles interchange, the voltage-coefficient resultant angle θ now becoming 180 degrees and the current angle becoming 0 degrees. This means that the effective reflecting termination at the match point shifts from an open circuit to a short circuit when the feed line is attached more than 45 degrees away from the E_{max} position at L = 0° on the transformer.

Consider now the effect of increasing the length of the transformer section still further, until the reactance component of the line impedance disappears by itself without requiring a stub to cancel it. The Vector Graph shows this condition to occur at L = 90°. At this point the load-mismatch reflected voltage wave is exactly 180 degrees out of phase with the incident wave and therefore no reactance component is developed here. The resistance component of the line impedance at this point is 0.33 x Z_0 on the chart. Based on our present reasoning, we may at this point connect a feed line F having an impedance Z_C = 0.33 (see

Table I - Matching characteristics with various transformer- and stub-section lengths.

Transformer length L°	Stub length S_L°	Resistance component R	Stub reactance jX	Angle of Reflection Coefficient: Voltage: Load mismatch θ°	Voltage: Stub & line mismatch θ°	Current: Load mismatch θ°	Current: Stub & line mismatch θ°
0	0	3.0	0.0	0	0	180	180
10	48.2	2.42	+1.12	-20	+20	+160	-160
22.5	52.9	1.38	+1.32	-45	+45	+135	-135
30	49.0	1.00	+1.15	-60	+60	+120	-120
45	38.7	0.60	+0.80	-90	+90	+90	-90
52	33.0	0.50	+0.65	-104	+104	+76	-76
60	26.6	0.43	+0.50	-120	+120	+60	-60
67.5	19.8	0.38	+0.36	-135	+135	+45	-45
90	0	0.333	0.0	180	180	0	0

Fig. 7D) and obtain a perfect match. No reflections will appear on the 0.33-ohm feed line. How come? Again, because of the canceling reflections in the transformer section! Note the present length of the transformer section -- 90 degrees, or a quarter wavelength. The transformer section alone is now using the *entire length* of the line section, and the stub section has disappeared. Simply by moving the feed point along the transformer to the point where the line reactance vanishes, the resistance component becomes Z_C /SWR = 1/3, and we slip smoothly from the stub form into the series quarter-wave transformer form of matching. (See Table 1.) Remember, the characteristic impedance of the transformer is still $Z_C = 1.0$, which has become the geometric mean between the input and output impedances that it is matching $\sqrt{0.33 \times 3.0} = 1.0$. Looking from inside the transformer, the impedance level at the input terminals is stepped down 3:1 (giving us short-circuit reflection behavior), just as the output impedance (load) is stepped up 1:3 (for an open-circuit behavior at the output terminals). It is therefore evident that a λ/4-transformer section of line having a Z_C equal to the geometric mean of its two end-terminal impedances has equal mismatches at both ends, and thus produces reflections of equal magnitude at both ends. These reflections from each end cancel each other out at the feed-line transformer-input junction, because waves reflected at the output mismatch return to the input junction exactly 180 degrees out of phase with the waves reflected at the input mismatch. This is because the load-mismatch reflected wave has traveled 90 degrees from the input point to the load mismatch, and an additional 90 degrees in returning to the input. To clarify further what is happening here, we recall that previously, when the feedline F and transformer T were of equal impedance Z_C (Fig. 7B), the canceling reflected wave was generated entirely by the stub mismatch. In the present case where the stub length is zero, the canceling reflected wave is generated entirely by the 3:1 feed-line transformer-junction mismatch. The voltage reflection coefficient angle of this feed-line junction-mismatch reflected wave is $\theta = 0°$ referenced to the feed-line Z_C because the transformer Z_C is three times greater. However, after both the load- and input-junction-mismatch reflected waves have joined to cancel one another, the input-junction reflection no longer travels toward the generator, but is rereflected into the transformer toward the load in the same manner as with the previous stub-reflected wave. Therefore, referenced from within the transformer, the reflection coefficient angle of the junction reflection is $\theta = 180°$, as indicated in Table 1. In a later section we will see why the λ/4 series transformer is an impedance inverter for any complex terminating load, and is not restricted to purely resistive loads.

Part 5 - Low SWR for the Wrong Reasons

IN PART 1 of this series of articles the statement was made that misconceptions concerning SWR and reflections are rampant among amateurs, both in print and on the air. So to reiterate further, this series has been written with one primary goal in view -- to identify some of the misconceptions and to provide correct answers in the hope of clarifying some of the confusion resulting from the misconceptions. One outstanding area of confusion concerns the nature of reflected power and how it is accounted for in the circuit. In short, is it *real* or is it *fictitious*, and where does it go? Now the nature of reflected power was discussed in Part 3, where it was shown why reflected power is real power. And in Part 4 we delved into the question of where the reflected power goes, as the role of reflections in conjugate matching was discussed. There the stub form of matching was used to illustrate the wave action which accomplishes the matching function and which also derives the total incident power from the combined source and reflected power. We will recall that learning of this wave action stripped away the mystery of how a mismatched load can *absorb all of the power delivered by the source*. We learned this as we saw how the reflected power adds to the source power at the conjugate match point so that the reflected power can be subtracted from the total, enlarged, incident power at the mismatch point to leave a net power in the load equal to the source power.

Now that we have established this relationship between the source, reflected, and incident powers in terms of the wave mechanics of the conjugate match, we have the necessary background and tools for identifying some of the improper usage of SWR, and for clarifying in greater detail the reasons for the misunderstanding that still prevails concerning what happens to the power reflected from an antenna that is mismatched to its feed line. Further clarification of the misconceptions will enhance the appreciation of the mismatched feed line as simply an impedance-transforming device, particularly as we see somewhat later how the Transmatch type of feed-line matching network and the pi-network tank circuit of the transmitter perform the conjugate matching function in the same manner as the stub. Additional perspective in relating the discussion to practical feed-line operation will be gained as some of the thoughts presented in Parts 1 and 2 of this series are now expanded.

If it appeared to some readers that the importance of SWR was overly minimized or downgraded in the treatment accorded it in Part 1, it was not so intended. The intent there was to focus attention on the importance of understanding the subject of reflection and SWR correctly and in such depth that we may retain complete control over them in our antenna system design engineering. Thus, instead of letting SWR become king to take control and deprive us of a breadth and flexibility, we may use SWR in the system design choices in ways which many are unaware exist.

How many of us have acquiesced to the King in pruning an 80-meter dipole, with great pains to obtain the best possible match to a half-wavelength feed line at a specific frequency, but cannot operate more than a few kHz from that frequency without fear of the King's apparent dire consequences? But how many are aware that King SWR can be outwitted and his consequences averted without pruning either the dipole or the feed line? And how many have been aware that the matching operation can be performed at the transmitter end of the line at any frequency within the entire 75-80 meter band without suffering any significant loss in power in spite of the SWR remaining on the feed line? Although it contradicts the word published in many articles during the past two decades, this revelation is true, and is indicative of the flexibility or freedom that really is available in our choice of antenna systems designed for all the hf bands, simply by having a better understanding of SWR and reflection.

Valid Reasons for Low SWR

There are good and valid reasons for being concerned with SWR and reflection, from both the amateur and commercial viewpoints -- with this there can be no argument. As we well know, these reasons are concerned basically with voltage breakdown and power-handling capability, efficiency and losses, and with line-input impedance as it relates to transmitter output coupling. In amateur practice, power-handling capability and voltage breakdown don't become serious problems unless we try to shove the legal limit of power through RG-58/U or RG-59/U at a high SWR. Losses and efficiency concern us, but to a much smaller degree than is generally realized, and for a different reason than many are aware, as will be shown very shortly.

The chief reason why the amateur should be concerned (but not alarmed) with SWR is in its relation to line-input impedance and transmitter coupling. This will be discussed in great detail in a later section. There we will see how to tame impedance and coupling for any reasonable value of SWR, and in that discussion the relative *unimportance* of having a resonant antenna will also become evident. But it is of great importance that we first clarify some of the prevalent misunderstandings of SWR and reflected power, because they are causing many amateurs to strive for a low SWR for wrong, invalid reasons, and often needlessly. Probably the most serious and widespread misconception concerning SWR prevailing throughout the amateur fraternity is the erroneous notion that there is a direct one-for-one relationship between reduction in reflected power and a resulting, increase in radiated power. In other words, every decreased watt of reflected power is thought to provide an additional watt of increased output. Not So, but the tremendous number of amateurs who have been misled to believe this invalid, unscientific, and untenable premise is simply unbelievable.

Another related concept, popular, but also erroneous, is that when terminated in a mismatch, the coaxial feed line becomes part of the radiator, causing radiation from the feed line due to the standing wave (ref. Part 2 of this series, statement 18). This is untrue because the line voltages and currents, and the standing waves resulting from the mismatch, are entirely contained in and between the outer and inner conductors, inside the coax. *No standing wave develops on the outside because of mismatch.* However, feed line radiation may result from standing waves on the outside of the coax caused by current unbalance if a balanced dipole is fed with coax and no balun is used. This feed-line radiation may or may not be of any consequence, but the topic is covered well by McCoy *(ref. 45).*

Misunderstanding of how the benefits accrue, from a low SWR, and of just how little benefit is obtained, is driving many of us to attain SWR values far lower than where the benefits continue to be significant in relation to the efforts expended to attain them. It is for this reason that we often set an unrealistically low limit on SWR that needlessly restricts the operating bandwidth, or range of usable frequencies on either side of the antenna resonant frequency, to a far more limited range than is necessary. In rectifying a misunderstanding such as this, it often helps to learn first how the misunderstanding originated.

"Impedance" Bridges

One aspect of the misunderstanding has been created to a large extent by narrow and often erroneous interpretations of matching principles found in various instructions for instruments such as noise bridges[2 3] and the antenna-scope for determining the terminal "impedance" of an antenna. Contrary to what is stated in some of the instructions, these devices cannot measure *impedance* -- they can measure *resistance only* -- and then *only in the absence of reactance.* (Suggestion: Look up and compare the definitions of impedance and resistance; the term impedance is often misused when the correct term should be resistance. The reader is also invited to see *ref. 46).*

Consequently, in using these devices we have been coerced into finding only the resistance component of the antenna terminal impedance, and only at the resonant frequency of the antenna, because this is the only frequency where the impedance has zero reactance, or R + *j*0.

In following this tack, erroneous emphasis has been given to requiring the antenna radiator itself to be resonant, thus nurturing the misconception that it needs to be resonant to radiate all the power being supplied to it.[24, 25] Thus, many have been misled to believe that the antenna just won't perform properly at any frequency except the resonant frequency. (See Part 2 of this series, p. 21, statements 5, 6 and 7 and *refs. 20, 21, and 24.)* In addition, emphasis on the further necessity for obtaining a resistance component reading equal to the line impedance Z_C has in many cases caused us to go to extreme lengths, such as adjusting the antenna height above ground in small increments to achieve that exact resistance reading in quest of the perfect 1.0 match.[26] (See Part 2 of this series, p.22, statement 15.) Adjusting heights in large increments to obtain control of radiation in the vertical plane is realistic. But controlling radiation resistance by adjusting the height is neither necessary, realistic, nor practical, because the efficiency thought to be gained through this action is illusory.

The truth of this will become evident somewhat later as we see why there is no justification whatever for expending any matching effort *at the load* to improve a mismatch of 2:1 or less, simply to remove the standing wave with the expectation of improving efficiency. Furthermore, because of the reactance that appears as we depart from the resonant frequency, the sacred but overrated perfect match found at some carefully adjusted height can be obtained at only one frequency without retrimming the radiator length, thus continuing the vicious cycle. However, the widespread practice of this philosophy in antenna-system operation has completely conditioned us to think only in terms of using a $\lambda/2$ transmission line with no reflection, and to obtain its perfect 50-ohm nonreactive input impedance by operating only at the resonant frequency. So we have, in effect, been deterred from learning of the real effect of reactance in antenna impedance, and of how the line transforms any terminating impedance in a straightforward and predictable manner.

