A Commercial Triplexer Design

By George Cutsogeorge, W2VJN

The June 2010 issue of *QST* had an article on a home-made triplexer that allows two or three transmitters to feed a single triband antenna.[1] (This article is provided in the supplemental files section of the *ARRL Handbook* CD-ROM.) This design works well and was built by many people for Field Day and the WRTC championship activities. The author of the article, Gary Gordon, W6KV, was contacted and he had no intention of going commercial with his design. It seemed like a natural idea for a commercial product, so the folks at Inrad (International Radio — **www.inrad.net**) set out to make one.

As designed, the triplexer consists of three series-tuned circuits connected together at the output. Each is resonant on a different band; 20, 15 and 10 meters. The property of a series-tuned circuit that allows the triplexer to work is that the impedance is minimum at resonance and increases for frequencies above and below resonance. So, when connected together at the output or antenna input, the three circuits act nearly independently. The actual component values were selected as a compromise between insertion loss and adjacent band rejection. The resulting rejection is between 13 and 20 dB. *External band-pass filters must be used — the isolation provided by the triplexer by itself is not sufficient to prevent receiver damage!*

Looking at it as a commercial product showed two factors that were problematic. First, the *QST* design used variable capacitors, which are expensive. Also, the capacitor was not grounded, so it needed to be insulated from the enclosure and from the person adjusting it.

Second, analysis of the circuit indicated that the voltage on the capacitor shaft would be 140 volts peak and between the capacitor and inductor it would be about 2000 volts peak for 200 watts input on each single band. This level of voltage could be dangerous to the operator and an ungrounded shaft adjustment is never a good idea.

Both of these concerns were addressed by using three band-pass filters connected together at their outputs to replace the tuned circuits. Various circuit arrangements were tried in *ELSIE*, the filter design program from Tonne Software (**www.tonnesoft.com**) that is provided on the CD-ROM included with this book. The mesh circuit has two positive characteristics for this job. It has series elements so the input and output impedances rise out of the pass band. Also, the element values are perfect for the task at hand. Two tuned circuits provide 15 to 40 dB adjacent band rejection and minimum insertion loss.

Once the schematic was firmed up, circuit analysis was performed with *LTspice* (**www.linear.com/designtools/software**) to determine the component voltage and current requirements for 200 watts per band. The coil voltage requirements allowed selection of powdered-iron toroids that would not overheat and the currents were used to select the wire size. Capacitors were a bit harder to find until multi-layer ceramic capacitors (MLCC) were tried. These capacitors are presently available with up to 500-volt ratings and also have maximum current vs frequency plots. In some cases, series-parallel capacitors were used to meet desired voltage and current requirements. A safety factor of 1.5 to 2 allowed for some variations in antenna SWR.

Capacitors and toroid cores were obtained to verify the selections made from manufacturer's data. Various coils were wound and tested on a Q meter. Type 17 cores had the best Q over the range of 14 to 30 MHz. The -130 size was initially selected. Q was over 200 on all three bands.

Several types of capacitor were tested using the Q meter. The leaded KG-type capacitors tested slightly better than DUR micas, but MLCC surface mount capacitors turned out to be substantially better than both DUR mica and KGE types. An added bonus is that there is actual data on their current performance over the frequency range of interest. This adds a degree of design confidence over other types.

Next, breadboard circuits were built up in a form very close to an actual PC board design. Extensive testing was done with a dummy load and triband Yagi antennas to determine if the design worked in real-world conditions. Some changes in SWR occurred, but the maximum value did not greatly exceed what the antennas showed by themselves.

Power testing followed. Most testing was done at 50% duty cycle for time periods that allowed the temperature to stabilize at a maximum level. The capacitors stayed well within maximum temperature specifications but the coils heated a bit more than desired. Larger cores were obtained and the power testing was repeated. This time the temperature rise was less.

The PC board was designed and a prototype was built. SWR and frequency response looked good, so we went into power testing again. With a full ground plane the temperature rise of the capacitors was much lower. Temperature rise in the coils was 25 °C over ambient. Otherwise performance was similar to the breadboard testing. In production there was some adjusting required to account for component tolerances. This was accomplished by spreading or compressing the toroid windings a small amount.

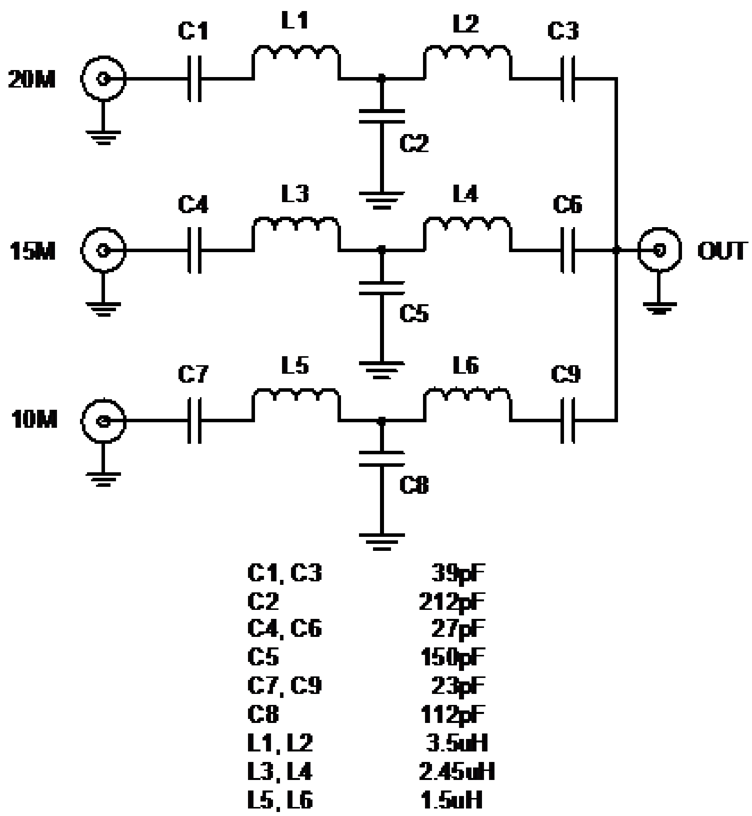


Figure 1 — The commercial triplexer schematic.

The basic schematic of the triplexer is shown in **Figure 1**. Each band consists of a 2-element mesh-type band-pass filter designed in *ELSIE*. The outputs are connected together and there is little interaction between them. Since MLCC capacitors are limited to 100 pF and 500 volts peak, each capacitor is made up of a series-parallel combination.

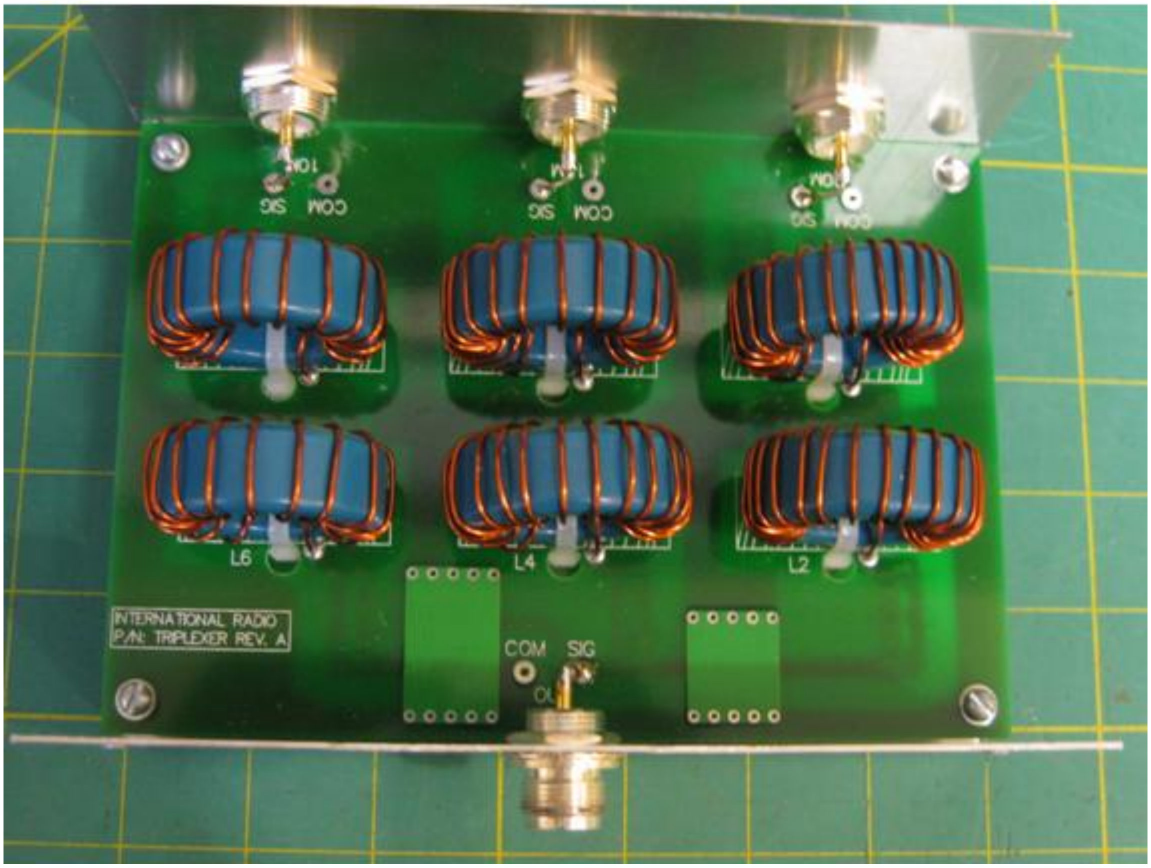


Figure 2 — The inside of the commercial production triplexer.

Figure 2 shows the inside view of a production triplexer. The values in the production unit vary slightly from those in Figure 1 to account for PC board strays. The basic schematic could be built with DUR mica capacitors and used for lower power operation.

References

1. G. Gordon, K6KV, “HF Yagi Triplexer Especially for ARRL Field Day,” *QST*, June 2010, p. 37.

Broadband Transformers

(excerpted from Chapter 14 of the ARRL Handbook, 2009 and previous editions)

Conventional Transformers

Fig 14.54A shows a push-pull amplifier that we will use to point out the main properties and the problems of conventional transformers. The medium of signal transfer from primary to secondary is magnetic flux in the core. If the core material is ferromagnetic then this is basically a nonlinear process that becomes increasingly nonlinear if the flux becomes too large or if there is a dc current through the winding that biases the core into a nonlinear region. Nonlinearity causes harmonics and IMD.

Push-pull operation eliminates the dc biasing effect if the stage is symmetrical. The magnetic circuit can be made more linear by adding more turns to the windings. This reduces the ac volts per turn, increases the reactance of the windings and therefore reduces the flux. For a given physical size, however, the wire resistance, distributed capacitance and leakage reactance all tend to increase as turns are added. This reduces efficiency and bandwidth. Higher permeability core materials and special winding techniques can improve things up to a point, but eventually linearity becomes more difficult to maintain.

Fig 14.54B is an approximate equivalent circuit of a typical transformer. It shows the leakage reactance and winding capacitance that affect the high-frequency response and the coil inductance that affects the low-frequency response. Fig 14.54C shows how these elements determine the frequency response, including a resonant peak at some high frequency.

The transformers in a system are correctly designed and properly coordinated when the total distortion caused by them is at least 10 dB less than the total distortion due to all other nonlinearities in the system. Do not over-design them in relation to the rest of the equipment. During the design process, distortion measurements are made on the transformers to verify this.

The main advantage of the conventional transformer, aside from its ability to transform between widely different impedances over a fairly wide frequency band, is the very high resistance between the windings. This isolation is important in many applications and it also eliminates coupling capacitors, which can sometimes be large and expensive.

In radio-circuit design, conventional transformers with magnetic cores are often used in high-impedance RF/IF amplifiers, in high-power solid-state amplifiers and in tuning networks such as antenna couplers. They are seldom used any more in audio circuits. Hybrid transformers, such as those in Fig 14.8, are often "conventional."

Fig 14.54D considers a typical application of a conventional transformer in a linear Class-A RF power amplifier. The load is 50 Ω and the maximum allowable transistor collector voltage and current excursions for linear operation are shown. The value of DC current and the sinewave limits are determined by studying the collector voltage-current curves (or constant-current curves) in the data manual to find the most linear region. To deliver this power to the 50 Ω load, the turns ratio is calculated from the equations. In an RF amplifier a ferrite or powdered-iron core would be used. The efficiency in this example is 37%.