In becoming so conditioned, many of us have forgotten that we can obtain the desired 50-ohm nonreactive input impedance from the line-transformed antenna impedance with a simple line-input matching network in the shack, often more easily than it can be obtained at the antenna. In fact, in some transmitters the impedance seen by the transmitter at the line input for SWR values of 2:1 or higher can be matched for optimum loading by adjustment of the transmitter tank circuit itself. If a transmitter does not contain sufficient matching range, a separate line-matching network between the transmitter and line input offers a more judicious matching arrangement than playing games out at the antenna. We will see why there are many situations where this same matching approach should be considered when the load mismatch yields SWRs of even 5:1 or higher, as one departs from the self-resonant frequency of the radiator *(ref. 24).*

One further misconception exists that has also resulted in needless and unwarranted reliance on the $\lambda/2$ feed line to repeat the resonant antenna resistance at the transmitter. This one concerns the effect of line-input reactance on tank-circuit resonance when the line with reflections is fed directly by the pi network. Consider a tank circuit which is first loaded and tuned to resonance with a resistive load, and then when the load is changed to one containing reactance. If the tank components have sufficient retuning range to compensate for the reflected reactance and return the circuit to resonance at the proper load level, all is well; the tubes still see a proper resistive load as before. The misconception about this point has been generated by some writers who apparently don't understand resonant circuits, for they proclaim that the retuning introduces reactance that detunes the circuit, causing improper loading, and increases plate current and dissipation. Not so -- much more detail on this point will appear in a subsequent part.

Low SWR for the Wrong Reasons

We have discussed "low SWR for the wrong reason," as practiced (often unwittingly) in using the perfectly matched antenna operated only at the self-resonant frequency of the radiating element. But another wrong reason for desiring a low SWR is interpreting feed-line SWR as the sole criterion for indicating the quality of an antenna's radiating performance across a band of frequencies, with low SWR across the band getting the raves and high SWR getting the boos. This is a definite misuse of SWR, because there are cases where the low and high SWRs occur in just the opposite relation, with respect to indicating antenna efficiency over a given bandwidth, for reasons which will be explained shortly. As a result of this misuse of SWR, *good* antennas are too frequently rejected as "bad" because the feed-line SWR swings relatively high, and *poor* antennas are accepted as "good" when the SWR remains relatively low.

In most cases the use of feed-line SWR alone to indicate antenna efficiency is completely invalid, because SWR indicates only the *degree of mismatch*, not efficiency. However, we will see presently how a relative change in SWR, to a value either lower or higher than a previous value known to be correct in a given antenna system, can indicate that a change has occurred somewhere in the system. That change may affect its radiating efficiency. The popular vertical antenna having from two to four ground radials (an insufficient number for efficient operation), or perhaps having only a buried water pipe or a driven rod for a ground terminal, is one case where lower-than-normal SWR obtained over a frequency range indicates a poor quality of radiating efficiency, rather than a good one. But conversely, improving the ground system by adding a sufficient numberof radials can increase the radiating efficiency to nearly 100 percent, and this improvement will be accompanied by a significant increase in SWR readings over the same frequency range to higher values, which are the normal or expected values.

With an adequate ground system, the SWR is predictable over the frequency range, because a load impedance of any specific R + *j*X value yields an exact SWR on a given feed line, and because we know approximately what the antenna impedance should be at whatever frequency we may wish to use *(refs. 4 7, 48, 49, and 58, p. 3-1)*. But when the ground system is inadequate there is an unknown ground-loss resistance added to the known antenna impedance, which changes the SWR to some lower, unpredictable value. Yet, without being aware of these facts, we often tend to be happier in the discovery of an unsubstantiated low SWR than we do in determining whether we have SWR values that *should* be obtained with the existing configuration. This is a very important concept that requires a clear understanding if we are to avoid misinterpretation of SWR data in our effort to optimize radiated power.

It will help in understanding this concept if we have a clear physical picture of how the ground-loss resistance develops. It appears that we have still another misconception here, this one concerning the current and field behavior in the vertical-over-ground antenna system. Most of us know that conventional grounding techniques used for lightning protection, such as rods or pipes driven deeply into the ground, provide an excellent low-resistance current path for the lightning current. Many are unaware, however, that these techniques are totally inadequate for conducting the entirely different pattern of current flow of the vertical antenna system.

Vertical Radiator over Earth

Let us digress a moment for a brief look into the field and current behavior of the vertical antenna system, to see what type of ground system it takes to meet the current-pattern requirements. Consider a base-fed vertical antenna; one terminal of a generator is connected to the base of the vertical radiator and the other generator terminal is connected to ground, just below the base of the radiator. During the half cycle in which the conduction current in the antenna radiator flows upward, all the current returns to ground through displacement currents, which follow the lines of force in the rf electric field through the radiator-to-ground capacitance. See Fig. 8. The electric field surrounding the antenna, which excites the displacement currents, fills the entire volume of space surrounding the antenna in the shape of an oblate or somewhat squashed hemisphere. This hemisphere intersects the ground to form an imaginary circle having a radius of slightly over 0.4λ for radiators of $\lambda/4$ in physical height. (The radius decreases as the physical height of the radiator decreases.) The displacement currents enter the ground *everywhere over the entire surface within the circle* and then flow back *radially* to reach the grounded generator terminal. Although some of the current penetrates somewhat more deeply, most of the flow at frequencies above 3

MHz is restricted by skin effect to the upper few inches of the ground.

Now a ground system comprising only a simple water pipe or a driven rod or two is simply a terminal -- the ground-feed terminal of the antenna system. So all the returning currents must flow entirely through the poorly conducting ground from all directions everywhere within the circle to reach the terminal. This ground system is often measured to have an "acceptably low" resistance *at dc* (which may be satisfactory for lightning protection), but it injects a series loss resistance in the antenna circuit *at rf*. The rf resistance often exceeds the radiation resistance of the antenna itself! Adding two to four wire radials to the system will provide good conductivity toward the ground terminal for the currents which reach those radials, but only a tiny amount of the total current entering the surface inside the circle is intercepted by the radials. Thus, all the remaining currents *still* flow only through the lossy ground, and the result is that we still have a high loss resistance.

Now if a sufficient number of equally spaced radials (90 to 100) extending out to 0.4λ are present to intercept all the currents, all the returning displacement currents find highly conductive paths everywhere within the circle, which lead the currents through negligible loss resistance directly back to the ground terminal of the generator. This can be visualized by examining Fig. 8. Currents which do enter the ground between the closely spaced radials quickly diffract to a radial wire, and thus travel only a short distance through lossy earth before reaching a good conductive path. Thus, with sufficient radials, we have a nearly perfect ground system which adds only a negligible amount of resistance to the true antenna impedance measurable between the radiator base and ground *terminals (refs. 20; 50; 51 and 57, pp. 115-124)*. From this we can see why the lightning-type ground system, although in prevalent use, is unsatisfactory for an efficient antenna system *(ref. 57, p. 82).*

Now we are not suggesting that λ/4 antennas with less than ideal ground systems should not be used, nor that fair results cannot be obtained without their use. But the difference between no or few radials (3 or 4), compared to 100, can amount to over 3 dB. This is far in excess of the loss resulting from an SWR of 4:1 or 5:1 on the average coaxial feed line used by amateurs. The

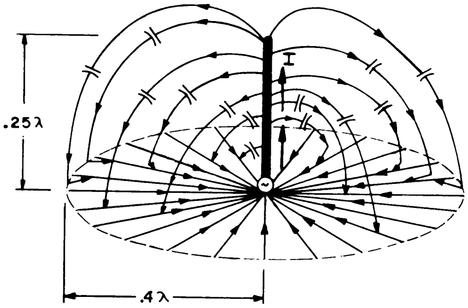


Fig. 8 - The hemisphere of current which flows as a result of capacitance of a λ/4 vertical radiator to the earth or a radial system. At frequencies above 3 MHz, rf currents flow primarily in the top few inches of soil, as explained in the text. Ground rods are of little value at these frequencies, and spikes or large nails are sufficient to secure the outside end of each radial wire. With a sufficient number of radials, annular wires inter-connecting the radials offer no improvement in antenna efficiency, as the current path is radial in nature.

point being emphasized here is that the value of ground resistance is unknown and unpredictable in systems using less than an adequate number of radials. This makes the resulting SWR readings unpredictable and therefore useless for the purpose of evaluating the *absolute* quality of the system, unless some means is available for determining what the change in SWR would be if the loss resistance could be switched in or out.

In practical amateur installations, the ground resistance will be sufficiently low if only 40 to 50 radials are used with a λ/4 radiator. The small improvement in radiated power for the addition of still another 40 or 50 radials *with the* λ/4 *radiator* will probably not justify the extra cost and effort. However, if a *short* vertical antenna (from λ/8 or less to λ/4) is contemplated, it should be remembered that the radiation and terminal resistances decrease as the radiator is shortened. The ground resistance now becomes a larger part of the total resistance, decreasing the efficiency. Thus the ground resistance should be kept as low as possible for the full capability of the short antenna to be realized *(refs. 51; 56; 57, pp. 18-29).* There is practically no difference between the radiation capabilities of the λ/4 antenna and a radiator even shorter than λ/8, except for the effect of ground resistance and the loss in the resistance of the coil used to cancel the capacitive reactance in the terminal impedance of the shortened antenna. The professional literature is replete with references confirming this point *(refs. 20, 24, 52, and 53).*

Resistive Losses and SWR

In this section we will see how *any* additional resistive losses that are separable from the true antenna load impedance affect the true load SWR. By separable are meant such losses as ground-loss resistance, corroded connectors and other poor connections, cold-solder joints, and so on. These all contribute loss resistances that we can control or reduce. In contrast is the resistive component of the antenna terminal impedance, which comprises both the radiation resistance and the inherent conductor-loss resistance in the radiating element. In most cases the conductor-loss resistance in practical radiating elements is negligible, unless excessively small wire is used.

There are several useful relationships between load impedance Z = R + *j*X, line impedance, Z_C, and SWR. For example, it is well known that when the load impedance is a pure resistance R, equal to the line impedance Z_C, the reflection coefficient ρ is zero, and the standing-wave ratio is thus one to one. But the reflection is no longer zero and the SWR becomes equal to the ratio R/ Z_C when the resistance is larger than Z_C, or Z_C /R when the resistance is smaller than Z_C. It is also well known that ρ and SWR increase with the addition of any reactance component in the load impedance that increases the total reactance, whatever the resistive component may be (see Eq. 1, Part 3 of this series). And as noted previously, any combination of R + *j*X yields an exact value of SWR, when terminating a line of given impedance Z_C. We also know that the reactance, X, appearing in the impedance at the terminals of an antenna contributes more to the rise in SWR at frequencies away from the antenna resonant frequency than does the change in resistance. This is because the reactance changes more rapidly than the resistance during the change in frequency *(ref. 58, p. 3-1).*

There is, however, an interesting relationship which is not generally well known between the resistance and reactance components of a load impedance. This relationship sheds light on how the two components affect mismatch reflection and SWR, and also explains why the unknown ground resistance and other losses mentioned above reduce the usefulness of SWR readings. When reactance is present in the load impedance, the minimum possible SWR occurs when the resistance R is greater than Z_C. The value of the resistance that yields the lowest SWR in combination with a given value of reactance in the load (which we will call the *minimum-SWR resistance*) is dependent solely on the reactance present. This value may be obtained from the relationship:

$$r = \sqrt{x^2 + 1} \quad \text{(Eq. 11)}$$

where *r* is the minimum-SWR resistance and *x* is the reactance present in the load, with both values normalized to the system Z_C. (See footnote 22, Part 4 of this series regarding normalized impedances. Also see Feedback, p. 46 of QST for Nov., 1973.) It can be seen from Eq. 11 that when *x* becomes zero, *r* = 1, for an SWR of 1:1, but it is interesting to know that the resulting SWR always equals exactly the arithmetic sum of the minimum SWR resistance value *r*, and the reactance value, *x*. This latter relationship will help us to understand how unwanted loss resistance, separable from the true antenna-load impedance, affects SWR. In the case of the vertical radiator over earth, these unpredictable losses change the SWR from a predictable value, based on known available antenna impedance data *(ref. 57, p. 82),* to some unpredictable and usually lower value. A general application of the relationship is presented in the following statements. When the resistance component of the true load impedance is lower than the minimum-SWR resistance, as determined for any reactance component also present in the true load, adding of resistance separate from the true load impedance will cause the SWR to decrease from the value obtained with the true load. This is true until the total resistance is equal to the minimum-SWR resistance. Further addition of resistance will cause the SWR to rise again. These statements apply especially to the vertical antenna of λ/4 heights or less in proving why ground resistance which reduces the efficiency also reduces the SWR. This is because the true antenna resistance component, R, is generally less than the impedance, Z_C, of normally used feed lines, while the minimum-SWR resistance R is always equal to or greater than Z_C.