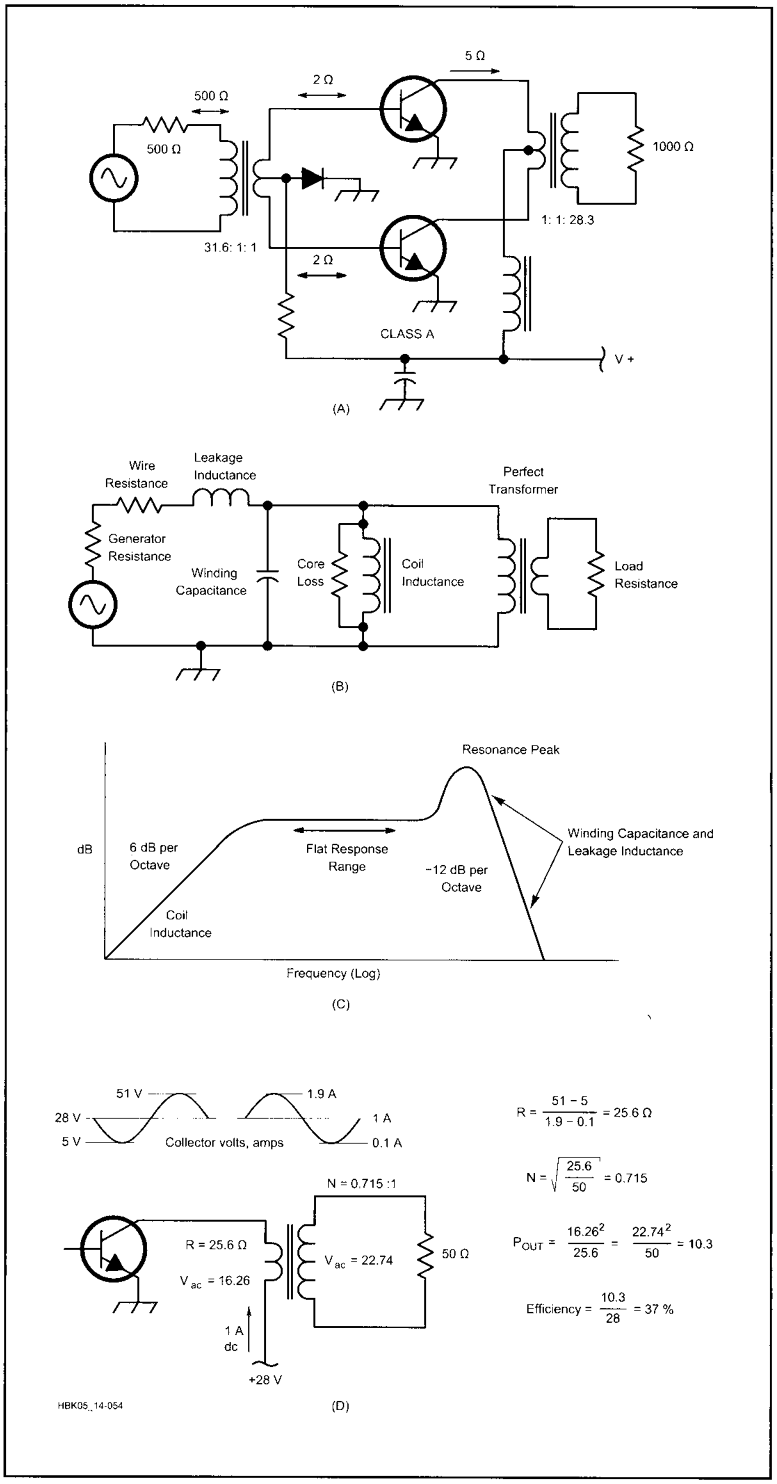


Fig 14.54 — Conventional transformers in an RF power amplifier. Leakage reactances, stray capacitances and core magnetizations limit the bandwidth and linearity, and also create resonant peaks. D shows design example of transformer-coupled RF power amplifier.

Transmission Line Transformers

The basic transmission line transformer, from which other transformers are derived, is the 1:1 choke (or current) balun, shown in **Fig 14.55A**. We consider the following basic properties:

• A pair of close-spaced wires or a length of coax (ie, a transmission line) wraps around a ferrite rod or toroid or through a number of beads. For the 3.5 to 29.7 MHz band, type 43 ferrite (μ = 850), or equivalent, is usually preferred. Other types such as 77 (at 1.8 MHz, μ = 2000) or 61 (at VHF bands, μ = 120) are used. The Z_0 of the line should equal R.

• Because of the ferrite, a large impedance exists between points A and C and a virtually identical impedance between B and D. This is true for parallel wires and it is also true for coax. The ferrite affects the A to C impedance of the coax inner conductor and the B to D impedance of the outer braid equally.

• The conductors (two wires or coax braid and center-wire) are tightly coupled by electromagnetic fields and therefore constitute a good conventional transformer with a turns ratio of 1:1. The voltage from A to C is equal to and in-phase with that from B to D. These are called the *common-mode voltages* (CM).

• A common-mode (CM) current is one that has the same value and direction in both wires (or braid and center wire). Because of the ferrite, the CM current encounters a high impedance that acts to reduce (choke) the current. The normal differential-mode (DM) signal does not encounter this CM impedance because the electromagnetic fields due to equal and opposite currents in the two conductors cancel each other at the ferrite, so the magnetic flux in the ferrite is virtually zero.

• The main idea of the transmission line transformer is that although the CM impedance may be very large, the DM signal is virtually unopposed, especially if the line length is a small fraction of a wavelength.

• A common experience is a CM current that flows on the outside of a coax braid due to some external field, such as a nearby antenna or noise source. The balun reduces (chokes) the CM current due to these sources. But it is very important to keep in mind that the common-mode voltage across the ferrite winding that is due to this current is efficiently coupled to the center wire by conventional transformer action, as mentioned before and easily verified. This equality of CM voltages, and also CM impedances, reduces the *conversion* of a CM signal to an *undesired* DM signal that can interfere with the *desired* DM signal in both transmitters and receivers.

• The CM current, multiplied by the CM impedance due to the ferrite, produces a CM voltage. The CM impedance has L and C reactance and also R. So L, C and R cause a broad parallel self-resonance at some frequency. The R component also produces some dissipation (heat) in the ferrite. This dissipation is an excellent way to dispose of a small amount of unwanted CM power.

• The main feature of the ferrite is that the choke is effective over a bandwidth of one, possibly two decades of frequency. In addition to the ferrite choke balun, straight or coiled lengths of coax (no core and almost no CM dissipation) are used within narrow frequency bands. A one-quarter-wave length of transmission line is a good choke balun at a single frequency or within a narrow band.

• The two output wires of the balun in Fig 14.55A have a high impedance with respect to, and are therefore “isolated” from, the generator. This feature is very useful because now any point of R at the output can be grounded. In a well-designed balun circuit *almost* all of the current in one conductor returns to the generator through the other conductor, despite this ground connection. Note also that the ground connection introduces some CM voltage across the balun cores and this has to be taken into account. This CM voltage is maximum if point C is grounded. If point D is grounded and if all “ground” connections are at the same potential, which they often are not, the CM voltage is zero and the balun may no longer be needed. In a coax balun the return current flows on the inside surface of the braid.

We now look briefly at a transmission line transformer that is based on the choke balun. Fig 14.55B shows two identical choke baluns whose inputs are in parallel and whose outputs are in series. The output voltage amplitude of each balun is identical to the common input, so the two outputs add in-phase (equal time delay) to produce twice the input voltage. It is the high CM impedance that makes this voltage addition possible. If the power remains constant the load current must be one-half the generator current, and the load resistor is 2V/0.5I = 4V/I = 4R.

The CM voltage in each balun is V/2, so there is some flux in the cores. The right side floats. This is named the *Guanella* transformer. If Z_0 of the lines equals 2R and if the load is pure resistance 4R then the input resistance R is independent of line length. If the lines are exactly one-quarter wavelength, then $Z_{IN} = (2R)^2 / Z_L$, an impedance inverter, where Z_{IN} and Z_L are complex. The quality of balance can often be improved by inserting a 1:1 balun (Fig 14.55A) at the left end so that both ends of the 1:4 transformer are floating and a ground is at the far left side as shown. The Guanella can also be operated from a grounded right end to a floating left end. The 1:1 balun at the left then allows a grounded far left end.

Fig 14.55C is a different kind, the *Ruthroff* transformer. The input voltage V is divided into two equal in-phase voltages AC and BD (they are tightly coupled), so the output is V/2. And because power is constant, $I_{OUT} = 2I_{IN}$ and the load is R/4. There is a CM voltage V/2 between A and C and between B and D, so in normal operation the core is not free of magnetic flux. The input and output both return to ground so it can also be operated from right to left for a 1:4 impedance stepup. The Ruthroff is often used as an amplifier interstage transformer, for example between 200 Ω and 50 Ω. To maintain low attenuation the line length should be much less than one-fourth wavelength at the highest frequency of operation, and its Z_0 should be R/2. A balanced version is shown in Fig 14.55D, where the CM voltage is V, not V/2, and transmission is from left-to-right only. Because of the greater flux in the cores, no different than a conventional transformer, this is not a preferred approach, although it could be used with air wound coils (for example in antenna tuner circuits) to couple 75 Ω unbalanced to 300 Ω balanced. The tuner circuit could then transform 75 Ω to 50 Ω.

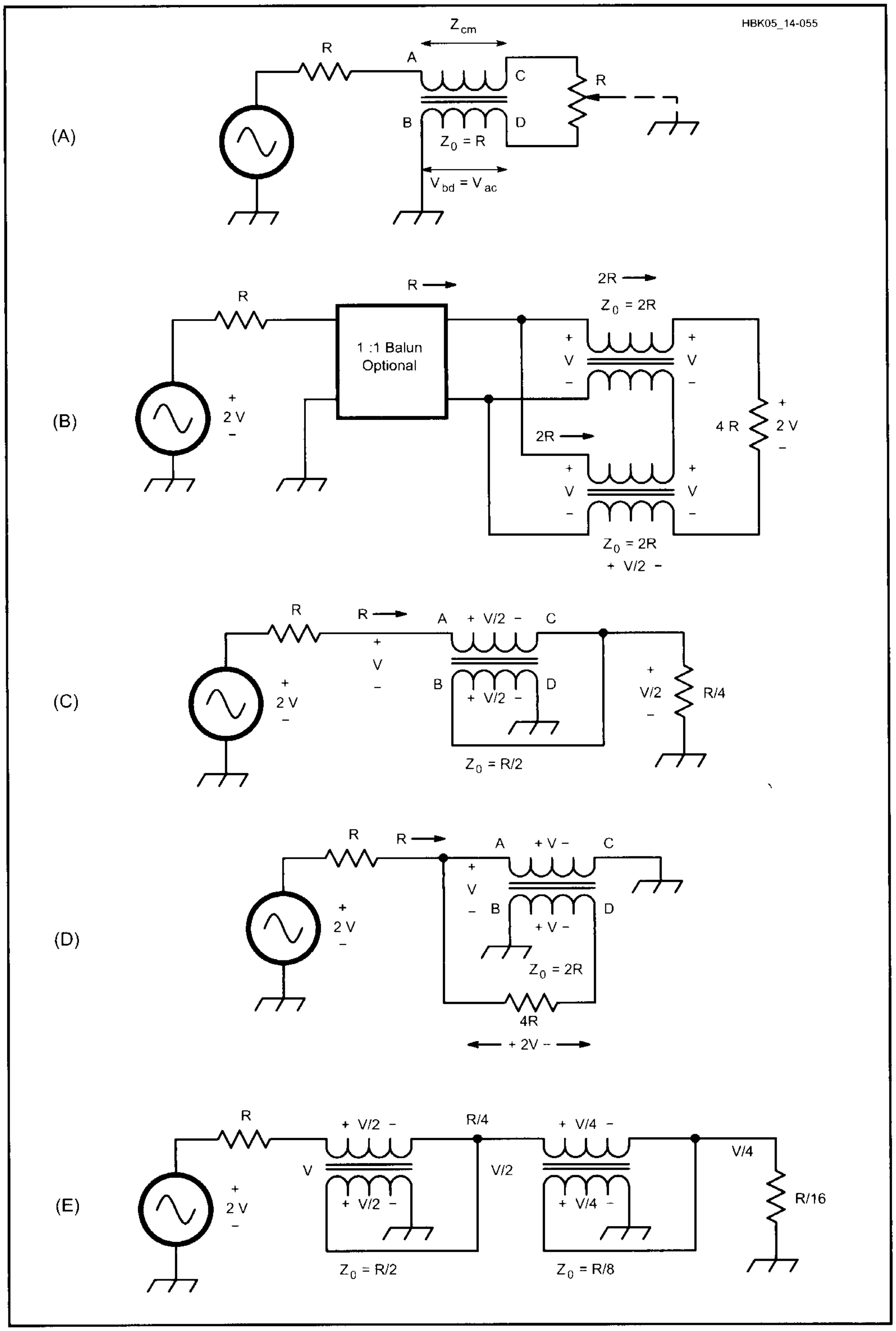


Fig 14.55 — At A, basic balun. At B, 1:4 Guanella transformer. At C, Ruthroff transformer, 4:1 unbalanced. At D, Ruthroff 1:4 balanced transformer. At E, Ruthroff 16:1 unbalanced transformer.