The effect of reactance in the antenna impedance raises an additional factor of importance in understanding the relationship between SWR values and antenna performance. As stated earlier, the rate at which SWR rises as the operating frequency departs from the resonant frequency of the antenna depends on the resulting change in the impedance at the antenna terminals, which in turn is dependent on the *Q* of the antenna. One factor that has a primary influence on antenna *Q* is the amount of capacitance between the opposite halves of the dipole. (Although it is more commonly called a monopole, a vertical antenna over ground can also be considered as a dipole, because the lower half is simply the image of the upper half, with the opposite polarity.) This dipole capacitance is determined by the ratio of the radiator length, L, to its diameter, D.

The L/D ratio *(refs. 1; 58, p. 3-1)* found in the usual simple thin-wire dipole is very high, resulting in a low dipole capacitance and high *Q*, causing a rapid change in impedance, reflection, and SWR as frequency changes. This is why a thin-wire dipole is considered a narrow-band device. However, specific broad-banding steps may be taken to increase the dipole capacitance and thus reduce the *Q* and thereby the rate of change of SWR. One such step, for example, is decreasing the L/D ratio by using a multiwire cage configuration for each dipole half, or by fanning out multiple wires from the feed point. SWR curves vs. frequency are valid here in comparing bandwidths obtained while experimenting with different radiator configurations. However, any separable loss resistance must now be either minimized or held constant to prevent it from introducing unknown variables. Otherwise, the unknown variables can caused *differing errors* in the SWR readings obtained with different configurations, and thus render the results of the experiment invalid. But unless actual broad-banding steps have been taken to reduce the *Q*, the rate of change in SWR as frequency changes will not differ dramatically between various types of dipoles having roughly equivalent *Q* values. (These types include the so-called inverted V.) If a dramatic difference is noted with no *valid* broadbanding steps taken, troubleshooting is called for to determine the cause. More than likely some unwanted loss resistance will be flushed out, if this is the case.

The writer has seen SWR curves published, along with descriptions of quite simple antennas, where it would have been impossible for the SWR to remain as low as indicated over the frequency range shown; the antenna *Q* of the configuration presented would have simply been too high. Two possible explanations for this sort of contradiction are that (1) perhaps the readings were obtained using an inaccurate SWR indicator -- many read on the very low side *(refs. 40, 54),* or (2) as suggested above, an unrecognized trouble existed somewhere in the antenna system which was lowering the *Q* by means of a separable loss resistance. Yet these articles were published because the antennas they described were purported to have "'improved SWR characteristics." How many times have you heard someone praise his newly hung skywire by simply telling how low the SWR indicator reads across an entire band? It should now be clear that it cannot be emphasized too strongly that an unrecognized and unwanted loss resistance is an antenna system can *cause* a low SWR reading *when it should not he low!* So in a later section we will explore the relationship between antenna impedance and SWR in detail so that we may determine what is a proper SWR for given conditions.

Reflected Power and SWR

Let us now return to the subject of why we worship low SWR for a wrong and invalid reason. As stated earlier, the misunderstanding of this aspect of reflected power is based primarily on the prevalent, but erroneous, idea that any reduction in SWR or reflected power effected on a line feeding an antenna results in a direct one-for-one increase in radiated power. The erroneous reasoning in this idea is in the assumption that if the power is being reflected, it therefore cannot be absorbed in the load or radiated, and that the power which is reflected returns to be lost by dissipation in the transmitter. The assumption is false on both points, for the truth is, because of the reflective conditions in the circuitry used in coupling the transmitter to the line, *all power that enters* the line is absorbed by the load (except that dissipated in the line itself due to attenuation). This is *true even when the load is not matched* to the line impedance *(ref. 55)*. Complete absorption in the mismatched load (and line) of all the power delivered by the transmitter is obtained, because the power reflected from the mismatch is conserved and returned to the load by rereflection from the line-coupling or matching circuitry, in accordance with the principles discussed in Part 4.

Let us consider a lossless line for a moment, in light of the above statement; here it is axiomatic that if all power delivered to the line is already being absorbed in the load (because none can be absorbed in a lossless line), a reduction of the reflected power cannot have any effect whatever on the amount of power taken by the load. And obviously, there is no power left over to be dissipated in the transmitter. Following this same reasoning in a real line having attenuation, *all losses in powe*r must be attributed to the basic I^2R and E^2/R losses arising from the line resistance or attenuation. These losses are unavoidable, even when the load is perfectly matched. The only *additional power losses* which can be attributed to SWR or reflection occur because the same resistive attenuation is encountered by the reflected power wave as it travels along the line from the load to the input. The amount of power lost in this manner is very small, indeed, at frequencies in the hf range when good-quality low-loss line is used because, during its return to the input, the reflected power suffers only the same *rate* of line-attenuation loss (in dB) as the incident power suffers in its forward travel toward the load. And as previously stated, all the reflected power which arrives back at the input now becomes part of the incident power. Another way of explaining the relation between SWR and lost power is to recall from Parts 3 and 4 that because the incident power is the sum of the source and reflected powers, the incident power is greater than the source power wherever the SWR is greater than 1.0. Thus for a given source power, the resistive losses are somewhat higher in the portion of line where the incident power is higher than the source power, simply because the *average* line current *I* and voltage *E* are higher in that portion.

So from this discussion concerning improper usage of SWR we learn that from the viewpoint of efficiency, our concern for SWR involves only the loss due to line attenuation. Thus we can tolerate a higher SWR when the attenuation is low, but when attenuation is high the SWR limit must be lower for the same amount of additional power lost from SWR. The exact relation between SWR and the power loss caused by SWR for different values of line attenuation is shown graphically in Fig. 1, Part 1, taken from the ARRL *Handbook* and the *Antenna Book*. From this figure we can easily see that the amount of power actually lost is in sharp contrast to the amount mistakenly assumed to be lost in the improper concept of SWR, where it is thought that a reduction in SWR or reflected power results in a direct equivalent decrease in the amount of power lost in the system.

There is a great deal of irony behind these various misunderstandings of reflections that have engendered the wrong interpretation or usage of SWR. The irony is that the *correct* reasons why SWR should be considered, as previously recited, are frequently overlooked in the wrong usage, while the basis so generally accepted in support of the wrong usage doesn't even exist in the coupling methods used by amateurs to transfer power from the transmitter to the antenna. A part of this obtuse logic originated from the confusion among both amateurs and engineers in the meaning of a " matched generator" -- to some it implies being matched in *only one direction*, and to others it means being matched in *both directions*. In transmitter operation, where conjugate coupling is usually used to deliver optimum power to a load through a line, the match is in one direction only -- forward.

Part 6 - Low SWR for the Wrong Reasons (continued)

PART 5 OF THIS SERIES concluded with the statement that in transmitter operation, where conjugate coupling is usually used to deliver optimum power to a load through a line, the match is in one direction only -- *forward*. The generator (transmitter) is matched to the line, but, looking back into the generator coupling circuitry during all times that the generator is actively supplying power through the conjugate coupling to the line, the line is totally mismatched. 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 + *j*X looking in one direction, and the equal but opposite-sign impedance R - *j*X in the opposite direction. (The net reactance of zero obtained from these two impedances proves the *system* is resonant!) But 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. In this case the impedance now terminating the generator end of the line will be seen as a dissipative load while measuring impedance in the generator direction. (See footnote 5, Part 2 of this series.)

The fact that dissipation occurs in the impedance which *replaces* the generator impedance during these measurements is largely responsible for the erroneous inference that power reflected in the generator direction is also dissipated in a similar manner in the generator impedance. However, when the generator is active, its internal impedance is never seen as a load for power reflected from a mismatched load terminating the line because of the interaction between the source wave, the load-reflected wave, and the canceling wave, as described in detail in Part 4. The fine is thus totally mismatched looking in the generator direction.

In laboratory work, on the other hand, the generator is usually matched in both directions. Here the generator is isolated from the line with a resistive pad, or attenuator, having an insertion loss of about 20 dB, and having the same impedance as the generator output and the line Z_C. In this case the generator sees a match looking into the line-terminated pad, and the line also sees a match looking back 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, the reflected power reaching the source is about 40 dB below, or 1/10,000 of the power delivered by the generator, when the load is a totally reflecting short- or open-circuit termination, and even less with practical terminations that are dissipative. This amount of reflected power reaching the generator is negligible from the viewpoint of adding to the source power and modifying the line-input impedance. Thus, for all practical purposes, the pad appears to the generator as either an infinitely long line, or a line having a perfect R = Z_C termination, and both a constant generator loading and constant incident voltage are maintained as line loading is changed, to satisfy laboratory requirements. (See Part 1 of this series, p. 37, last paragraph, and *ref. 19, pp. 7 and 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.

Reflected Versus "Lost" Power

This erroneous concept of reflected power 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 K8ZVF, and the other, an SWR-indicator review by W2AEF, 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 which two important ingredients that have been overlooked for a long time will be revealed. In so doing, we will see not only why statements concerning lost power published in the two articles mentioned above are incorrect, but also 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 have been misled into seeking low SWR for the wrong reason.

Assume a lossless transmission line having a perfectly matched load termination. Assume also a matched condition between the generator or transmitter and the line characteristic impedance, Z_C. With these conditions there is no reflected power in the line and therefore no *reflection loss*. 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 the 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 $Z = E/I$ all along the line, as stated in Part 3, and as shown for an SWR = 3.0 in Fig. 4. When the reflected wave reaches the input terminals of the line the generator is presented with a change in line input impedance from the Zc value to some new value determined by the *E/I*-vector relationship 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 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.

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 can be referred back along the line to the generator. Thus, reflection loss is simply a non-dissipative type of loss representing only the *unavailability* of power to the load due to the generator's making less power available to the line as a result of the mismatch of impedances caused originally by the mismatch at the load terminating the line. (That reflection loss represents only the unavailability of power to the load will become evident as it is now shown 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 is totally rereflected toward the load, adding to the source power exactly the same amount as the reduction in power made available by the generator. Since incident power equals source power plus reflected power, the incident power reaching the mismatched load remains the same as before the generator made less power available. The reflection loss therefore now equals the reduction in generator power.

If a conjugate match is now provided anywhere along the line, even at the input terminals, the reflected-power wave is prevented from traveling past the match point toward the generator, as explained in Part 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. Also, the generator no longer sees a mismatch, and again delivers its *maximum-available* matched power to the line. The conjugate match has thus provided a negative reflection, commonly called "reflection gain," which exactly equals and cancels the reflection loss. But it has also been shown that 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*; Part 2 para. 3, statement 9; Part 4, para. 2.)

So we now ask the question, how does this situation relate to the K8ZVF nomograph, where reflected power is stated to be "lost power," and to the "useful power" table from the Knight SWR indicator review? (See footnotes 27 and 28.) It is this: The 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 Part 3, ρ^2 is the measure of reflection loss or power *reflected*, which, as stated above, equals the reduction in power made available from the transmitter, calculated directly from the mismatch between the line Z_C and the load impedance Z_L. The reflected power is the square of the voltage- or current-reflection coefficient, p, from Eq. 1, Part 3 (see also Fig. 4), but remember further that it is a