Fig 14.56 illustrates, in skeleton form, how transmission-line transformers can be used in a push-pull solid state power amplifier. The idea is to maintain highly balanced stages so that each transistor shares equally in the amplification in each stage. The balance also minimizes even-order harmonics so that low-pass filtering of the output is made much easier. In the diagram, T1 and T5 are current (choke) baluns that convert a grounded connection at one end to a balanced (floating) connection at the other end, with a high impedance to ground at both wires. T2 transforms the 50 Ω generator to the 12.5 Ω (4:1 impedance) input impedance of the first stage. T3 performs a similar step-down transformation from the collectors of the first stage to the gates of the second stage. The MOSFETs require a low impedance from gate to ground. The drains of the output stage require an impedance step up from 12.5 Ω to 50 Ω, performed by T4. Note how the choke baluns and the transformers collaborate to maintain a high degree of balance throughout the amplifier. Note also the various feedback and loading networks that help keep the amplifier frequency response flat.

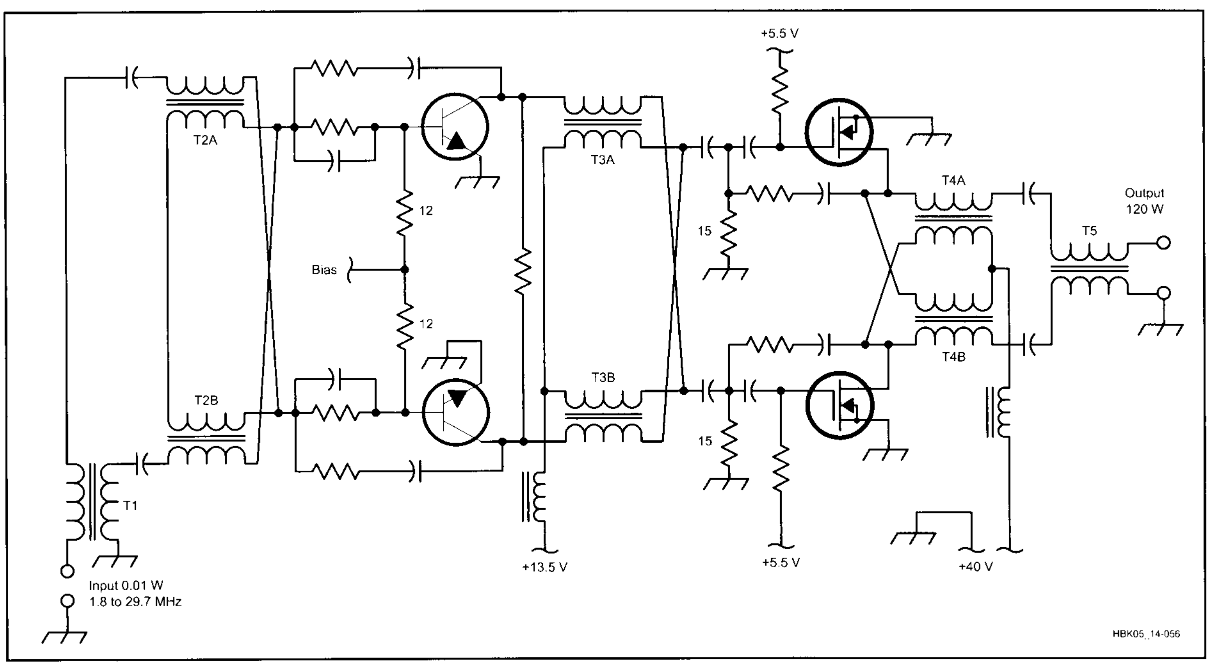


Fig 14.56 — This illustrates how transmission-line transformers can be used in a push-pull power amplifier.

Tips on Toroids and Coils

Some notes about toroid coils: Toroids do have a small amount of leakage flux, despite rumors to the contrary. Toroid coils are wound in the form of a helix (screw thread) around the circular length of the core. This means that there is a small component of the flux from each turn that is perpendicular to the circle of the toroid (parallel to the axis through the hole) and is therefore not adequately linked to all the other turns. This effect is responsible for a small leakage flux and the effect is called the "one-turn" effect, since the result is equivalent to one turn that is wound around the outer edge of the core and not through the hole. Also, the inductance of a toroid can be adjusted, also despite rumors to the contrary. If the turns can be pressed closer together or separated a little, inductance variations of a few percent are possible.

A grounded aluminum shield between adjacent toroidal coils can eliminate any significant capacitive or inductive (at high frequencies) coupling. These effects are most easily noticed if a network analyzer is available during the checkout procedure, but how many of us are that lucky? Spot checks with an attenuator ahead of a receiver that is tunable to the harmonics are also very helpful.

There are many transformer schemes that use the basic ideas of Fig 14.55. Several of them, with their toroid winding instructions, are shown in **Fig 14.57**. Because of space limitations, for a comprehensive treatment we suggest Jerry Sevick's books *Transmission Line Transformers* and *Building and Using Baluns and Ununs*. For applications in solid-state RF power amplifiers, see Sabin and Schoenike, *HF Radio Systems and Circuits*, Chapter 12.

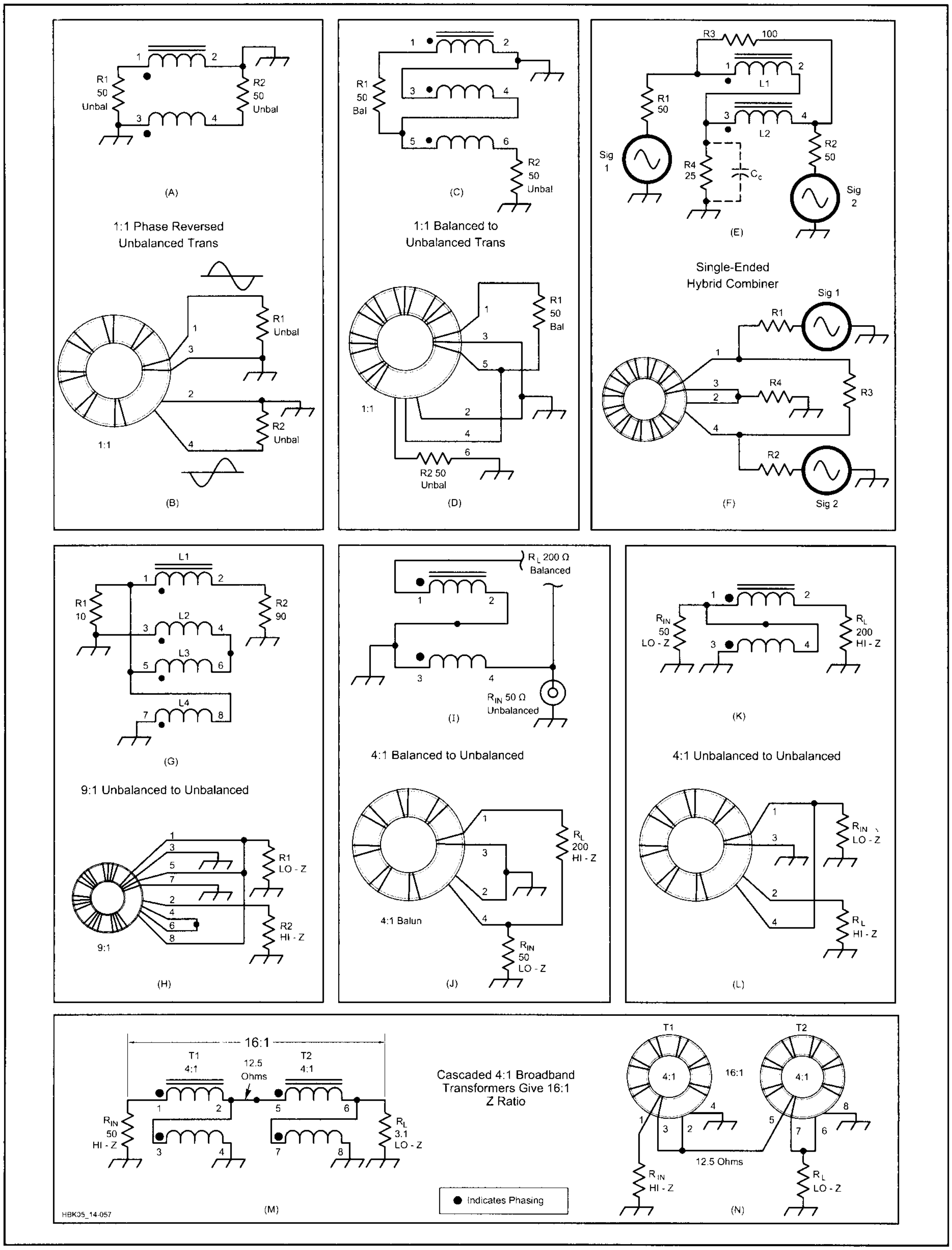


Fig 14.57 — Assembly instructions for some transmission-line transformers. See text for typical magnetic materials used.

Modules in Combination

Quite often the performance of a single stage can be greatly improved by combining two identical modules. Because the input power is split evenly between the two modules the drive source power can be twice as great and the output power will also be twice as great. In transmitters, especially, this often works better than a single transistor with twice the power rating. Or, for the same drive and output power, each module need supply only one-half as much power, which usually means better distortion performance. Often, the total number of stages can be reduced in this manner, with resulting cost savings. If the combining is performed properly, using hybrid transformers, the modules interact with each other much less, which can avoid certain problems. These are the system-design implications of module combining.

Three methods are commonly used to combine modules: parallel (0°), push-pull (180°) and quadrature (90°). In RF circuit design, the combining is often done with special types of "hybrid" transformers called splitters and combiners. These are both the same type of transformer that can perform either function. The splitter is at the input, the combiner at the output. We will only touch very briefly on these topics in this chapter and suggest that the reader consult the **RF Power Amplifiers** chapter and the very considerable literature for a deeper understanding and for techniques used at different frequency ranges.

Fig 14.8 illustrates one example of each of the three basic types. In a 0° hybrid splitter at the input the tight coupling between the two windings forces the voltages at A and B to be equal in amplitude and also equal in phase if the two modules are identical. The 2R resistor between points A and B greatly reduces the transfer of power between A and B via the transformer, but only if the generator resistance is closely equal to R. The output combiner separates the two outputs C and D from each other in the same manner, if the output load is equal to R, as shown. No power is lost in the 2R resistor if the module output levels are identical.