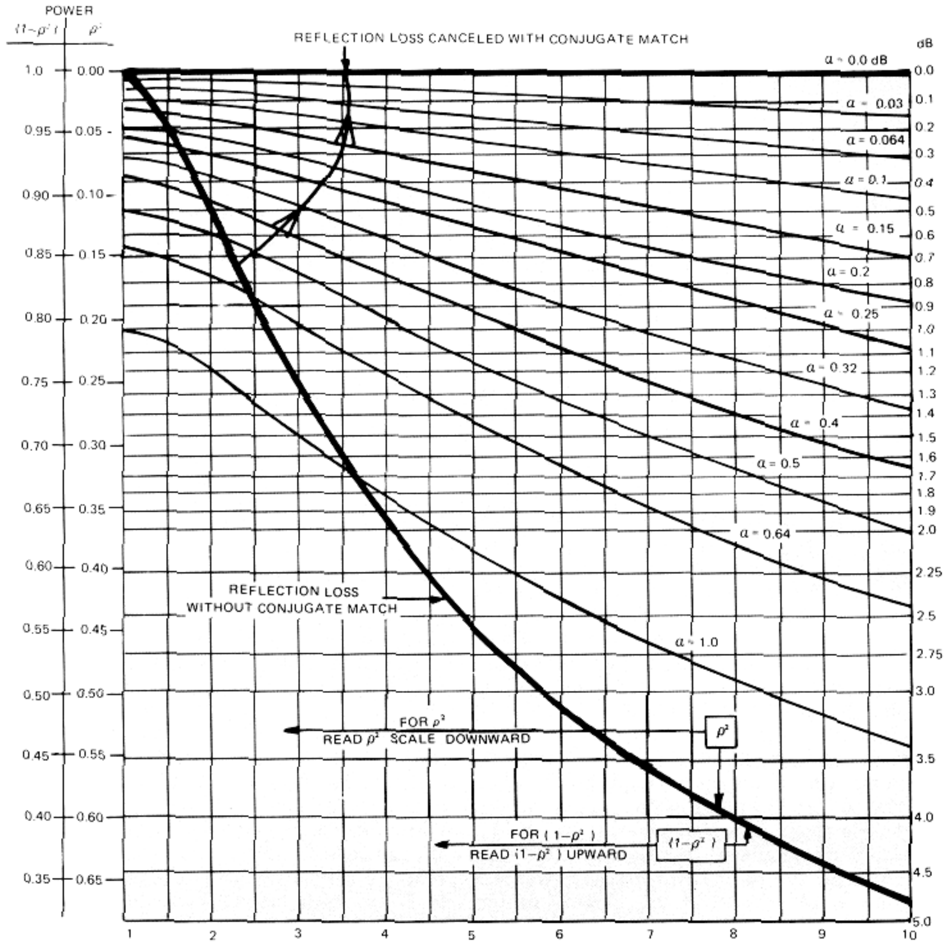


Fig. 9 - Reflection loss versus SWR and matched-line loss of rf transmission lines. Total attenuation in a line operating with SWR may be determined from the dB scale at the right of the chart. The calibration scales at the left are discussed in the text. The a curves in the body of the chart represent the matched-line loss for a particular length of line at a particular frequency. For example, the following types and lengths of line would exhibit the a attenuation factors indicated. Each of these examples is for a frequency of 4 MHz: a = 0.03 dB - 100' of No. 12 open-wire line; a = 0.064 dB 20' of RG-8/U; a = 0.1 dB - 100' of Amphenol Twin-Lead, No. 214-022; a = 0.2 dB - 62-1/2' of RG-8/U1; a = 0.32 dB - 50' of RG-59/U, 100' of RG-8/U, or 200' of RG-17/U; a = 0.5 dB - 87'' of RG-59/U or 175' of RG-8/U; a = 0.64 dB - 100' of RG-59/U or 200' of RG-8/U; a = 1.0 dB - 119' of RG-58/U, 350' of RG-8/U, or 700' of RG-17/U. The curves are plots of the following expressions:

	INCIDENT, OR FORWARD POWER (multiply by source power delivered)			POWER REFLECTED		POWER ABSORBED	
	POWER AT CONJUGATE MATCH POINT	POWER AT LOAD (after line attenuation)					
	$\frac{1}{1-\rho^2 e^{-4a}}$	$\frac{e^{-2a}}{1-\rho^2 e^{-4a}}$	$-$	$\frac{\rho^2 e^{-2a}}{1-\rho^2 e^{-4a}}$	$=$	$\frac{\left(1-\rho^2\right)e^{-2a}}{1-\rho^2 e^{-4a}}$	(Eq. 12)
with lossless line (a = 0)	$\frac{1}{1-\rho^2}$	$\frac{1}{1-\rho^2}$	$-$	$\frac{\rho^2}{1-\rho^2}$	$=$	1	(Eq. 13)

where ρ = magnitude of voltage-reflection coefficient (see Part 3, para. 1)

a = line attenuation in nepers (= $\frac{\text{dB}}{8.686}$)

e = 2.71828, the base of natural logarithms

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

nondissipative power, because it all eventually reaches the load, as explained in Parts 4 and 5.

The tabularized data in the SWR-indicator review article correctly lists percentage of reflected power ρ^2 for corresponding values of SWR. But 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, this 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 matching the transmitter to the new load and consequently establishing the reflection gain $\left(\frac{1}{1-\rho^2}-1\right)$ 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 rereflected at the generator terminals, either with or *without* the reflection gain.

It is now evident that the information presented by K8ZVF and W2AEF is not specifying "lost" power at all, but only the *reflection loss* the amount of power made unavailable by the transmitter until the conjugate match provides the reflection gain which cancels the loss and permits the transmitter to deliver its maximum-available matched power. And as stated on several previous occasions, the conjugate match is automatically attained (sometimes unwittingly) either by proper tuning of the transmitter tank circuit to the line impedance *E/I*, or (sometimes knowingly) by use of a line-matching network if the transmitter tank lacks sufficient range to obtain the match by itself. How the tank performs the conjugate match, and the effects of undercoupling, overcoupling, and possible reactive loading of the tank which can result in the absence of the conjugate match will be explained in detail later in this series.

Reflection Gain

Now refer to Fig. 9 (previous page). This figure was developed to illustrate the *reflection-gain* concept, 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 of us wrongly believe that "getting the SWR down" is the most important factor in "getting the power into the antenna." We fail to realize that, *whatever the SWR*, with a low-loss feed line, the reflection gain has canceled the effect of the load mismatch if the transmitter can be made to tune and load properly into the line, and all the transmitter power is already being taken by the antenna. And therefore, as explained previously in Part 5, we have also been unaware that no more power of any significance reaches the antenna in achieving the lower SWR. We have also been unaware that line attenuation is the key to whether the SWR level has any *practical* impact on efficiency at all (ref. Part I of this series, p. 38).

In Fig. 9 the heavy curve marked ρ^2 and $(1 - \rho^2)$ is based on lossless-line conditions. It is also an exact replot of the K8ZVF "lost power" nomograph, and indicates both the reflected and so-called "useful power" columns from the review article. The curve represents reflected power, ρ^2, versus SWR, if one reads downward from the top of the chart. 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. 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 above, when the reflection loss is canceled by the reflection gain of the conjugate match obtained by retuning the transmitter, a new curve, $\alpha = 0.0$ dB, which now represents this matched condition, follows the heavy straight line across the top of the graph. This indicates that 100% of the power is being made available, and is also taken by the load *regardless of the SWR value*. Suddenly the "lost" power is found!

As stated previously, power can be "lost" in a transmission line only through line attenuation - if the attenuation is zero, lost power is also zero, as shown along the $\alpha = 0.0$ dB curve, Fig. 9. The more attenuation, the more power is lost, as shown by the various loss curves marked $\alpha = 0.03$ dB and so forth. Since no allowance for the attenuation factor was made in either case in the material referenced in footnotes 27 and 28, we have, therefore, still another reason why the terms "lost power" in one case and "useful power" in the other are incorrect and misleading*. (Ref. 55)*

A later part of this series will deal in detail with attenuation effects, and will show how to perform some of the pertinent calculations based on the W2DU- derived equations associated with Fig. 9. However, the loss curves in Fig. 9 are in a form practical for visualizing the correct relationship between the losses actually encountered in different feed fines of various lengths, values of attenuation, and values of SWR. The curves indicate the total loss encountered on the line due to attenuation. The curves represent the condition in which the transmitter is conjugately matched at the input to 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. The loss value along each curve is seen to increase logarithmically as SWR increases on the line due to increasing mismatch of the line termination. Thus, the difference between the loss incurred with a 1.0-SWR termination 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. Here is further graphical evidence that when the line attenuation is low, the additional loss due to reflection is surprisingly small, even when the SWR is quite high.

Examine the SWR region between 1:1 and 2:1. Do you see enough difference in power level on any of these lines 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 pages 36 through 38 and page 40 of Part 1 (April *QST*, 1973) is now both pertinent and appropriate to emphasize how the use of these concepts can broaden our design flexibility. It is also suggested that the reader check the efficiency values of both the spacecraft feedline examples on page 37 and the 80- and 40-meter dipole examples on page 40 in the Fig. 9 graph. In the NAVSAT example, comparison of the 66% value of reflected-power with only 1.15 dB actual loss will be especially revealing.

Fig. 10 provides additional line-attenuation data which permit us to extend the use of the Fig. 9 curves to other frequencies and lines. Fig. 10 may also be supplemented with further data available in the ARRL *Handbook* and the *Antenna Book. (refs. 1 and 2).*

Radiation Resistance

In Part 2 of this series, statement 26 says, in effect, that no significant amount of power will be saved by using matching circuitry between the feed line and the antenna terminals of a mobile antenna system in the 80- through 10-meter bands. 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 rather widely misunderstood, so this is an appropriate time to clarify both of these statements, because these concepts also fit the category of "low SWR for the wrong reasons."

It is well known that the radiation resistance of the short 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 1.0 ohm in this band, as shown by Belrose *(ref. 60).* The capacitive-reactance component in the terminal impedance of this short antenna, which ranges from -*j*3000 to -*j*3500 ohms in typical 80-meter models (also shown by Belrose and confirmed from measurements made by the author), is canceled by the equal +*j* 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, 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 it is erroneously thought by many amateurs that the one-ohm radiation resistance alone comprises the entire antenna-terminal impedance, and also that a matching device is thereby required 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, loss resistance in the loading coil and any ground-loss resistance both add to the radiation resistance, causing the terminal resistance to be much higher than is commonly realized, though still lower than the impedance Z_C of normally used feed lines. Thus the actual mismatch value is much *lower* than is usually appreciated.

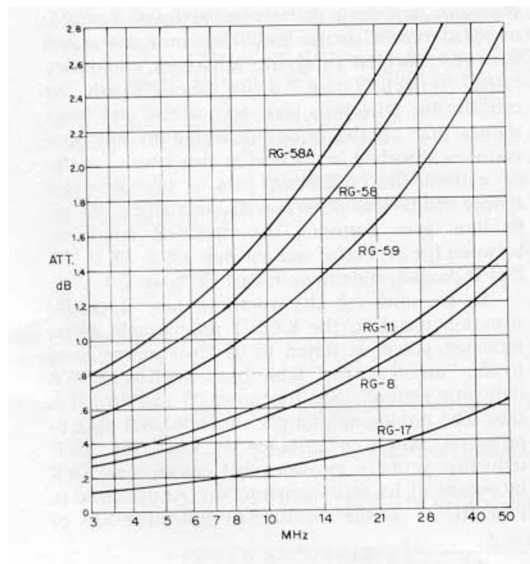


Fig. 10 - Attenuation in decibels per hundred feet for various coaxial cables.

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 exists, because, excepting Belrose, most writers neglect to mention it or consider it in their system analysis. For example, see Swafford, "Improved Coax Feed for Low Frequency Mobile Antennas," *QST* for Dec., 1951. The *Mobile Handbook*, p. 100, (Cowan Publications, 1st Edition) not only fails to recognize the existence of ground resistance, but mistakes what is actually the combined ground and radiation resistances as radiation resistance alone. (In subtracting 6 ohms of loading-coil and whip resistance from the 14 ohms of measured terminal resistance, the 8-ohm difference is simply taken to be radiation resistance, with no mention of any ground resistance.) Thus, the ground resistance, which cannot be ignored, is unknowingly and improperly being included as a portion of the radiation factor in the efficiency expression, instead of as a loss. 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 plus a 1-ohm radiation resistance subtracted from the 14 ohms measured leaves *7 ohms*, which requires an explanation of where the remaining 7 ohms of resistance came from. From further study of the Cowan text, it appears that an assumption of a greater amount of radiation from the coil than what this author considers 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 unrealistically high radiation resistance (8 ohms) and efficiency values (8/14 = 57%) that 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 considered in the Cowan *Mobile Handbook* as being radiated is actually dissipated as heat in the ground. Belrose shows a proper analysis *(ref. 60)*, supported by this author's measurements.

Now we will see why it's practically impossible even to obtain a mismatch of sufficient magnitude that will require 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*, the many remotely controlled luggage compartment tuning and matching arrangements notwithstanding.

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 resistance encountered with conventional low-band mobile antenna installations ranges from about 5 ohms for low, wet ground to around 12 ohms for high, dry ground; average ground yields around 7 ohms. The ground-loss resistance in the mobile setup is less than that found in an antenna of a full λ/4 in physical height with no radials, because the radius of the circle where the minimum currents return to the ground is shorter with the shorter antenna and therefore the currents travel a shorter distance through lossy earth (ref. Part 5, Fig. 8). (The current-flow pattern in the mobile system is also described by Belrose.) Thus, the total terminal resistance is nowhere near one ohm but lies in the range from an absolute minimum of around 14 ohms when using loading coils of high *Q* to as high as 44 or 45 ohms when higher ground loss and the use of a poorer quality coil occur simultaneously. Referenced to a 50-ohm feed line, these resistances result in SWRs (at the resonant frequency) from around 3.5:1 (with the lower loss resistances) down to around 1.1:1 with the higher losses.