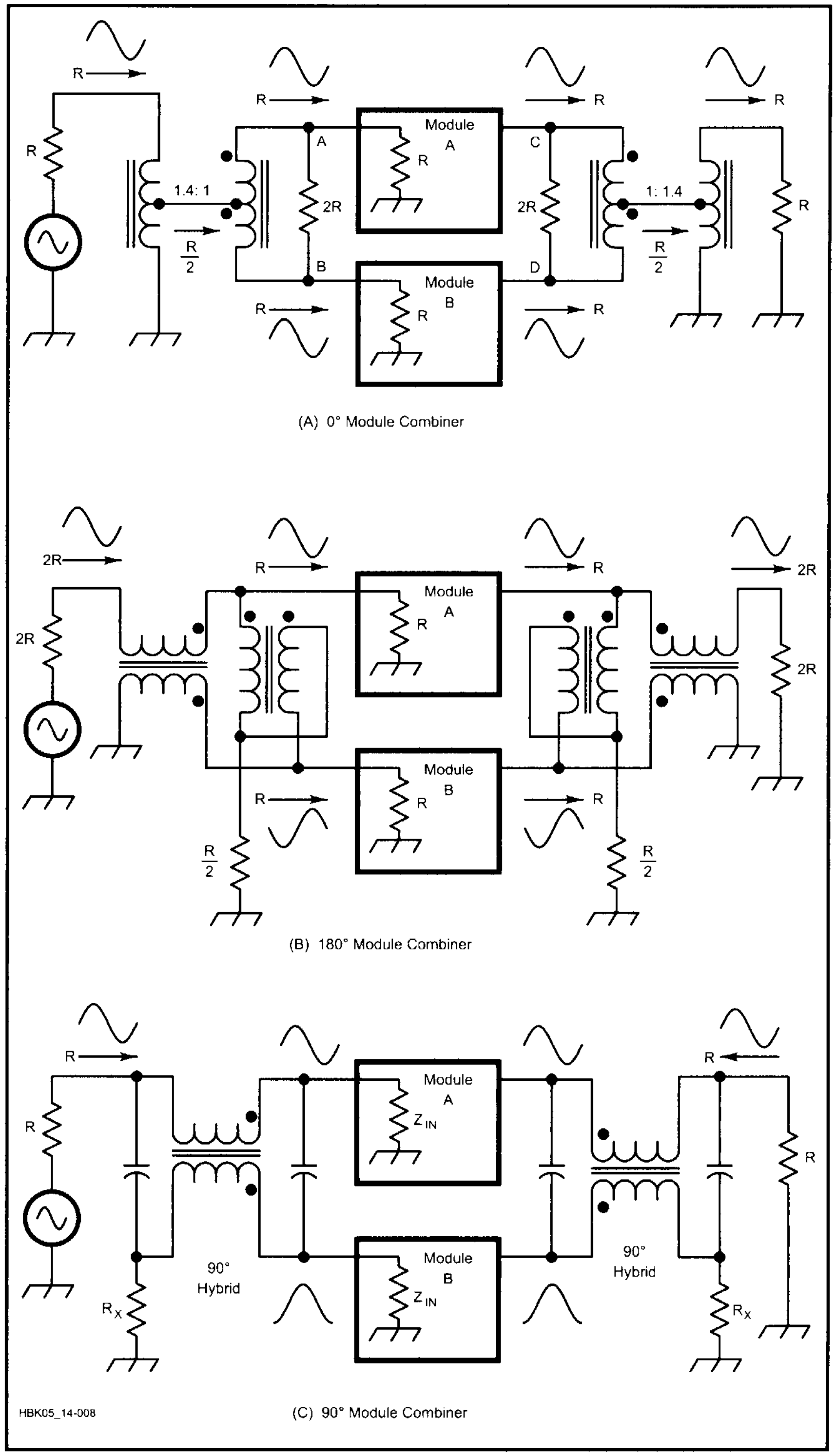
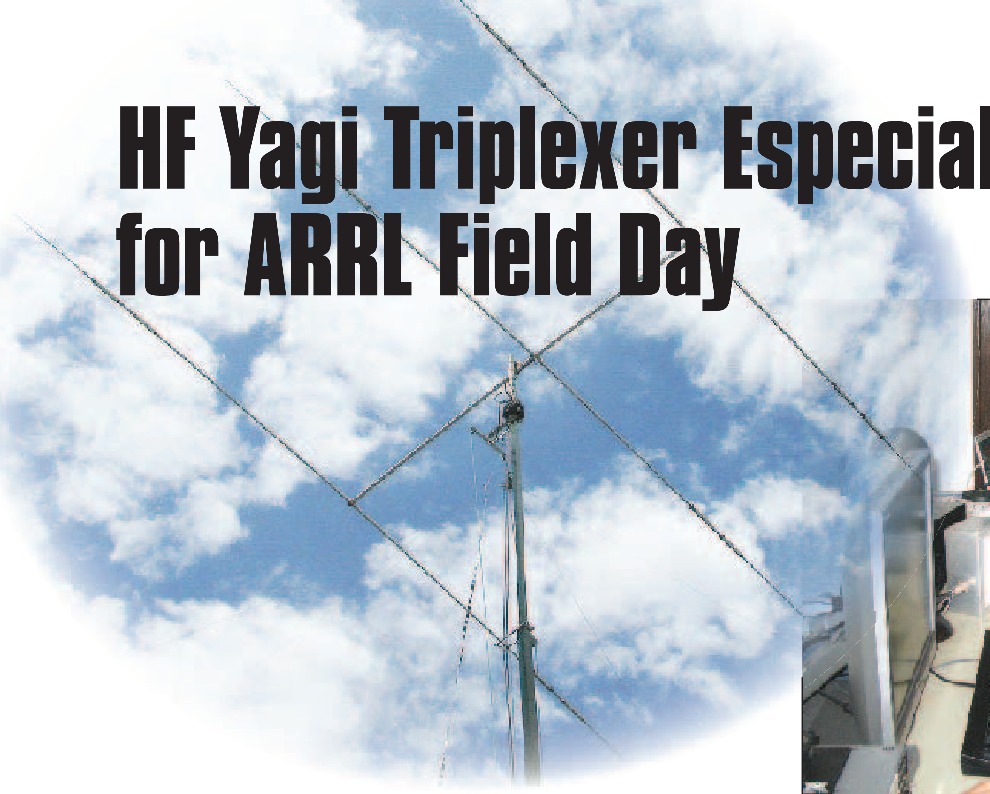


Fig 14.8 — The three basic techniques for combining modules.

The 180° hybrid produces push-pull operation. The advantages of push pull were previously discussed. The horizontal transformers, 1:1 balun transformers, allow one side of the input and output to be grounded. The R/2 resistors improve isolation between the two modules if the 2R resistors are accurate, and dissipate power if the two modules are not identical.

In a 90° hybrid splitter, if the two modules are identical but their identical input impedance values may not be equal to R, the hybrid input impedance is nevertheless R Ω, a fact that is sometimes very useful in system design. The power that is "reflected" from the mismatched module input impedance is absorbed in RX, the "dump" resistor, thus creating a virtual input impedance equal to R. The two module inputs are 90° apart. At the output, the two identical signals, 90° apart, are combined as shown and the output resistance is also R. This basic hybrid is a narrowband device, but methods for greatly extending the frequency range are in the literature (W. Hennigan, W3CZ, "Broadband Hybrid Splitters and Summers," Oct 1979 *QST*). One advantage of the 90° hybrid is that catastrophic failure in one module causes a loss of only one half of the power output.



HF Yagi Triplexer Especially for ARRL Field Day

K6KV

Figure 1 — Kenneth Finnegan, W6KWF, in foreground and Phil Verinsky, W6TQG, at Phil's station, competing in the July 2009 NAQP contest in the multioperator two transceiver category. Using the triplexer standing on the table between them, Kenneth and Phil operate on both 20 and 15 meters sharing Phil's triband antenna.

This easy-to-build project lets up to three transceivers on 10, 15, and 20 meters share the same antenna.

Gary Gordon, K6KV

It all started during a WVARA ARRL Field Day discussion.[1] Svend Jensen, KF6EMB, was asking Jim Peterson, K6EI: "If a triband Yagi works on three bands, why do three transceivers need three separate antennas? Why can't they all share the same antenna?"

Whoa — connect my receiver to your transmitter's antenna? It sounded like asking for big trouble. Even with separate antennas, just having one station near to another can cause plenty of interference. Unless precautions are taken, signals from one station will invariably find their way into the other and cause overload or possibly damage.

Fortunately, the nearby radio problem was solved years ago with the introduction of band-pass filters. You connect one between each rig and its antenna, and they'll block signals on other bands from getting through. It's little surprise they've become standard fare on most contest outings.

You might wonder if simply paralleling several band-pass filters together might allow different rigs to share the same antenna. Unfortunately it won't, partly because more isolation is required but mainly because their design is such that they'd simply short out each other's signals. What will work, however, is using a decoupling network in conjunction with band-pass filters as described in this article.

This article describes two designs, one rated for 5 or 10 W and the other for 100 W. Both use commercial band-pass filters to greatly simplify the project. Each has insertion losses, including the band-pass filters, of less than 1 dB. The 100 W triplexer was contest proven during the 2009 NAQP RTTY and SSB contests at W6TQG (see Figure 1). Both versions were used during WVARA's 2009 QRP Field Day. With a triplexer, every transceiver could be operated as though it owned the antenna, even to the point of tuning it for minimum SWR. There was never a hint of interference, and all that was left for the operators to decide was where to point the antenna. One limitation, of course, is that it a single antenna can only point in one direction at a time.

Rig Protection

By now you might be asking, just how safe is my receiver? The short answer is, as long as the band-pass filters stay connected to the decoupling network of this article, it's virtually impossible to come up with any scenario that could damage a transceiver. Rigs are generally safe with overloads up to approximately 1 W of RF power at their antenna connectors; crosstalk from using a triplexer is far below this level. Under the worst conditions with 100 W transmitters, a receiver will never see more than 2 mW of RF, or 1/500 of the damage level.

During normal operation, signals from other transmitters are attenuated approximately 50 dB through the action of both the band-pass filters and the decoupling network portion of the triplexer. Although not immediately obvious, the risk is equally benign should you inadvertently tune your receiver to a band where another rig is transmitting, because 2 mW of RF just isn't hazardous. If another rig inadvertently transmits on your band, your receiver will see even less than 2 mW, because besides the normal isolation the triplexer provides, the offending rig will shut down from being unable to find an impedance match.

If a component fails or arcs over, perhaps caused by running excessive transmitter power, a resonant circuit in the triplexer will either short out or open up. Artificially introducing these failures only lowered the crosstalk, which never measured more than 0.1 mW.

We can also calculate the stress on the feed line. With 100 W PEP transmitters, the feed line might see average power levels around 100 W, and peak potentials of 300 to 500 V.

[1]Notes appear on page 40.

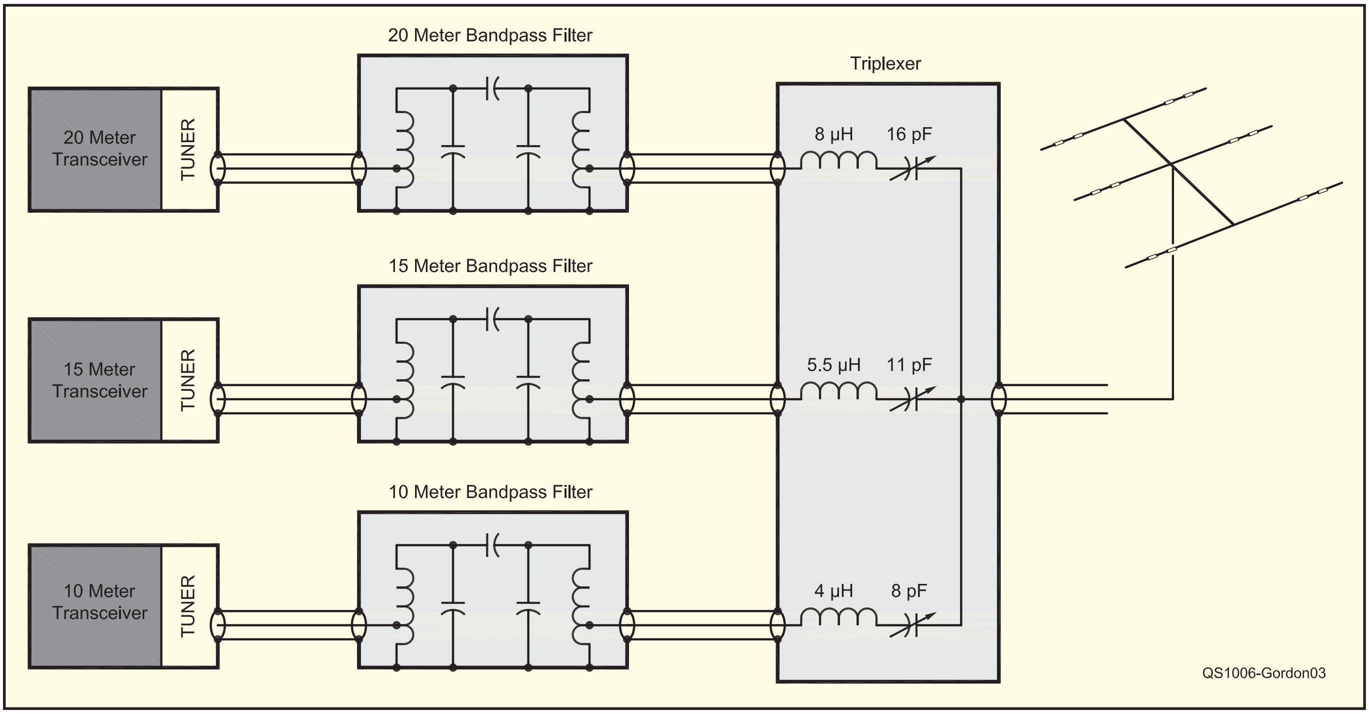


Figure 2 — Triplexer schematic. See the article regarding choosing the capacitors and winding the inductors.

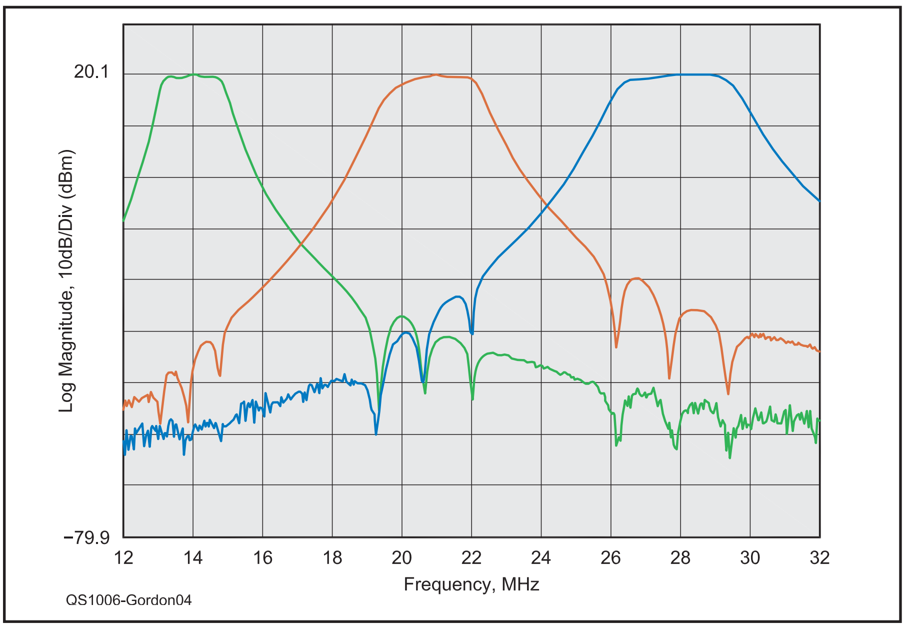


Figure 3 — Frequency response of the triplexer decoupling network and band-pass filters. Next-band signals are attenuated by approximately 50 dB.

These are safely within the 430 W rating for 0.2 inch diameter RG-58/U coax cable and the 1000 V rating for PL-259 UHF connectors.

Note that the portion of this project that you can build, the decoupling network box, is not the entire triplexer, and will not by itself protect transceivers. Only when band-pass filters are connected to it will the transceivers be safe. If you plan to try any automatic antenna or band switching, then be sure to do it outside this regime. That said, it seems inconceivable that any component failure, loose connector or band switching mistake could put a transceiver at risk. With this setup, a receiver should never be subjected to a power level any stronger than 2 mW.

Circuit Description

The decoupling network portion of the triplexer uses three series resonant circuits, one between each input and the common antenna feed line connector (see Figure 2). For example, the 20 meter (top) series tuned circuit is tuned to resonance at 14 MHz to pass signals on the 20 meter band while attenuating signals on all other bands. Each resonant circuit has a loaded Q of 5, chosen to keep insertion losses low while providing sufficient other band attenuation. Figure 3 shows the frequency response for the three channels of the triplexer with Dunestar model 300s band-pass filters, as measured on an HP 2588A spectrum analyzer.[2]

Table 1 shows the insertion losses at five frequencies across the 20 meter band, measured using a JRC JST-245 transceiver, a Daiwa CN-620B power meter, and an MFJ-264 dummy load. Except for the top end of the 10 meter band the insertion loss never exceeded 1.0 dB, one sixth of an S-unit. Figure 4 shows the SWR measurements for the three bands, which rarely exceeded 1.5:1, as measured using an HP 8591E Spectrum Analyzer and a directional coupler.

Construction

Figure 5 shows a 100 W version of the triplexer. A 5 W version was also built. The only differences between them are the voltage ratings of the variable capacitors and the way they are mounted. Either version can be built in a die cast or other metal enclosure that provides shielding and ground continuity between the inputs and the output. An example is the Hammond 1590E boxes that are available as Digi-Key part number HM155. The input and output connectors are SO-239 UHF jacks. Their 5/8 inch mounting holes can easily be drilled using a step drill or 5/8 inch countersink.

Table 1

20 Meter Insertion Losses for the 100 W Triplexer Decoupling Network and Band-pass Filter

Frequency (MHz)	*Transmit Power (W)*	*Filter Loss (dB)*	*Triplexer Loss (dB)*	*Total Loss (dB)*	*Power Out (W)*
14.00	175	0.70	0.30	1.00	139
14.10	175	0.68	0.29	0.97	140
14.20	175	0.68	0.29	0.97	140
14.30	175	0.68	0.30	0.98	139.5
14.35	175	0.73	0.27	1.00	139

Table 2

Winding Specifications for Inductors Suitable for Power Levels up to 150 W

See text. All coils 1 inch in diameter.

Inductor	*Band (Meters)*	*Inductance (μH)*	*Turns*	*Wire Size (AWG)*	*Winding Pitch (TPI)*
L1	20	8	19.5	20	17.5
L2	15	5.3	13.5	18	13
L3	10	4	12	16	11

Inductor Details

The inductors can be wound on 2 inch lengths of 1 inch polycarbonate tubing (not brittle polystyrene), available from Tap Plastics (**www.tapplastics.com**). Small holes can be drilled near the ends of the forms to provide for anchoring the ends of the windings. Short lengths of ¾ inch polycarbonate rod can be glued into one end and tapped for mounting screws.

The wire can be bare, enameled, tinned, or Teflon insulated wire, depending on what is available. The suggested wire gauges in Table 2 do not need to be followed exactly. A wire kit is being offered by VE2VBR that consists of 25 foot lengths each of #16, 18 and 20 AWG wire. Alternately, short lengths may be available on Internet auction sites and in stores.[3]

If your wire is either bare or has thin insulation, the coil turns should to be spaced apart slightly. You can space the coil turns by interspersing fishing leader between them, added after the wire has been wound, and then coating everything with an adhesive. For my inductors I used a threading lathe to cut a shallow 0.03 inch spiral groove across much of the length to guide the wire, something a machinist could do in half an hour. In either case winding the coil is much easier if the wire is first straightened. Cut generous lengths, perhaps 7 feet for the 10 meter coil and 10 feet for the 20 meter coil, anchor one end to a vice or anything rigid and then use pliers to give the other end a sharp yank, stretching the wire an inch or so. If your wire has Teflon insulation, the turns do not need to be spaced apart.

Capacitor Details

Look for variable capacitors with maximum capacitances ideally of 20 or 30 pF or slightly more.[4] For the 5 W version) the first choice is the miniature APC style. These variable capacitors are compact, inexpensive, relatively available as industrial and military surplus, and adjusted using a screwdriver. Look for capacitors with 500 V or higher ratings, such as the Fair Radio Sales 35 pF 1G-35.

Higher power triplexers call for capacitors with larger plate spacings since, for example to handle 100 W, they will be subjected to at least 1200 V peak. One excellent choice for this power level is the RF Parts 41 μF 48APL41S.

Table 2 lists the recommended voltage ratings for different power levels, and how to estimate the voltage rating based on the plate spacing. If you come across capacitors with higher than optimum maximum capacitances, you might consider modifying them by removing either rotor or stator plates by sawing off a fraction of the lengths. I'll mention in passing that a completely different but more complicated way to make tuned circuits is using fixed capacitors and toroidal inductors that are adjusted by sliding their turns, as described in a *QST* article.[5]

The variable capacitors are not panel mounted but instead insulated from the metal box. In the 100 W version shown in Figure 5, they are mounted to a 1 inch wide strip of aluminum. The aluminum strip is insulated from the box by stand off insulators such as the threaded phenolic spacers manufactured by Keystone. These are available in the ¾ × ¼ inch size from Digikey as part number 386K. This mounting method also provides electrical continuity between the three capacitors. The previously suggested 41 pF part has a ¼ × ⅜ inch shaft, which should be slotted for screwdriver adjustment access through ¼ inch holes drilled into the box. Recessing the adjustment in this manner will discourage knob twiddling, since once the capacitors have been initially tuned they never again need to be touched.

Band-pass Filters

To complete the project you'll need a set of three 40 dB band-pass filters. These are available commercially, or can be home made.[6-9] If you really want to build your own filters I recommend using the step-up transformer approach used in the commercial products and in the referenced articles in order to achieve sufficient Q, although my strongest recommendation is to keep this a simple project by borrowing or buying a set.

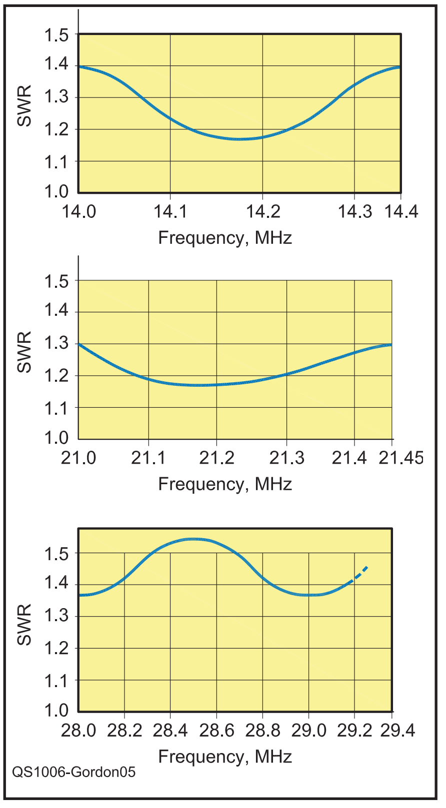


Figure 4 — SWR measurements for the triplexer decoupling network and band-pass filters

Note that band-pass filters have their own power limitations. Those manufactured by Dunestar are rated for intermittent use with transceivers up to 200 W, and with the additional advice that they be operated into low SWRs and with an antenna always connected.

I must reemphasize that in order to protect the transceivers from damage, band-pass filters must be attached to the decoupling network box of this article, and a note to this effect should be placed on your unit. To ensure they stay a permanent part, screw their free ends down to a strip of wood or metal, which also protects connectors from getting broken if

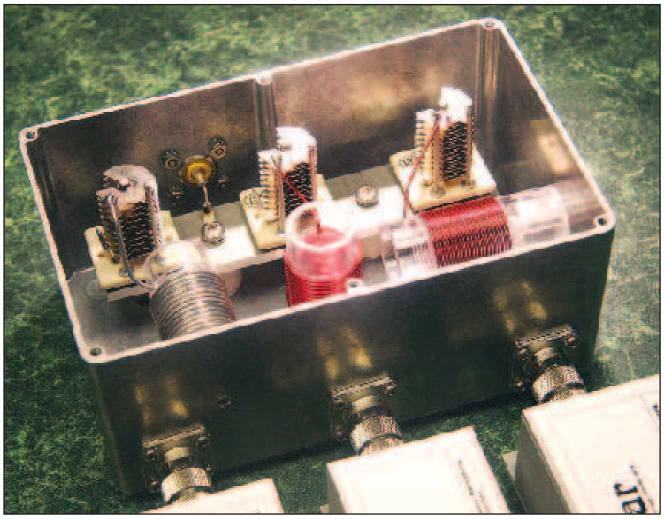


Figure 5 — The 100 W triplexer decoupling network with its cover removed.

Table 3

Recommended Minimum Capacitor Ratings for Different Transmitter Power Levels

Transmit Power (W)	*Capacitor Rating (V)*	*Plate Spacing (inches)*
150	2500	0.062
100	2000	0.050
25	1000	0.025
6	500	0.015

cables get yanked on. It would seem a good idea to also label the outboard end of each filter's feed line with its intended band.

Adjustment

Tuning the triplexer's decoupling network consists of adjusting each tuned circuit for resonance on its particular band. The easiest way is to do this is to connect it to a dummy load and adjust the capacitors for maximum power to the load. Start by connecting a transceiver to the input of the 20 meter band-pass filter and the triplexer's output to a power meter and dummy load. You can set your transceiver frequency to either the center of the band or in your favorite segment, although its bandwidth will be wide enough to cover the entire band. Set a low power level, and turn off any antenna tuner.

With the power meter set to a sensitive range, transmit a CW signal and adjust C1 for maximum power using an insulated screwdriver. If you are using an analog power meter, increase the transmitter power if necessary to move the needle up the scale and make the adjustment more sensitive.

The 15 and 10 meter circuits are tuned in a like manner. If your three variable capacitors are identical you should see their engagement angles look something like those in Figure 6, where the capacitance values will be roughly 8, 11, and 16 pF. As a final check of performance, confirm that the insertion losses are not higher than approximately 1 dB (20%), by comparing dummy load power readings with and without the triplexer.

First On Air Test

Since the concept was inspired by ARRL Field Day 2009, the first tests took place at that event, at an open space preserve atop Mora Hill in Cupertino, California. Needing to adjust to the park ranger's new and more restrictive rules regarding how much aluminum we could have, we decided on two gain antennas, one triband Yagi each for CW and SSB. Since the skip was always to the East, neither antenna needed to be rotated, although that advantage might not apply to your location. Each antenna ended up servicing three separate transceivers, giving us nearly the firepower of six monoband Yagis. We placed 12th overall nationally, second in our division and third nationally in the low power (QRP) category.

No Interference (Really?)