Thus, as strange as it may seem, the *higher* the minimum SWR attainable *at resonance*, the greater will be the power radiated (for the same power delivered to the feed line by the transmitter). It will not seem so strange, however, when we consider that the low radiation resistance of around one ohm is the *only* portion of the total resistance that contributes to radiation -- and it 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 in heat, 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 is now being spent in heating the coil.[29] However, when using the higher-*Q*, lower-loss loading coil, even though its lower loss resistance results in a larger load mismatch and higher feed-line SWR, the increase in radiated power is still proportional to the decrease in total resistance. Any additional loss from the higher SWR is so small 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.

In Fig. 9 the loss curve $\alpha = 0.064$ dB represents the loss characteristics of a typical mobile feed line -- 20 feet of RG-8/U at 4.0 MHz. The curve shows a matched-line attenuation of 0.064 dB, plus an additional loss of 0.056 dB at SWR = 3.5, for a total loss of 0.12 dB. When this condition is accompanied by the 1:14 ratio of radiation resistance to total 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 0 to 14 ohms), the radiating efficiency is 7.14%, or 11.46 dB below the transmitter power delivered (excluding line loss). With the 1- to 45-ohm resistance ratio (with the higher loss coil and poor ground) the radiating efficiency is 2.22%, or 16.53 dB 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 1.1:1.

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 matching 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 Part 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 power from the transmitter into the line is to be performed at the input end or 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 his convenience and the accessibility to adjustment, and not on an arbitrary, low SWR dictated for the wrong reason by a decree of a 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.

Finally, here is a suggestion which may be helpful in tuning a mobile antenna. The use of the Grid Dip Oscillator (GDO) for determining the resonant frequency of the coil and radiator combination can introduce errors of sizeable magnitude, especially when disconnecting the line to make the measurement, or when measuring at the input terminals of the line. An SWR indicator connected directly at the input of the feed line can provide a more accurate indication of resonance, providing the instrument is reliable and accurately calibrated for the impedance of the fine *(ref. 59)*, because minimum SWR for a given terminating load of the type we're discussing occurs at the resonant frequency of the load, *regardless of the line length* (ref. Part 2 of this series, statement 24, and Part 5, the paragraph concerning "minimum-SWR resistance").

Low SWR for the Right Reasons

In concluding our remarks concerning low SWR for the wrong 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 images from appearing in the received picture images which would result from reflections on the feed line. Similarly, low SWR on the feed line is a necessity in fm stereo broadcasting to avoid cross contamination between the 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 feedline:

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

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

c) 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. (A detailed enlightenment on this crucial point has been promised for later presentation.)

And from reexamination of Figs. 9 and 10 it is evident that we *do not need* an SWR lower than 2:1 *on any line* to avoid any significant loss in efficiency, or with SWRs considerably higher than 2:1 when using feed lines having low attenuation. It would seem that there aren't too many reasons for needing a low SWR in amateur operations at hf (ref. Part 1, p. 40 and Part 2, statements 11 through 17). 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 50 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) A conjugate match placed anywhere on the line between a mismatched load and the transmitter compensates for the mismatch at the load, with the resulting effect that a conjugate match now exists *everywhere* on the line *(ref. 17, p. 243)*. In other words, if a mismatched load $Z_L = R + jX$ terminating the line is conjugately matched somewhere on the line, the reflection generated by the complementary mismatch at the conjugate matching point causes the impedance looking into the termination end of the line to change from Z_C to $Z = R - jX$ (see Part 4, pp. 22 and 23).

3) Now let's take advantage of the increased bandwidth immediately available to us simply in our knowledge that nothing magical or miraculous happens in "bringing the SWR down" to 1.0. When using coaxial line to center feed a dipole, the operation is usually for one amateur band only. But now we 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 we will define presently).[30] To minimize the increase in mismatch and SWR resulting from the frequency excursion to either of the band ends, the dipole should be cut to resonate near the center of the band. On the 75-80 meter band where the percentage of frequency excursion is the greatest, the mismatch at the band ends will be somewhat less severe with a 75-ohm feed line than with a 50-ohm line. Of the smaller sized cables, RG-59/U is preferred here over RG-58/U, because the combination of the lower maximum SWR and lower matched-line attenuation with the RG-59/U permits either a greater frequency excursion away from the self-resonant frequency of the dipole, or a longer line for the same loss factor. Of the larger size cable, either RG-8/U or RG-11/U gives nearly equal results, because the matched-line attenuation of the RG-11/U is a little greater than with RG-8/U, 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 factor than 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 is 1.0 dB. So, to find the SWR which reduces the radiated power by 1.0 dB, first use Fig. 10 to find the attenuation per hundred feet of the correct feed-line type at the desired operating frequency. Then apply the correction for the actual feed line. Now go to Fig. 9 and find the α-loss curve which 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. Read the SWR at this point. 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 will be presented later, but the following are fair rule-of-thumb values for dipole SWRs to be expected at the band ends when the dipole is cut for resonance at the band center:

Freq.	*Max. SWR Value*
3.5 to 4.0 MHz	5 or 6:1 (50-ohm line) 4 to 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

Applying these data to Fig. 9 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 1.0 dB of additional loss will seldom be encountered, and therefore, no "pile-up punch" will be sacrificed in obtaining the increased operating-bandwidth flexibility.

5) At an SWR of around 4:1, the additional loss due to SWR just equals the perfectly matched line attenuation, thus, in effect, multiplying the matched loss by a factor of two. As an example, this statement translates: The power lost in 175 feet of RG-8/U, or in 87 feet of RG-59/U, 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 0.5 dB.

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 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 at 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 which is matched at the load.

Since it has now been shown that any required matching can be performed at the input to the feed line, instead of at the load, no SWR bandwidth limit for amateur use (such as the commonly used, low, *arbitrary* 2:1 limit) 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 has been 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, these networks could be built into the transmitters, giving us 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, because the conjugate matching at the line input then basically involved a simple adjustment of the link position to achieve the proper *degree* of coupling, and the retuning of the plate-tank capacitor to cancel the reflected reactance. Using this technique, we often loaded our transmitters into lines with high SWR 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. 9 and 10, and 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!

Although reader response to this series of articles has been excellent, some have 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 any reader who still entertains any skepticism of these entire proceedings concerning SWR is reminded that the information presented herein is not simply a recitation of the ideas or opinions held by the writer, but has been taken directly from the professional scientific and engineering literature (see extensive bibliography), 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 text-book authors) the principles involved are completely understood and are based on true, proven scientific facts which are not subject to divergent opinions as found in politics or religion.

Apparently many have forgotten that this story was told for the amateur in *QST* no less than twice prior to this series, by two well-known experts in this subject area. They are Mr. George Grammer, W1DF, engineer and retired former Technical Editor of *QST*, and Dr. Yardley Beers, WØJF, formerly a professor of physics and Chief of the Radio Standards Physics Division, National Bureau of Standards, and presently Senior Scientist, Quantum Electronics Division of the National Bureau of Standards. Their illuminating contributions, listed as references 6, 16, and 22 of the bibliography appearing at the end of Part 1 of this series, should be reviewed, even if it means a trip to the library the trip will be very rewarding.

Part 7: My Transmatch really does tune my antenna

The birth of this series touched off a bombshell aimed at SWR misconceptions, emitting shock waves impacting on novices and old-timers alike. The myth believers among them were caught with their feed lines dangling at 1.05:1 -- and with the low-viswarmaniacs screaming at them to "Get that SWR *down*!" and Maxwell shouting "If ya do, it'll kill ya!" They didn't know whether to burn their noise bridges behind them or to lock their VFOs on the resonant frequency of the antenna. Probably the misconception causing the greatest shock of all concerns reflection loss -- the "lost reflected power" excuse offered by low-viswarmaniacs in defense of their insistence on low values of SWR on the feed line. This misconception was blasted to extinction by explaining the concept of conjugate matching in an unusual way -- from the viewpoint of reflections and wave action on the line. From this viewpoint (which involves both reflection loss and reflection *gain*), the conjugate match concept emerged as a basis for understanding the lost-reflected-power myth. However, while reflections from this series show that the blast dispelled the myth for many readers, some are apparently jousting for another round. So let's take aim for the seventh round by reiterating a basic principle of matching, after which we'll contemplate some enticing aspects of line-matching networks. In this part we will see in an intriguing way how transmitters are matched to mismatched feed lines by the pi-section network (Fig. 12), or by an external network (Fig. 11).

The principle that maximum efficiency is obtained when a feed line is perfectly matched to an antenna -- no reflections on the feed tine, and a 1:1 SWR -- is so well known it hardly needs repeating. Even so, it is recited in practically every good textbook on the subject. So it is important to appreciate that no statement in this series violates this principle, nor suggests any disagreement with it whatever. The misconceptions we are attempting to clarify concerning, lost reflected power stem simply from misuse of the perfect-match principle in its practical applications.

Ironically, textbooks may be a bit responsible for this misuse. While extolling the virtues of the perfect match, many authors fail to explain how much (or how little) one loses when the load is mismatched, if compensated with the *conjugate* match. Those authors generally present the case for the ideal load match in terms of single-frequency operation, and practically ignore the unique, multifrequency operations of the radio amateur. We have frequency *bands*, not single, specific frequencies - and we want to operate our VFOs anywhere within those bands. Since the antenna impedance changes as we change frequency, relaxation of any antenna/feed-line matching restriction is necessary if we are to enjoy such operating freedom. Although many engineering texts discuss loss versus load mismatch, few textbooks discuss multifrequency antenna-matching situations where conjugate matching usually exists. Consequently, an overly rigorous application of the perfect-load matching principle has unwittingly been thrust upon us by dozens of misleading statements appearing in various amateur journals, Add to this a prevalent misconception of pi-network loading principles. Result? The '"lost reflected power" syndrome and the mania for low SWR. Thus, providing guidance for the amateur concerning match quality and efficiency in his multi-frequency operation is the primary purpose of this series.

For example, the graph in Fig. 9, Part 6, displays transmission loss in dB versus SWR for various values of line attenuation. This graph shows that maximum, efficiency is indeed obtained with a perfect match. On the other hand, it also provides dramatic evidence that when using conjugately matched, low-loss feed line, the difference between having either a perfect load match or a moderately high SWR is practically insignificant in terms of power transferred to the antenna. In other words, through the use of conjugate matching, the antenna accepts the maximum available power from the transmitter, even when the feed line and antenna are mismatched, and with the antenna off resonance.[31] The conjugate match allows us to tolerate this load mismatch because the reflection loss caused by the mismatch is compensated by reflection gain provided by the conjugate match. Consequently, the transmitter is properly coupled to its desired load impedance and the reflected power is conserved. Underlying these intrinsic characteristics of the conjugate match is *system resonance*, which compensates for the effect of the off-resonant state of the antenna.

For the myth believer who is still unconvinced of the wondrous compensating powers of conjugate matching, let me quote Everitt on the Maximum Power-Transfer Theorem from classical network theory *(Ref. 17, page 49)*. Its relation to feed-line antenna matching is indicated in the parentheses: "The maximum power will be absorbed by one network (the antenna) from another (the feed line) joined to it at two terminals, when the impedance of the receiving (load) network (the antenna) is varied, if the impedances of the two networks at the junction are conjugates of each other." (Everitt then presents the proof.)

The expressions in Eqs. 12 and 13 (plotted in Fig. 9, Part 6) illustrate this theorem for the case where a feed line is the sending network. These expressions show that when the networks are conjugately matched, meaning the transmitter is tuned to the line, (1) there is no loss at the terminals joining the networks, (2) there is no loss in the sending network, the feed line, when that network attenuation is zero, and (3) if the sending network has attenuation other than zero, the transmission loss results only from the attenuation. For further discussion the reader is referred to Part 2, p. 21, Part 4, p. 25, mid Part 5, p. 163, of this series.

The wave action through which conjugate matching and reflection gain are achieved was illustrated in Part 4 using the stub form of matching to develop the wave action. The stub form was used for the illustration because it is easy to visualize. But since it's not a practical form to use if frequent changes in frequency are contemplated, let's see how conjugate matching is performed using devices which are easily adjusted to perform at any desired frequency. Such devices are the Ultimate Transmatch *(Ref. 41)*. a *T* network, *L*, pi, and other types of networks. We will see that these devices can perform the matching function at the *input* of a feed line, and that the feed line can be of *any* length, We will also see how, in some instances, line-input matching is performed by the transmitter tank circuit itself.