Before Field Day arrived, Jim Peterson, W6EI, and I ran a series of interference tests using six different midrange transceivers each running 100 W. The short takeaway is that as long as his triband Yagi was connected we saw no interference whatsoever, regardless of how we mixed and matched the rigs and bands.

That made us curious as how much safety margin existed, and since it wasn't possible to increase transmitter power, we decided instead to eliminate atmospheric noise by replacing his antenna with a dummy load. While this represented an artificial situation, nonetheless it was a useful way to expose underlying interference issues, which we did see in most cases. The only exception was an absence of any interference between two Elecraft K3 transceivers.

Interference can be caused by both transmitters and receivers. If you're looking for a clean transmitter, look for one with low phase noise, as revealed by having low "composite transmitted noise" in the ARRL Product Reviews. For a resistant receiver, according to Elecraft, look for one that has a high out-of-band signal rejection, as measured by its second order intercept point (IP2). That and other useful transceiver specifications can be found at **www.elecraft.com/K2_perf.htm**. Nonetheless, while it's smart to test one's equipment before any important event, you should not expect to encounter any interference whatsoever when using this triplexer with modern transceivers.

Conclusion

This is an easy to build construction project that will reduce the amount of antenna hardware you'll need for your next multioperator contest. I look forward to others sharing their experiences and improving upon the design.

I am pleased to acknowledge the encouragement and testing help provided by Jim Peterson, K6EI, and to Rene Morris, K6XW, of Elecraft for the frequency response test of Figure 3.

Notes

[1]WVARA, West Valley Amateur Radio Association, San Jose, California. **www.wvara.org**.
[2]**www.dunestar.com**
[3]Wire: Conception R.B., **www.conceptionrb.com/boutique/index.php?cPath=46**, **www.conceptionrb.com/boutique>English>Catalog**.
[4]Variable capacitors: RF Parts, Fair Radio Sales (see their 1G-35) and eBay.
[5]E. Wetherhold, W3NQN, "Clean Up Your Signals with Band Pass Filters," Part 1, *QST*, May 1998, pp 44-51, Part 2, *QST,* Jun 1998, pp 39-42.
[6]**www.arraysolutions.com/Products/wx0bbpf6.htm**
[7]See Note 2.
[8]E. Wetherhold, W3NQN, "Receiver Band-Pass Filters Having Maximum Attenuation in Adjacent Bands," *QEX,* Jul 1999, pp 27-30.
[9]See Note 5.

ARRL member Gary Gordon, K6KV, took a liking to electricity in grade school after reading a copy of The Boy Electrician. *A few years later in Milwaukie, Oregon he built his first Tesla coil and in 1955 became WN7ZKG. The following year, finding war surplus parts both cheap and plentiful, he picked up several 803 pentodes for $.50 apiece and built a 300 W linear amplifier. That, a full size antenna and a swamp front location made for a quite respectable signal on 80 meters.*

Gary went on to study electrical engineering at University of California at Berkeley, and Stanford University. He then enjoyed a productive career at Hewlett Packard where, among other things, he co-invented the optical computer mouse. You can reach Gary at 21112 Bank Mill Rd, Saratoga, CA 95070 or at **gary1@gary-gordon.com**.

Measuring Ferrite Chokes

By Jim Brown, K9YC

A ferrite choke is a parallel resonant circuit, formed by inductance and resistance coupled from the core and stray capacitance that results from interaction of the conductor that forms the choke with the permittivity of the core, and, if the choke is wound, the capacitance between turns.

These chokes are very difficult to measure for two fundamental reasons.

1) The stray capacitance that forms the parallel resonance is quite small (typically 0.4 – 5 pF), which is often less than the stray capacitance of the test equipment used to measure it.
2) Most instrumentation measures RF impedances by measuring the reflection coefficient in a 50-Ω circuit.

Reflection-based measurements have increasingly poor accuracy when the unknown impedance is more than about 3x the impedance of the analyzer, because the value of the unknown is computed by differencing analyzer data, and even very small errors in the raw data cause very large errors in the computed result. While the software used with reflection-based systems use calibration and computation methods to remove systemic errors like fixture capacitance from the measurement, these methods have generally poor accuracy when the impedance being measured is in the range of typical ferrite chokes.

A simple method that works

The key to accurate measurement of high impedance ferrite chokes is to set up the choke as the series element of voltage divider (see Figure 1), using a well-calibrated voltmeter to read the voltage across a well-calibrated resistor that acts as the voltage divider's load resistor.

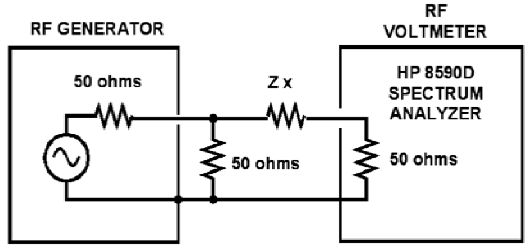


Figure 1 – Choke measurement setup

The fundamental assumption of this measurement method is that the unknown impedance is much, much higher than the impedance of both the generator and the load resistor. The voltage divider is driven by an RF generator, which must be terminated by its calibration impedance (because the generator is calibrated only when working into its calibration impedance, and the test circuit is a high impedance). An RF spectrum analyzer with its own internal termination resistor can serve as both the voltmeter and the load. Alternatively, a simple RF voltmeter or scope can be used, with the calibrated load impedance being provided by a termination resistor of known value in the frequency range of the measurement.

Making measurements

With the unknown in place, obtain values for the voltage across the load resistor as the generator in increments of about 5% over the range of interest, and record the data in a spreadsheet. Use the same frequencies for all chokes so that data can be plotted and compared. Using the spreadsheet, solve the voltage divider equation backwards to find the unknown impedance. Plot the data as a graph of impedance (on the vertical axis) vs frequency (on the horizontal axis). Scale both axes to display logarithmically.

Accuracy

This method yields the magnitude of the impedance, but no phase information. Accuracy is greatest for large values of unknown impedance (worst case 1% for 5,000 ohms, 10% for 500 ohms). Accuracy can be further improved by correct for variations in the loading of the generator by the test circuit. Measure the voltage at the generator output with the unknown connected and use that voltage as the input voltage. The voltmeter must be un-terminated for this measurement.

Obtaining R, L and C values

In a second spreadsheet (or another page in your measurement spreadsheet), create a new table that computes the magnitude of the impedance of a parallel resonant circuit for the same range of frequencies as your choke measurements. Set up the spreadsheet so that you enter values for R, L, and C at the top of the spreadsheet, with the calculations for each frequency based on the values of R, L, and C that you have entered. Also at the top of your spreadsheet, compute the resonant frequency and Q for values you have entered. Create a graph that plots this computed impedance over the same range of frequencies as the measurements, and with the same plotted scale as the measurements.

Now, 1) set R equal to the resonant peak of the measured impedance; 2) pick a point on the resonance curve below the resonant frequency that's about one-third of the impedance at resonance, and compute L for that value of inductive reactance; 3) pick a value for C to produce the same resonant frequency of the measurement. 4) if necessary, adjust the values of L and C until the computed curve most closely matches the measured curve. These R, L, and C values can be used in circuit models (NEC, SPICE) to predict the behavior of circuits using ferrite chokes.

Capacitance errors

This setup can be constructed so that its stray capacitance is small, but it won't be zero. A first approximation of the stray capacitance can be obtained by substituting for the unknown a non-inductive resistor whose resistance is in the same general range as the chokes being measured, then varying the frequency of the generator to find the -3dB point (where $X_C = R$). This test for the setup in the photograph yielded a stray capacitance value of 0.4 pF. **(***A surface-mount film resistor will have the least reactance, but may be impractical due to its small size. Low-wattage carbon composition resistors are acceptable if leads are kept as short as possible – Ed.*)

Accounting for stray capacitance

Since the measured curve includes stray capacitance, the actual capacitance of the choke will be slightly less than the computed value. If you have determined the value of stray for your test setup, subtract it from the computed value to get the actual capacitance. You can also plug this corrected value into the theoretical circuit to see how the choke will actually behave in a circuit (that is, without the stray capacitance of your test setup). You won't see the change in your measured data, only in the theoretical RLC equivalent.

Dual resonances

In NiZn materials (types #61, #43), there is only circuit resonance, but MnZn materials (#77, #78, #31) have both circuit resonance and dimensional resonance. The dimensional resonance of #77 and #78 is rather high Q and clearly defined, so R, L, and C values can often be computed for both resonances. This is not practical with chokes wound on #31, because the dimensional resonance occurs below 5 MHz, is very low Q, is poorly defined, and blends with the circuit resonance to broaden the impedance curve. The result is a dual-sloped resonance curve – that is, curve fitting will produce somewhat different values of R, L, and C when matching the low-frequency slope and high frequency slope. When using these values in a circuit model, use the values that most closely match the behavior of the choke in the frequency range of interest.

Measuring Isolation Between Radios

By George Cutsogeorge, W2VJN

When two or more radios are operated in close proximity, some care is needed to be sure they can coexist without damage. When antennas are close together and high power is being used, there is a danger that a receiver will be damaged by absorbing too much energy. This situation is common on Field Day and DXpeditions where multiple transmitters are operating at the same time. In multiop contest stations and in single op, two radio (SO2R) stations it is also a concern.

Without knowing anything about the receiver's input circuit, it may be a good assumption that it consists of components such as resistors and diodes that are rated at 1/10 to 1/4 watt. To be safe, let's say we would like to input no more than 50 milliwatts to prevent receiver damage. This is about 45 dB below 1500 watts and about 33 dB below 100 watts.

How can we tell if we have interference levels that will cause any of these problems? Do we just fire up our 1500-watt transmitter in radio A and see what happens to radio B? I don't recommend that. All we need are the two radios plus a little arithmetic to figure it out.

The voltage level that causes an S9 reading in most radios is typically 50 microvolts. Let's look at an example of one station, radio A, on 40 meters and a second on 20 meters, radio B. Each is connected to its normal antenna. Radio A uses a 40 meter dipole and radio B has a triband Yagi. Both antennas are on the same tower.

Set up radio A on 40 as the transmitter and radio B as the receiver. Turn the radio A power all the way down before transmitting. Set radio B to 40 meters leaving the tribander connected. Adjust the radio A power output to one watt and tune it in with radio B. Rotate the tribander for the maximum S meter reading on radio B. Add attenuation to radio B so that an S meter reading below full scale occurs. As an example, assume the S meter reads S9+55 dB with 18 dB of attenuation inserted.

The isolation between radios can be calculated with this equation:

Isolation (dB) = Ptx – attn – (–73 dBm) – dB above S9

where

Ptx = transmitter power in dBm (1 watt = +30 dBm)

attn = attenuator setting in dB

–73 dBm = S9 = 50 µV

In this case, isolation = +30 – 18 + 73 – 55 = 30 dB

Clearly radio B is in danger if the transmitted power is 1500 watts, so more isolation is required. This could take the form of filters and/or stubs. For a 100-watt transmitter the isolation is still 3 dB less than we would like. If the tribander is rotated there will be positions that provide a bit more isolation and this may be adequate for a temporary situation.

The same measurements should be made with radio B transmitting on 20, 15 and 10 meters and radio A receiving on the same band. Since a 40 meter dipole is a good 15 meter antenna, this will most likely be the band with the least isolation.

W1GHZ

MICROWAVELENGTHS

Microwave Plumbing

For the most popular microwave bands, 10 GHz and lower, waveguides are too bulky and difficult to use inside our equipment. RF connections are made with coaxial cable, but ordinary coax, with braided shield, is lossy and the familiar PL-259 coax connectors have poor performance.

The cable of choice is semirigid 50 Ω coax, which has a copper tube for the outer conductor, Teflon dielectric for lower loss, and a solid copper inner conductor, usually silver-plated. This construction seems like it would be quite difficult to bend, but at the smaller diameters used in microwave work, it can be bent with care to a desired shape. The most common size is 0.141 inches in diameter, sometimes designated RG-402. Another frequently used size is 0.085 inches in diameter, favored for tighter bends, but with higher losses. Other sizes are available, both larger and smaller, but are not often seen. Impedances other than 50 Ω occasionally appear as well.

SMA connectors are a good match for semi-rigid cable. Connectors designed for the cable are readily available, affordable, and have excellent performance at microwave frequencies. Mating chassis-mount and PC-mount connectors and adaptors to other connector types are all readily available. Even better, short cable assemblies with an SMA connector on each are ubiquitous at microwave swap sessions and flea markets.

Whenever I see them for $1 or less, I buy a handful — it takes a bunch to interconnect a microwave transverter or to set up test equipment. A selection is shown in Figure 1. When shopping, look inside the connectors and choose clean ones. Some of the cables have gold, silver, or tin plating on the outside for solderability or looks, but it isn't important, since no RF travels on the outside.