Match at the Line Input

Before going further the reader may ask, "'Why match at the *input*?" The answer is that without matching at the feed-line input we have very little operating flexibility. In the absence of a line-matching network we are restricted to operating in a narrow portion of a band (especially 80 meters) unless effective measures for broadbanding the antenna itself have been taken. We are restricted because, as we deviate from the antenna-resonant frequency, the resulting increase in feed-line/antenna impedance mismatch is transferred to the line input as an increased transmitter/feed-line impedance mismatch. As a result, the transmitter load impedance varies beyond acceptable limits; the transmitter fails to load property, and it can be damaged by overloading or by arc-overs caused by underloading. These phenomena (plus an unawareness of the remarkable performance capabilities of line matching) are largely responsible for the traditional low SWR mania. On the other hand, simple impedance matching at the feed-line input provides stupendous improvement in operating flexibility because the matching network compensates for the impedance changes at the feed-line input, and provides the correct load impedance for the transmitter at whatever frequency we select within an entire band. Correct load impedance is obtained by simply adjusting the network, conveniently located at the operating position.

So the next question is, "Why not broadband the antenna and avoid having to retune a matching device when changing frequency?" The answer is that we can, but only to a limited degree. This is because, for example, broadbanding techniques which would permit coupling the average amateur transmitter directly into the feed line over the entire 80-meter band (with no adjustments other than retuning the transmitter) are not practical in the average amateur situation. This includes the coaxial dipole (sometimes called a "double bazooka".) which, contrary to prevalent opinion, fails to deliver any significant bandwidth improvement over a simple dipole when it is fed with the usual 50-ohm feed line. While this statement may appear a bit incredible, a revealing analysis by the writer appears in *Ham Radio* for August, 1976 (Ref. 62).

An explanation of how either a final tank network or an external line-matching network performs conjugate matching from the viewpoint of reflections is really a continuation of Part 4. However, to dramatize both the availability and the advantages of matching at the feed-line input, we digressed in Parts 5 and 6 to highlight some of the wrong reasons for using low SWR. These reasons show that, in the mania for obtaining low SWR (in the feed line, we have often put misplaced emphasis on matching at the wrong end of the line. As stated previously, the principles of wave reflection found in stub matching are also applicable to other matching schemes, such as series quarter-wavelength transformers, *L*, *T*, pi networks and so on. Since some of the concepts we will be discussing were presented in detail in Part 4. you may wish to refer again to that section.

The Intermediate Role of the Quarter-Wavelength Transformer

How many remember the "Johnson *Q* Match"? And how many have used a quarter-wavelength section of 70-ohm line to match a 100-ohm resistive load to a 50-ohm feed line? Both are examples of series λ/4 transformers. In addition to requiring an electrical length of 90-degrees, impedance matching is accomplished in these transformers because of a specific relationship between their characteristic impedance Z_C, and the impedances being matched. To perform the matching, the transformer impedance Z_C, must have a ratio relative to its input impedance Z_i which is the inverse of the ratio relative to its output impedance Z_O, e.g.,

$$\frac{Z_i}{Z_c} = \frac{Z_c}{Z_o} \qquad \text{(Eq. 14)}$$

In other words, the impedance required of the transformer is the geometric mean value of the two impedances being matched, stated simply by the well-known expression

$$Z_c = \sqrt{Z_i \times Z_o} \qquad \text{(Eq. 15)}$$

Both the impedance and the length of the λ/4 transformer play important roles in clarifying the principles underlying all forms of line-matching networks. Since these roles are not generally were understood, let's examine them.

It was shown in Part 4 that reflections play a necessary role in impedance-matching operations. We saw that conjugate matching is obtained by canceling the reflections from a load mismatch by wave interference. The interference is set up by new, separate reflections generated by a separate mismatch introduced at a desired matching point. The mismatch introduced at the matching point is tailored to complement the load mismatch, so that the new reflection will have the same magnitude and opposite phase (at the matching point) as the reflection generated by the load mismatch. The reflection generated by this complementary mismatch is called either a complementary, or a canceling reflection. In stub matching, the complementary mismatch is introduced by the stub. While investigating complementary reflections generated by stubs, we observed the wave action through which the λ/4 transformer performs the matching (Part 4, Table 1 and Fig. 7D). While analyzing the effects of using various stub and transformer lengths in matching a resistive load to various feed lines having different values of impedance, we found that when the feed-line impedance is equal to Z_i (relative to transformer impedance Z, and load impedance Z_O, in Eqs. 14 and 15), transformer length became 90 degrees, and the stub length became zero. In this case the complementary reflection is generated by the mismatch appearing at the junction of the feedline and the transformer. This mismatch results from the abrupt change in impedance (Z_i to Z_C) encountered by the forward wave as it propagates out of the feed line (Z_i) into the transformer (Z_C). This mismatch is complementary to the resistive load mismatch in magnitude because the ratio between the feed line and transformer impedances (Z_i / Z_C,) is the same as between the transformer and load impedances (Z_C / Z_O). Thus the reflections generated by both load mismatch and transformer-input mismatch are equal in *magnitude* (as required to obtain perfect cancellation), because both mismatches are equal in magnitude.

Since the input and output mismatches are physically separated by 90 degrees, they are also complementary in relation to the *phase* of the reflections appearing at the matching point (also required for cancellation). The wave reflected by the input mismatch has zero distance to travel relative to the matching point. But the 90-degree separation results in a travel of 180 degrees for the wave reflected by the load mismatch -- 90 degrees from the input to the load, plus the 90-degree return trip. Thus the load-reflected wave arrives at the matching point with a 180-degree phase difference relative to the input-reflected wave. We now have two complementary reflected waves -- equal in magnitude and opposite in phase at the matching point. Consequently the two waves mutually cancel, resulting in total reflection of both waves into the transformer to propagate in the forward direction, as explained in Part 4. The voltage and current components of both re-reflected waves are in phase with their corresponding components of the source wave; hence a conjugate match appears at the transformer input, all of the power reflected from the load mismatch which reaches the transformer input is again on its way to the load, and no reflected wave appears on the feed line,

This $\lambda/4$-transformer matching action deserves serious study. Why? Because it provides an intermediate step in understanding how matching can be achieved at the input of a line transformer of *any random length*, and which may have *any impedance* for its terminating load, such as the complex impedance $Za = Ra + jXa$ of a mismatched, off-resonant antenna. And this line transformer is none other than the feed line so many strive to operate with no reflections by simply restricting its load to a matched, resonant antenna!

Input Line-Matching Networks

Now let's examine external line-matching networks (such as the Ultimate Transmatch) from the viewpoint of conjugate matching with reflections. Referring to Fig. 7D of Part 4, we replace the 90-degree transformer (T) with the combination of an adjustable network and a line of *random* length to connect the mismatched load (antenna) to the network. This arrangement is shown in Fig. 11. The line connecting the network to the antenna will now be called the "transformer line." The line connecting the transmitter to the network. is the "feed line,'" as before. The matching point is defined as the junction of the feed line and the input of the network.

In a manner which will be explained later, the transformer line (which can have any value Z'_c) transforms the complex antenna impedance $Z_a = R_a + jX_a$ to a second impedance $Z_2 = R_2 + jX_2$ at the transformer-line input. The network then transforms Z_2 to a third impedance $Z_3 = R_3 + jX_3$ at the matching point, When the network is correctly adjusted, Z_3 is a pure resistance equal to the feed-line impedance Z_C. ($Z_3 = R_3 + j0 = Z_C$) Thus the antenna impedance Z_a is conjugately matched to the feedline impedance Z_C, which is a proper load for the *transmitter (Ref. 17, p. 243).* Without the network the transmitter load would be impedance Z_2, well known for deviating far beyond the range of matching capability for most transmitters. However, by adding the network we obtain the conjugate match by simply adjusting the network to transform R_2 to equal the feed-line impedance Z_C at the matching point, and to cancel any transformer-line reactance X_2 to zero at the matching point.

Let's now examine the transformation of impedance Z_2, more closely. The variations of resistance R_2 and reactance X_2 of impedance Z_2 are dependent on three different factors: Antenna impedance Z_a and both the length and the impedance Z'_c, of the transformer line. A study of Part 4, pages 27-28, will show that for given antenna and transformer-line impedances, a length of line can be found that will make $R_2 = Z_C$, but will yield reactance X_2 which requires canceling to obtain a match (for example, with a stub as shown in Fig. 7A or 7B). Another length can be found that will make X_2 become zero, but now R_2, will not equal Z_C (still no match). There is no length that will yield $Z_2 = R_2 + j0 = Z_C$. This situation illustrates the typical, endless cat-and-mouse game we play in trying to load the transmitter by changing line lengths. A fanciful solution to this problem would be a line of variable length, plus a device for dumping the unwanted reactance. So how can we possibly solve the problem using a fixed, random-length line? Because, as we now discover, the line-matching network is a star performer! We know from elementary transmission-line theory that a transmission line is made up of an infinite number of tiny distributed series inductances and shunt capacitances, so it should not be surprising that we can adjust the electrical length of a line of a given physical length by adding-lumped inductances or capacitances.

Indeed, by a proper selection and adjustment of reactances arranged in an appropriate L, T, or pi configuration we can simulate a line having any desired electrical length without specifying any physical length whatever *(Refs. 8-13; .19, p. 115: 21; 22; 24; 30; 31; 41; 61; 63).*

When a matching network is adjusted to obtain a conjugate match between the antenna and feed-line impedances, it performs the following two feats. First, it creates the effect of stretching the electrical length of the transformer line to make it reach the matching point, so that the resistance R_2 (at the input of the physical transformer line) is transformed to $R_3 = Z_C$. And second, it introduces reactance $-X_3$ to cancel the stretched-transformer-line reactance X_3 appearing at the matching point, in the same manner as a stub would perform in stub matching.

The introduction of reactance $- X_3$ at the matching point constitutes a complementary mismatch which generates a canceling reflection having equal magnitude and opposite phase relative to the reflected wave arriving at the matching point from the mismatched antenna. The matching network has thus provided the proper *overall* transformer length to obtain both the desired transformation of the resistance component, and a canceling phase relation between the load-reflected wave and the canceling reflected wave. And in addition, the network has provided the complementary mismatch ($-X_3$.) which generated the canceling wave. Consequently, the network has also transformed a dream into reality by conjuring up the fanciful variable-length line and means for dumping the unwanted reactance,

An ideal setup for observing the action while performing the tuning adjustments of the network includes an rf ammeter in the transformer line to indicate line current (a meter in each conductor if using balanced, two-wire line), and a dual-meter reflectometer to indicate forward and reflected power simultaneously in the feed line. It is important that the reflectometer be adjusted initially to indicate zero reflected power with the feed line terminated in a pure resistance equal to the feedline impedance. The tuning adjustments are complete when we obtain maximum current in the transformer line simultaneously with maximum forward and zero reflected power in the feed fine. Of course, because of standing waves on the transformer line, we will obtain different values of line current depending on where along the line the ammeter is inserted. However, our only interest is in seeing *changes* in relative current to indicate when maximum network output occurs during the tuning adjustments; thus neither absolute line current, nor where the meter is inserted, is Important.

The simultaneous indication of maximum current in the transformer line and zero reflected power in the feed line denotes four significant factors for comparing the wave actions involved in conjugate matching with a line-matching network and matching with the $\lambda/4$ transformer described earlier. First, proper network adjustment establishes the complementary mismatch between the feed-fine termination and the network input which produces the canceling reflection at the matching point. Second, at the matching point the canceling reflection is equal in magnitude and opposite in phase relative to the load-mismatch reflection and, the canceling reflection and the load-mismatch reflection nullify each other at the matching point. Consequently, a purely resistive impedance equal to the feed-line impedance Z_C appears at the network-input terminals, while reflections and a standing wave remain on the transformer line. And fourth, observing the transformer-line current rising to maximum while the reflected power in the feed line drops to zero provides visible evidence that the power reflected from the load mismatch is indeed rereflected by the complementary mismatch at the matching point.