Surplus cable assemblies only come in two lengths: too long and too short. And they are usually pre-bent, in the wrong place. Often, we choose to make do, adding additional bends and tolerating extra length and loss. But sometimes we don't have the space or can't tolerate the extra loss, and need to make one the right length and shape. This is how I do it.

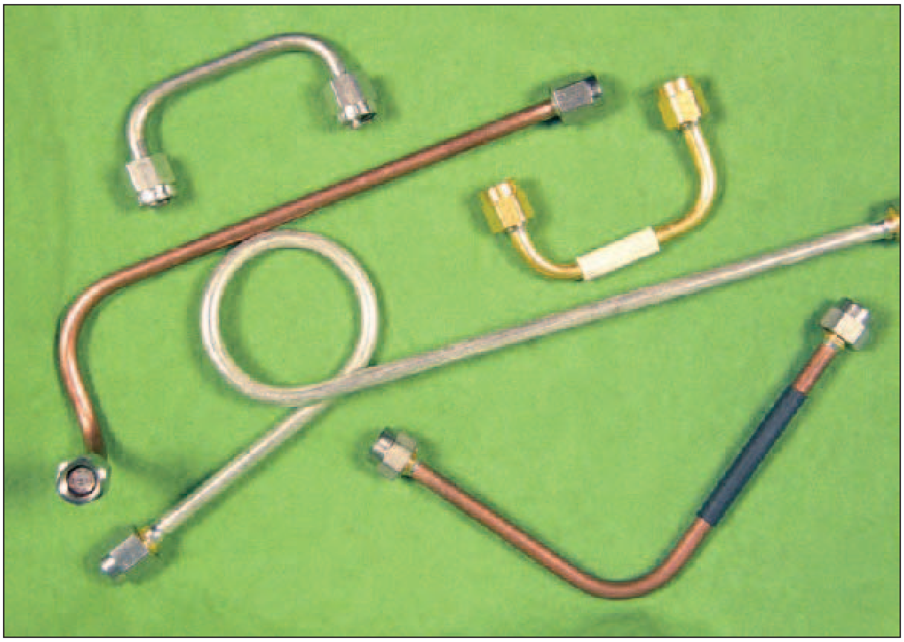


Figure 1 — Surplus semirigid cable assemblies come in all shapes and sizes.

Figure 2 — Score the outer tubing with a miniature tubing cutter.

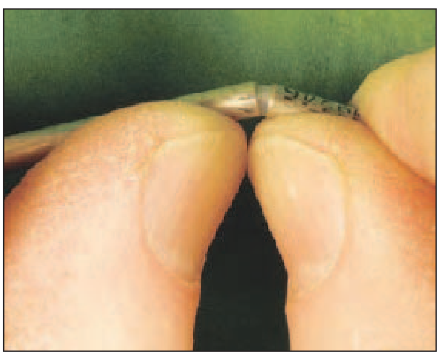


Figure 3 — Flex the cable with thumbs until the outer tubing snaps at the score line.

SMA Connector Assembly

To make things easier, I start with a cable assembly that is too long and make it into two shorter ones — that way, I only have to assemble one connector on each. For your first one, it might be a good idea to practice on the end of the less critical length first.

Some new connectors are needed — unsoldering used ones is difficult and not worth the effort. New connectors may be found surplus, but are also available from

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suppliers like Digi-Key (**www.digikey.com**) at quite reasonable prices. Either way, look at the data sheets on the Digi-Key Web site to get an idea of assembly dimensions for different styles.

Most SMA male cable connectors use the center conductor of the semirigid cable as the connector center pin, and make the end of the copper outer tube flush with the mating surface of the connector outer conductor. So the first operation is to cut the outer tube clean and square at the desired length, leaving about a ½ inch beyond the cut. I use a miniature tubing cutter to score the outer tube, as shown in Figure 2. The precision model at my local hardware store or from **www.smallparts.com** works really well — just a couple of times around scores the thin tube enough to snap the line with my thumb, as shown in Figure 3. The unwanted end of the tube can be wiggled off.

Figure 4 — Solder the connector body to the cable. Make sure to slide any necessary pieces on the cable first.

Figure 5 — Bending semirigid cable around a homemade bending jig.

Next, the Teflon dielectric is trimmed to length with a razor blade, being careful not to nick the center conductor. Some connector styles cut it flush with the outer conductor, while others leave a specified length protruding. Try and guess by the pictures on the data sheets.

The center conductor is cut to length to become the center pin. If the data sheets don't make the length clear, eyeball it from other SMA connectors. Filing a point on the end will make connections easier.

Before soldering, dry assembly is in order. Make sure everything fits together. Some styles require the threaded coupling nut to be slid on the cable before assembly, so that it can be added from the back after soldering.

When everything fits on the bench and in your mind, then put a small amount of rosin paste flux on the copper outer tube (I like Kester SP-44) and solder the connector on with a good-sized iron and a minimal amount of solder, as shown in Figure 4. After everything cools, clean up any excess flux.

The coupling nut may be loose, threaded in place, or held on with a snap ring, depending on the manufacturer. The rings just take some fiddling to get on. Once everything is in place, gently mate with a female SMA connector to check fit.

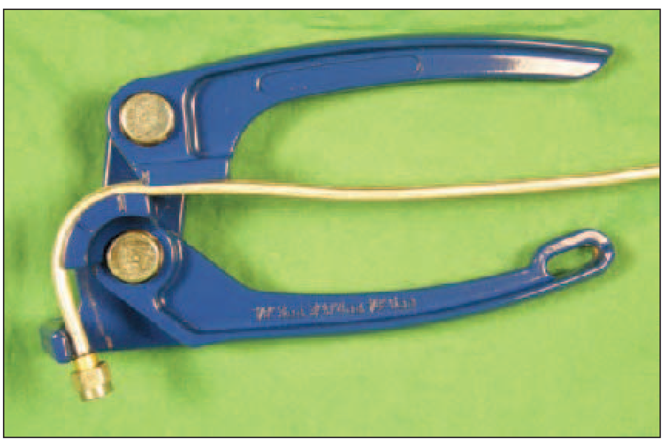


Figure 6 — Bending semirigid cable with a miniature tubing bender.

Bending Semi-Rigid Cable

Semi-rigid cable may be bent by hand, if the bend radius is fairly large. Just make a series of gentle bends with your thumbs. For tighter bends, perhaps any tighter than the size of a half-dollar, a bending tool is needed. The tool has a groove that the cable fits in to keep it round as it is bent so it isn't deformed, which would change the impedance. Professional tools are available, at a price, but aren't essential. Figure 5 shows a wooden bending jig that I made on a lathe, with a series of grooves for different bend radii, thumb operated. An inexpensive miniature tubing bender (Harbor Freight #94571; **www.harborfreight.com**) shown in Figure 6 also works, but only for one bend radius. Work slowly and don't force things.

Unbending existing bends in cable assemblies can be more difficult. Bending the cable work-hardens the copper, making it more brittle, so there is a danger of cracking the copper during additional bending. Work slowly, bending with your thumbs.

Connector Torque

In laboratories, SMA connectors are installed with a torque wrench set at 8 inch-pounds. The reason is to avoid damaging expensive precision connectors on test equipment, while making repeatable connections. Without a torque wrench, use one or two fingers on a small wrench until the connection is just snug. Before installing any SMA connector, inspect the mating surfaces to make sure they are clean, and wipe them clean with a cotton swab if needed.

However, equipment used for roving or up on a tower will soon have loose connections from vibration. I've seen many contacts missed because of loose connectors. So once you are finished testing and satisfied that everything is working, honk the connectors down! Before each roving season, I take all the covers off the equipment and make sure all connectors are tight.

Adapters

Sometimes it is necessary to connect to another type of connector, perhaps for test equipment, or because a surplus component came that way. Quality adapters from SMA connectors to other types may be found in surplus. Be wary of inexpensive imported adapters — some are made with inferior metals and dielectrics and are not suitable for microwave use.

Summary

Almost all microwave work requires some cabling — ready-to-operate transceivers aren't available. You might have noticed that all the work is done with thumbs, so even hams who are all thumbs can master SMA connectors and semirigid cable. QST

Multiband Operation with Open-Wire Line

By George Cutsogeorge, W2VJN

Despite the mechanical difficulties associated with open-wire line, there are some compelling reasons for its use, especially in simple multiband antenna systems. Every antenna system, no matter what its physical form, exhibits a definite value of impedance at the point where the transmission line is connected. Although the input impedance of an antenna system is seldom known exactly, it is often possible to make a close estimate of its value with computer modeling software.

As an example, the table lists the computed characteristics versus frequency for a multiband, 100-ft long center-fed dipole or "flat top" antenna, placed 50 ft above average ground. A non-resonant 100-ft length was chosen as an illustration of a practical size that many radio amateurs could fit into their backyards, although nothing in particular recommends this antenna over other forms. It is merely used as an example.

Modeled Data for a 100-ft Flat-Top Antenna

Freq (MHz)	*Antenna Impedance (Ω)*	*Input VSWR RG-213 Coax*	*Loss of 100 ft RG-213 Coax (dB)*	*Loss of 100 ft 450- Ω Line (dB)*	*Max Voltage RG-213 Coax at 1500 W*	*Max Voltage 450- Ω Line at 1500 W*
1.8	4.18 –*j* 1590	33.7	26.0	8.8	1507	10950
3.8	37.5 –*j* 354	16.7	5.7	0.5	1177	3231
7.1	447 +*j* 956	12.3	5.9	0.2	985	2001
10.1	2010 –*j* 2970	12.1	10.1	0.6	967	2911
14.1	87.6 –*j* 156	4.6	2.4	0.3	587	1747
18.1	1800 +*j* 1470	7.7	6.8	0.3	753	1600
21.1	461 –*j* 1250	4.6	3.2	0.1	585	828
24.9	155 +*j* 150	3.6	2.6	0.2	516	1328
28.4	2590 +*j* 772	6.7	9.4	0.5	703	1950

These values were computed using version 3 of the antenna modeling program, EZNEC (**www.eznec.com**). Antenna impedance computed using 499 segments and with the Real Ground model.

Examine the table carefully in the following discussion. Columns three and four show the SWR on a 50-Ω RG-213 coaxial transmission line directly connected to the antenna, followed by the total loss in 100 ft of this cable. The impedance for this non-resonant, 100-ft long antenna varies over a very wide range for the nine operating frequencies. The SWR on 50-Ω coax connected directly to this antenna would be extremely high on some frequencies, particularly at 1.8 MHz, where the antenna is highly capacitive because it is very much shorter than a resonant length. The loss in 100 ft of RG-213 at 1.8 MHz is a staggering 26 dB with an SWR of 33.7:1.

Contrast this to the loss in 100 ft of 450-Ω open-wire line. Here, the loss at 1.8 MHz is 8.8 dB. While 8.8 dB of loss is not particularly desirable, it is about 17 dB better than the coax! Note that the RG-213 coax exhibits a good deal of loss on almost all the bands due to mismatch. Only on 14 MHz does the loss drop down to 0.9 dB, where the antenna is just longer than 3/2-λ resonance. From 3.8 to 28.4 MHz the open-wire line has a maximum loss of only 0.6 dB.

Columns six and seven in the table list the maximum RMS voltage for 1500 W of RF power on the 50-Ω coax and on the 450-Ω open-wire line. The maximum RMS voltage for 1500 W on the open-wire line is extremely high, at 10,950 V at 1.8 MHz. The voltage for a 100-W transmitter would be reduced by a ratio of

$$\sqrt{\frac{1500}{100}} = 3.87:1$$

This is 2829 V, still high enough to cause arcing in many antenna tuners, although it only occurs at specific points that are multiples of 1/2 λ from the load. In practice, the lower voltages present along the transmission line are within the operating range of most tuners although you should remain aware that high voltages may be present along the line at some points.

The performance of our 100 foot flat top might be very poor on 160 meters for an additional reason. Not only will the radiated power will be much smaller than 1500 watts due to the high loss in the transmission line, the radiation will be maximum at 90 degrees of elevation (i.e. straight up), so this is a very poor DXing antenna.

Both of these drawbacks can be overcome very easily by changing the configuration of the system. If the two wires of the transmission line are tied together and fed against ground at the point where the line reaches ground, the antenna becomes a T which is a top-loaded vertical or monopole. This antenna has a radiation maximum at about 18 degrees and a null at 90 degrees. This is a very good DXing antenna on 160 meters. The impedance is low and easily matched with an L network or other type of antenna tuner.

Performance on 3.5 MHz as a top loaded vertical will also be good for DXing because the radiation peak is at about 19 degrees and there is a null directly overhead. Switched back to operation as a horizontal “flat top” close to resonance on 3.5 MHz, it has reasonable efficiency and makes a good high angle radiator for close in communications.