Incidentally, it is good practice to have an ammeter permanently connected in the transformer line. Here's why. If your matching network effectively comprises more than one *L*-section, you can generally obtain a match (zero reflected power in the feed line) with several different combinations of network *L* and *C* tuning. However, minimum network loss (which corresponds to highest maximum transformer-line current) usually occurs while using the maximum *C* and minimum *L* at which a match can be obtained. Monitoring transformer-line current during tune-up lets you select the *L-C* combination yielding the highest output line current; it also ensures against inadvertently obtaining an *L-C* combination which gives a false indication of a match, but yields little output and instead heats the network inductance. To ensure quick resettability and to minimize on-the-air tune-up tune, *L* and *C* settings for the best combination Should be logged whenever the network is tuned to a new frequency. The transmitter should be tuned initially into a dummy load, then the network tuned into the antenna using the lowest power at which the reflectometer provides a satisfactory indication.

Tank Circuit Line Matching

Let's now examine the case where the final amplifier tank circuit performs the line-matching function. To differentiate from the external network we will call this network the "tank network." Comparing Figs. 11 and 12, we see that, in general, the transformations of impedance from Z_a to Z_3 are identical whether using the tank network or the external network. The principal difference is the range of the transformations performed by the two networks. With the external network, impedance Z_2 is transformed to a value of Z_3, matching the feed-line impedance Z_C. When the tank network is used alone, impedance Z_2 is transformed to match directly the load impedance Z_L of the generator. Now a requirement for a generator (tube or transistor) to deliver its maximum available power into a load (the loaded tank circuit) is for it to see an impedance which we call, the *optimum load impedance*, Z_L, (not the same as *internal* impedance, Zg.). In practice, impedance Z_L, is usually resistive, so that $Z_L = R_L + j0$. Thus the generator is property loaded when it sees impedance $Z_3 = R_L + j0$ (which is $R_3 + j0$) looking into the tank network. The generator is underloaded with R_3 larger than R_L, overloaded with R_3 less than R_L and it will have higher than normal dissipation if Z_3 contains any reactance X_3.

When using the tank network alone to perform the matching, the impedance value Z_3 is determined by two factors: The impedance value of Z_2 loading the network, and the impedance transformation ratio of the network. The transformation ratio is somewhat variable (using the tuning and loading capacitors, C1 and C2), and providing a *range* of impedance-matching capability. This matching range allows impedance Z_2 to be any value which the network can transform to the impedance $Z_3 = R_L + j0$ by adjusting the tuning and loading controls.

Now we can summarize the principal operating conditions involving transformation of antenna impedances to the optimum load impedance for the generator.

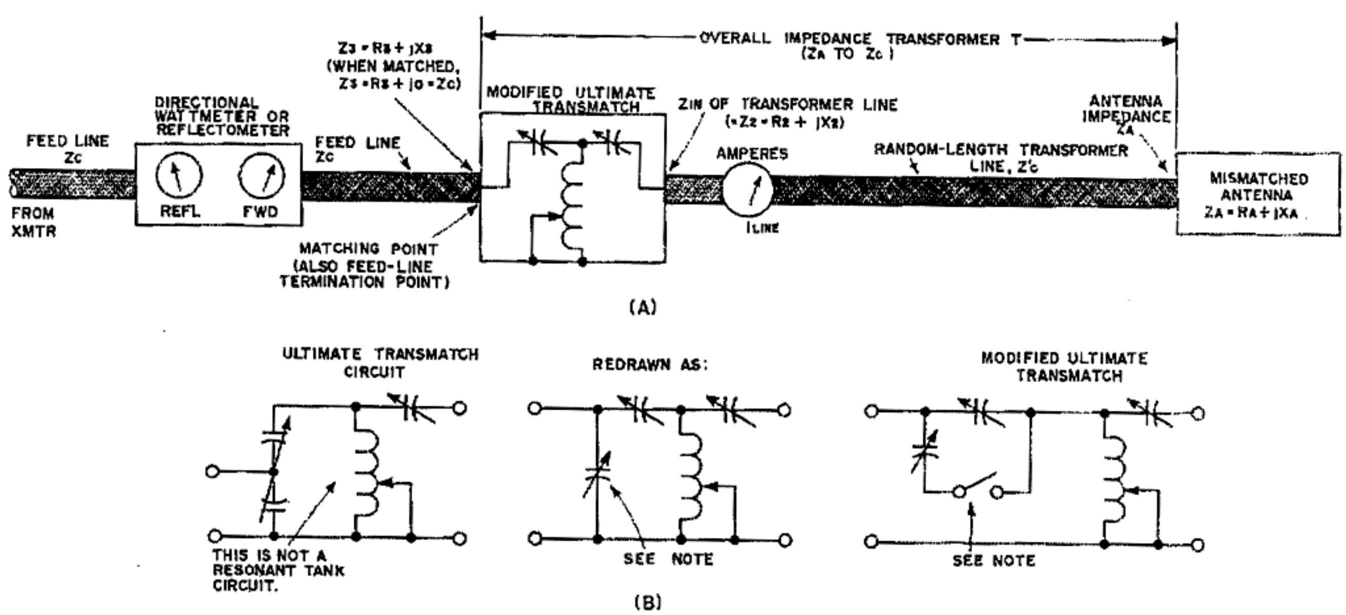


Fig. 11 - Conjugate matching with a Transmatch or other matching network external to the transmitter. The impedance of the feed line, Z_C, is usually 50 or 75 ohms in most amateur applications, but the impedance of the transformer line, Z'_C, may be any value, and may be of open-wire or coaxial type. The matching network as A is a modified Ultimate Transmatch (Ref. 41), as shown at B. NOTE: In the redrawn circuit at B (center) the shunt input capacitor is not needed to obtain a match, and the capacitor range may be increased by reconnecting as shown at the right.

Case 1

Antenna impedance $Z_a = R_a + j0$ (resonant) is matched to the transformer line, and the tank network is used alone -- no external matcher: Here antenna impedance Z_a, equals the impedance of the transformer-line, Z_C, the SWR on the line is 1:1, and impedance Z_2 is equal to the line impedance. If the transformer-line impedance Z_C is the commonly used 50 ohms, the tank network yields the proper generator-loading $R_L + j0$ with the same tank adjustment settings as obtained when tuning up with a 50-ohm dummy load

Case 2

Antenna is operated somewhat off resonance. Its impedance $Z_a = R_a + jX_a$ yields a mismatch to the transformer line so as to transform Z_a to an impedance Z_2 that is within the matching range of the tank network (for transformation to $Z_3 = R_L + j0$). Again the tank network is used solo.

Case 3

Antenna operated off resonance beyond matching range of tank network. Here the tank network is used in conjunction with an external network. An impedance $Z_3 = R_L + j0$ cannot be obtained with the tank network alone, and the generator would be either underloaded, overloaded, or reactively loaded. We remedy this situation by inserting an external line-matching network (as in Fig. 11), which transforms impedance Z_2 to an impedance Z_3 that is within the range the tank network can transform to $R_L + j0$.

Case 1 needs no comment, so let's refer to Fig. 12 and examine the action in the tank network as it performs the matching function under the conditions of Case 2. As stated earlier, when our generator is the commonly used Class AB, B or C amplifier, its optimum load impedance Z_L and its internal impedance Z_g are not the same. Although the explanation is beyond the scope of this article, it can be shown that Z_g effects only a light, high-impedance loading on the input of the tank network. Thus the input of the network (the matching point) is practically open circuited to energy traveling toward the generator. Consequently, waves reflected from a mismatched antenna (which cause impedance Z_2 to deviate from Z_C) enter the network at its output terminals and become almost totally reflected on arrival at the open-circuited input. When the network is tuned to resonance, voltage and current components of the reflected waves are reflected in phase with the corresponding source-wave components emanating from the generator. Thus the generator sees a resistive load $Z_3 = R_3 + j0$, and the reflected power adds to the generator power. This is why a directional wattmeter (or reflectometer) indicates a higher forward power than the generator is actually delivering when reflections are present on the transformer line. (See Part 4.)

Adjustment of capacitance C2 (the loading-control capacitor) to the value which provides optimum generator loading adjusts the network to transform R_2 to $R_3 = R_L$. When R_2 changes, following a change in antenna impedance Z_a, C2 can then be varied to modify the network transformation ratio to transform the new value of R_2 to $R_3 = R_L$. The tuning capacitor, C1, is then readjusted to return the network to resonance an we again have optimum generator loading. The range of R_2 values which can be transformed to $R_3 = R_L$ by varying C2 is determined by the design parameters of the network *(Refs. 4, 63, 64).*

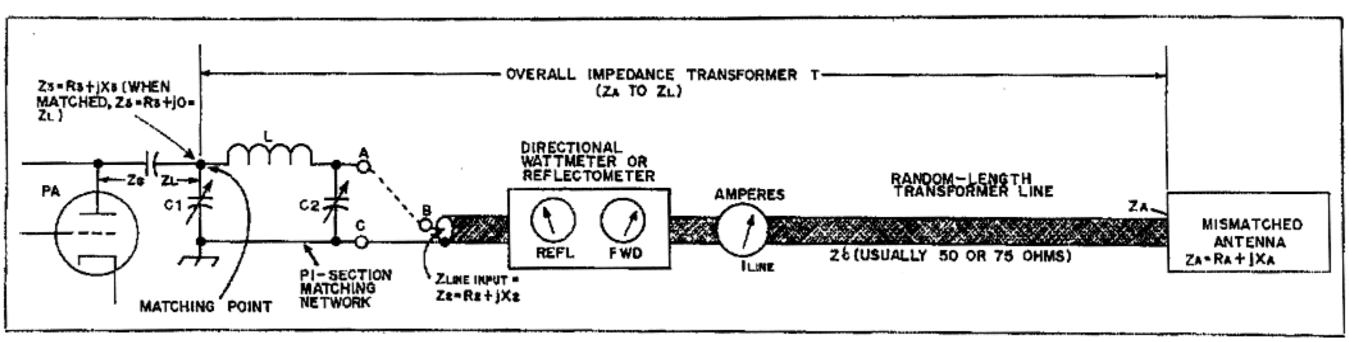


Fig. 12 - Conjugate matching with the transmitter final amplifier tank circuit. Z_L is defined as the optimum load impedance of the generator - the impedance into which the generator delivers maximum available power. This impedance must not be confused with the internal impedance of the generator, Zg; they are not the same.

However, when the line input impedance Z_2 loading the network contains reactance X_2, this reactance shifts the available range of capacitance provided by the loading control capacitor, C2. This shift occurs because reactance X_2 appears in parallel with the loading capacitor reactance. Consequently, for a proper setting of the loading capacitor where $X_2 = 0$, a capacitive X_2 increases the effective capacitance C2. (decreasing loading), while an inductive X_2 decreases C2 (and increases loading). Thus, to obtain a proper loading when X_2 is present, the setting of the loading-control capacitor must be shifted from the $X_2 = 0$ position to compensate for the additional line-reactance X_L. Providing R_2 is within the matching range, proper loading will be attainable as long as X_2 doesn't exceed the value which the loading-control capacitor range can accommodate to maintain the value of C2 required to transform R_2 to $R_3 = R_L$.

If reactance X_2 is too large for the loading capacitor to compensate, we have conditions as described in Case 3. However, there are simple remedies for this condition if R_2 is within matching range of the tank network, but X_2 is not. Here we can simply add a compensating reactance in series with the rf output (between points A and B in Fig. 12) which cancels the line reactance X_2. On the other hand, if the resistance component R_{2P} of the equivalent parallel-circuit of impedance Z_2 is within the matching range, the compensating reactance should be added in *shunt* across the rf output (between points A and C). Whether the compensating reactance should be capacitive or inductive can be determined experimentally by trying first one kind or the other to see whether loading is improved or worsened. The correct kind and value of compensating reactance has been found when proper loading can be obtained with a convenient setting of the loading-control capacitor. An excellent discussion of this subject by Grammer appears in the literature *(Ref. 4, Part 3).* When using this matching technique with the tank network alone, an adjustment of transformer line length can be of great assistance in obtaining values of impedance Z_2 which are most favorable to the matching range of the network. This subject of impedance transformation versus line length will be discussed in a later installment in this series of articles.

When operating under the conditions of Case 2, tuning up into a dummy load requires special care; it *can* be troublesome since the actual load often differs widely from the dummy-load value. If the tank network is first tuned into the dummy load and then switched to the transformer-line input impedance Z_2 (Fig. 12), the tank network must be *retuned* to the new impedance Z_2! Failure to retune to the actual operating impedance Z_2 after tuning up into the dummy load results in an improper load impedance Z_3 for the generator -- it is no longer $Z_3 = R_L + j0$ because the tank-network loading has changed. Without retuning, the generator is then either underloaded, overloaded or reactively loaded. In addition to the possibility of damaging the amplifier if operated in this mistuned condition, you also lose power! Because of this mistuning, the generator *delivers less power* by the amount of the reflection loss resulting from the line mismatch, as shown in Part 6, Fig. 9, on the curve labeled "Reflection Loss Without Conjugate Match" at the appropriate SWR ordinate. For example, if the line SWR were 3:1, the generator output drops by 25 percent while it sees the improper load. On the other hand, retuning the network to the actual operating impedance Z_2 establishes a conjugate match that restores the proper generator load impedance $R_L + j0$, and the generator again delivers its maximum available power. (See Part 6, pp. 14-15.)

But you ask, “How do we determine when proper loading is obtained, or that impedance Z_2 is within the matching range of the network?" The answer is simply by completing a normal tuning and loading operation in which you can obtain the same plate-current, plate-current dip (and screen-current) reading as those obtained with the dummy load. However, tuning- and loading-control settings, and the relative power (output voltage) will generally differ from those obtained with the dummy load, depending on how much Z_2 differs from 50 ohms. If normal plate current (and dip) cannot be obtained with any setting of the loading capacitor, Z_2 is outside the matching range and we have conditions as defined in Case 3. If proper loading cannot be obtained using the simple series or shunt reactance compensating technique described earlier, then a more complex network is required. However, I would like to emphasize that *any* value of $Z_2 = R_2 + jX_2$ can be transformed to a suitable value for loading the tank network by selecting a proper network configuration *(Refs. 19. p. 115; 30; 61)*

The range of impedances Z_2 which the tank network will transform to the value equal to the generator impedance $Z_L = R_L + j0$ raises an interesting point concerning the tuning procedure for the external network. The usual practice is to tune the tank network with the dummy load, and then switch in the external network and antenna and tune for zero reflected power in the feed line (not the transformer line). This procedure should be followed if there is a filter in the feed line. However, in the absence of a filter, it is necessary to adjust the external network only for an input impedance Z_3 which will bring it within the matching range of the tank network. This can be a time-saving feature when changing frequencies during contest operation! If both tank and external networks are adjusted for optimum match at midrange of the intended frequency excursions, only the tank network needs retuning with changes in frequency, providing the frequency excursions are within the range in which the external network yields a load impedance that the tank network can transform to $R_L + j0$.

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49) King and Blake, "The Self-Impedance of the Symmetrical Antenna," Proc. of the IRE, July, 1942, p. 335.

50) Sevick, "The Ground-Image Vertical Antenna," QST, July, 1971, p. 16.

51) Sevick, "The W2FMI Ground-Mounted Short Vertical," QST, March, 1973, p. 13.

52) Kraus, Antennas, McGraw-Hill, New York.

53) Jordan, Electromagnetic Waves and Radiating Systems, p. 415, Prentice-Hall, New York. 54) Fayman, "A Simple Computing SWR Meter," QST, July, 1973, p. 23.

55) Anderson, "SWR's Significance," CQ, October, 1970, p. 8.

56) Smith and Johnson, "Performance of Short Antennas," Proc. of IRE, October, 1947, p. 1026.

57) Laport, Radio Antenna Engineering, McGraw-Hill, New York.

58) Jasik, Antenna Engineering Handbook, McGraw-Hill, New York.

59) Schreuer, "Notes On Directional SWR Indicators," Ham Radio, December, 1969, p. 65.

60) Belrose, "Short Antennas for Mobile Operation," QST, September, 1953, p. 30. (Information also contained in The Mobile Manual for Radio Amateurs, ARRL, now out of print.)

61) Leo, "How To Design L-Networks, " Ham Radio, February, 1974, p. 26.

62) Maxwell, "A Revealing Analysis of the Coaxial Dipole Antenna," Ham Radio, August, 1976, p. 46.

63) Gray and Graham, Radio Transmitters. McGraw-Hill Book Company, 1961, pp. 115-133

64) Schottland, "Pi Networks as Coupled Circuits," Electronics, August, 1944.

Footnotes

[2] Forward, or incident power equals line-input power plus reflected power. Line-input power and incident power are equal only when the line is perfectly matched at the load, for zero reflected power.

[3] Transmitter manufacturers could easily have helped put an end to this myth by including two or three short paragraphs in their instruction manuals explaining the maximum and minimum coupling limits of the pi network, and how to extend them with an external capacitor, instead of the terse, uninformative "WARNING - DO NOT EXCEED 2:1 VSWR," and so forth.

[4] Many baluns have high leakage reactance, and thus cannot provide a true one-for-one impedance transfer. This reactance inserted between the antenna and its feed line can either improve or worsen the match, depending on the magnitudes and signs of the leakage and antenna-terminal reactances.

[5] To satisfy the conjugate conditions when the generator is a Class B or C amplifier, the impedance replacing the generator must be made equal to its optimum load impedance, the load into which the generator delivers its maximum power to the loaded tank. For a Class C amplifier this is roughly twice its internal impedance. The reason for this difference may be more fully appreciated when we consider that the classical network generator has a maximum efficiency limit of 50 percent because it delivers its maximum available power when its load impedance equals its internal resistance. But in the Class C amplifier the effective internal ac impedance is about half of its optimum load impedance because the current pulses which excite the loaded tank are of high peak value and short duration, while the instantaneous anode voltage during current flow is very low. This results in proportionately less power lost in the generator and more delivered to the load, enabling it to operate at an efficiency as high as 75 percent or more.

[6] With a conjugate match all reactance has been canceled, leaving the reflected voltage polarity either pure in- phase aiding, or out-of-phase bucking the source voltage. If bucking, the explanation of the text is sufficient, because a source of smaller voltage can never cause a reverse power flow through a larger voltage source. Whether bucking or aiding, if the loading adjustment leaves the generator impedance lower than the line-input impedance, the reflected wave is still totally rereflected, but the generator is undercoupled and not delivering all available power. But any loading adjustment or matching error which leaves the fine-input impedance lower than the generator impedance results in overcoupling, or overloading, and this, not reflected power, directly causes excessive generator dissipation and lowered efficiency.

[7] Detailed instructions (plus examples) on calculating the additional loss because of SWR for given line attenuation and SWR values will be presented later. The data may also be taken from part 1, Fig. 1, which comes from the Transmission Lines chapters of The ARRL Antenna Book and the ARRL Handbook, or from bibliography ref. 33 p. 573.

[8] Smith charts may be obtained at most university book stores. They may be ordered (50 for $2.50 postpaid when remittance is enclosed) from Phillip H. Smith, Analog Instruments Co., P.O. Box 808, New Providence, NJ 07974. For 8 1/2 x clinch paper charts with normalized coordinates, request Form 82-BSPR. A brochure of Smith charts and accessories will be sent upon request. NOTE: Smith charts with 50-ohm coordinates (Form 5301- 7569) are available at the same price from General Radio Co., West Concord, MA 01781.

[9] Footnote 9 was the Editor's note, included within the text, for continuity

[10] Reference ASA Y10.9, 1953 (American Standards Association). Prior to the adoption of this standard, either Γ or *k* was frequently used to indicate the reflection coefficient, while ρ was

often used for standing-wave ratio. In studying the literature, the reader should exercise care to avoid confusing symbols used before and after adoption of the standard.

[11] Remember that motor-generator action results from mutual motion between a field and a conductor. Here the field is changing even though the conductor is not moving, as in a transformer.

[12] The characteristic impedance, Z_C of lossless line has zero reactance, and low-loss line has so little reactance that it is neglected.

[13] Incident voltage and current should not be confused with line voltage and current, because they are not the same except when the line is perfectly matched and no reflections exist.

[14] Smith charts my be obtained at most university book stores. They may be ordered (50 for $2.50, postpaid when remittance is enclosed) from Phillip H. Smith, Analog Instruments Co., P.0. Box 808, New Providence, NJ 07974. For 8-1/2 X 11-inch paper charts with nomalized coordinates, request Form 82-BSPR. A brochure of Smith charts and accessories will be sent upon request. Smith charts with 50-ohm coordinates (Form 3801-7569) are available at the same price from General Radio Co., West Concord, MA 01781. Readers who are unfamiliar with the Smith chart are encouraged to consult the bibliography of Part 1 -for references; they will find the chart a valuable tool (See ref. 19, 25. 26, 27, 28, 29)

[15] The relative phase angle between the incident and reflected voltage waves is the angle of the reflection coefficient, referenced on the incident. See the "Angle of Reflection Coefficient" scale around the perimeter of the impedance plotting portion of the Smith chart.

[16] On the chart the dashed lines show the vectors after 180 degrees of travel on the line.

[17] Woods, "Power in Reflected Waves," Ham Radio, October, 1971, and Woods' correspondence on same, Ham Radio, December, 1972, p, 76.

[18] DeLaMatyr, "Reflections on Reflected Power,"' Technical Correspondence, QST for November, 1972, P. 46.

[19] Line Voltage, *E*, and line current, *I*, are the resultants of the forward and reflected *E* (+ and -) and I (+ and -) components (Fig. 4). Line *E* and *I* respectively, are measurable with a simple rf voltmeter across the line and an ammeter in series with the Line. Both quantities may be seen to vary along the line with the reflections (Fig. 6), and are generally reactive except for the two specific cases where they are in phase.

[20] See point 3, Part 1 of this series (p. 38 of QST for April, 1973).

[21] Alternatively it may be a short-circuit, depending on line and load conditions, which will be explained later.

[22] We are rather accustomed to thinking of 50 ohms as a standard system impedance because of the* preponderance of rf components and coaxial lines for that impedance. However, many calculations are greatly simplified by using normalized impedances, in which all impedance values have been divided by the system impedance. The system impedance is usually taken as the characteristic impedance of the transmission line, Z_C. Normalizing an impedance by dividing it by Z_C amounts to a change of scale such that the unit of impedance is Z_C ohms, rather than 1 ohm. The Smith chart Vector Graph, Fig. 4, uses the normalized system to take advantage of the simplification in the calculations. To obtain normalized values occurring in a system of any impedance, simply divide all impedances by the system impedance. For example, in a 50-ohm system 150 ohms becomes 3.0 ohms. To convert back to the 50-ohm system, simply reverse the process and multiply the normalized values by 50. For example, the normalized impedance $Z = 0.6 - j0.8$, found at $L = 45°$, becomes $Z = 30 - j40$. In the drawing for Fig. 6A, p. 24, the E^{-} resultant vector should be shown with a value of $\bar{\rho} = 0.5 \angle 0°$ rather than 60°. In the caption for Fig. 7, p. 27, the last few words should read". . . stub length goes to zero.

[24] See footnote 23.

[25] Glanzer, "More Words on Antennas," CQ July, 1957, p. 40; see p. 48 for material being referenced.

[26] See footnote 25.

[27] Houghton, "Convert SWR Into Watts," CQ, June, 1970, p. 36. Here it is stated 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, VE3AAZ (ref. 55), also in CQ, October, 1970, p. 8.

[28] Scherer, "CQ Reviews: The Knight Kit P-2 SWR Power Meter," CQ, March, 1963, p. 31. Reference is being made to data on p. 90 of that issue, where 100% minus the reflected power is erroneously shown as "useful" power. The succeeding paragraph on p. 90 also states incorrectly that the SWR indicator must be placed at a multiple of a half wavelength from the load to indicate the true SWR. The example demonstrates that the SWR indicator was either not properly adjusted to the impedance of the line, or that it was unreliable. See Part 2 of this series, statements 21 and 23, and refs. 38, pp. 25 - 26; 59.

[29] Many amateurs unknowingly select loading coils of low Q because "they produce a lower SWR than any other types of coils." They do, indeed, produce a lower SWR, because of their higher loss resistance, explained in the text.

[30] An exception is that satisfactory operation may be obtained on 15 meters with a 40-meter dipole.

[31] Remember, impedance mismatch does not cause current to flow on the outside of the coaxial feed line. (See Part 5, QST, April, 1974, p. 27.)

Feedback – Part 4

QST November 1973, pg. 46

Was it gremlins which crept into Part IV of Maxwell's series of articles, "Another Look at Reflections" (*QST* for October, 1973)? A portion of the information in footnote 22 (p.23) was omitted, garbling the meaning of the content.

[**Note** These corrections have been incorporated into the text in this electronic version – TIS]