Thus, an effective arrangement can be made by switching the antenna from a flat top to a top-loaded vertical configuration. For best efficiency, a good ground system is needed. Multiple radials on the ground are effective as are a few raised radials.

On 14.1 MHz this antenna comes very close to being a fair match to RG-213 and a very effective antenna. If it is made a bit longer to 107 feet, it will be resonant on 14.1 MHz where it is 3/2 λ long and has a feed point impedance of 103 Ω, according to *EZNEC 5*. Here, the SWR will be about 1.8:1 and the loss in RG-213 will be about 0.9 dB. This is quite acceptable considering that the resulting cloverleaf radiation pattern has a bit over 8 dBi of peak gain.

In general, such a non-resonant antenna is a proven, practical multiband radiator when fed with 450-Ω open-wire ladder line connected to an antenna tuner. A longer antenna would be preferable for more efficient 160 meter operation, even with open-wire line. The tuner and the line itself must be capable of handling the high RF voltages and currents involved for high-power operation. On the other hand, if such a multiband antenna is fed directly with coaxial cable, the losses on most frequencies are prohibitive. Coax is most suitable for antennas whose resonant feed-point impedances are close to the characteristic impedance of the feed line.

ARRL Handbook CD

Template File

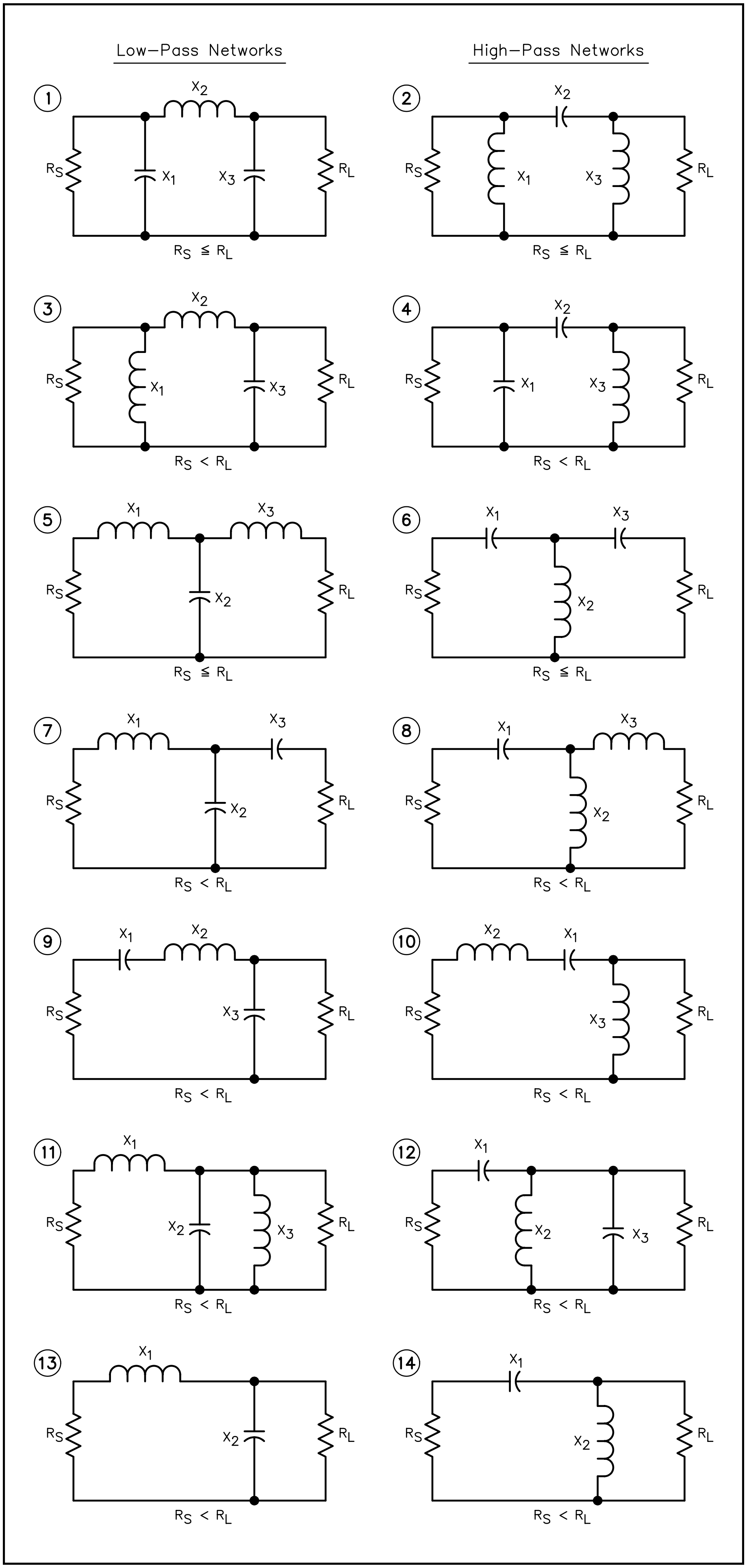
Title: NB Match

Chapter: 17

Topic: MATCH.EXE Diagrams

Template contains:

Diagram showing 14 possible network configurations to be analyzed by MATCH.EXE computer program.



This page accompanies the *ARRL Handbook* software package that is available on the *Handbook* CD. This figure shows the 14 possible network configurations used by MATCH.EXE. Refer to this figure and enter the desired network by number when you run the program. Note that low-pass networks have an odd identifying number while high-pass networks have an even identifying number. See the Broadband Transformers file on the Handbook CD-ROM.

George Cutsogeorge, W2VJN

Optimizing the Performance of Harmonic Attenuation Stubs

A quarter-wave (0.25 λ) shorted stub makes an effective high-power harmonic-reduction filter. It attenuates the harmonic by putting a very low resistance at the point of insertion in the line between an amplifier and the antenna for that particular frequency. The placement of the stub in the feed line from the transmitter output to the antenna can have a dramatic effect on stub effectiveness.

Usually we see the plots of such stubs taken with a network analyzer in a 50 Ω system. Attenuation can be 20 to 30 dB or greater for RG-213 in the HF bands. When the same stub is inserted randomly into an antenna system, however, the resulting attenuation can be higher or lower; it will depend upon the impedance at the point at which the stub is connected to the line between amplifier and antenna.

Most tube-type linear amplifiers have an LC matching network between the PA and the antenna. Solid-state amplifiers have low-pass or band-pass LC filters in the output circuit. These networks are designed to have 50 Ω output impedance at the operating frequency, but they almost always present a pure reactance at the second and higher harmonics. Some networks are capacitive, and some are inductive, ranging from a few ohms of reactance to several thousand ohms. Since the usual problem is to null a second harmonic, we will focus on that.

The circuit load impedance also can vary widely. For example, a triband beam driven on 20 meters will have a 10-meter impedance close to 50 Ω, but at the harmonics, a monoband antenna can present almost *any* impedance, depending upon matching method.

In a network analyzer the source and load impedances are both 50 Ω resistive. When checking a stub with a network analyzer, the driving impedance is the parallel equivalent of source and load. When a stub is placed in a real station somewhere between the amplifier and antenna, the impedance at that point determines how well the stub will work. When this is higher than 25 Ω at the stub null frequency, the attenuation provided will also be higher than that measured with the network analyzer. Conversely, when the system impedance is low at the connection point, the stub may only provide a few dB of attenuation — substantially less than expected.

We can measure the impedance at the stub insertion point by removing the stub and connecting a meter in its place. A simple one-port instrument such as the MFJ-259B or similar will work. If we measure something like 25 Ω or more at the null frequency, the stub will be doing a reasonable job and no further action is needed.

If we measure a low impedance, we can improve the performance of the stub by moving the connection point. Since the amplifier line and antenna line are connected at this point, we cannot tell which is causing the low impedance. By disconnecting the cable to the amplifier and the cable to the antenna, we can measure each by itself. If the impedance looking into either cable at that point in the line measures greater than 25 Ω, leave it alone. If the other presents a low impedance again at the harmonic, it can easily be increased by adding a short length of transmission line before reconnecting the T and the stub.

An example circuit and some circuit analysis can show what can happen (see Figure 1). On the left side is an analog of a power amplifier with a Pi-L network output circuit. The Pi-L is designed for 2000 Ω to 50 Ω. W1 connects the amplifier output to the T connector, which has a single stub, W3, attached. The output of the T goes through W2 to the antenna. The output is tuned to 14 MHz. The stub is a quarter-wave shorted line cut for 14 MHz. The antenna is simulated by a tuned circuit at 14 MHz in parallel with a 50 Ω load.

The impedance at the T looking toward the amplifier has maximum and minimum values that depend upon the length of W1. The same is true looking through W2 toward the antenna. For these particular values used in the simulation, the values come out to be

W1: Max = 5.5K Ω/Min = 5 Ω

W2: Max = 135 Ω/Min = 18 Ω

The circuit simulation was run under four conditions to determine the net stub attenuation:

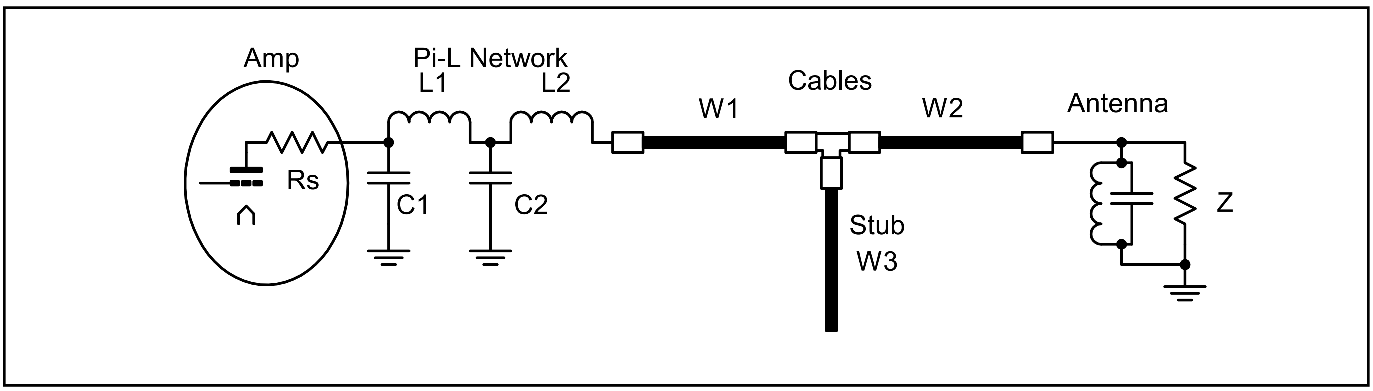


Figure 1 — An analog of a power amplifier with a Pi-L output circuit designed for 2000 Ω to 50 Ω is on the left. W1 connects the amplifier output to the T connector, with stub W3 attached. The output, tuned to 14 MHz, goes through W2 to the antenna. The stub is a quarter-wave shorted line cut to 14 MHz. A tuned circuit at 14 MHz in parallel with a 50 Ω load simulates the antenna.

W1 Max, W2 Max: Net attenuation = 51.1 dB

W1 Max, W2 Min: Net attenuation = 34.4 dB

W1 Min, W2 Max: Net attenuation = 6.4 dB

W1 Min, W2Min: Net attenuation = 6.2 dB

In order to arrive at these results, analysis under each of the four conditions was run twice — once with the stub in place, and once without the stub. The attenuation at 28 MHz with the stub was subtracted from the attenuation without the stub.

The results show a wide variation in the attenuation added by the stub. When the impedance at the stub insertion point is very low, the stub is not very effective. When it's high, the stub can be more effective than indicated by network analyzer measurements.

While we would not complain if our stub produced *more* than 50 dB second-harmonic attenuation — or even 34 dB — it would hardly be worth it to attain only 6 dB. Remember this is only an example, and results can vary a great deal from those shown. The point here is that we must make some simple measurements to be sure the stub is working well.

To prepare for the measurement we must first tune the amplifier for normal operation, and then remove the driver cable from the amplifier, so there is no possibility of RF. Next, activate the PTT to connect the amplifier tank circuit to the antenna output cable.

Now, let's say we remove the stub and measure a low impedance at the connection point. We then separate the cables from the transmitter and to the antenna. We know we will have to alter one or both of the cable lengths, and, because we are dealing with an existing installation, we will only consider adding cable, not removing it. We measure the impedance looking into the transmitter cable, and we read a low value — something lower than 25 Ω to 50 Ω at the stub null frequency. Now we increase the frequency a small amount and observe the direction of change. If the impedance goes *up* we can try adding about 0.125 λ of cable at the null frequency and then re-measure the impedance. It should be higher. If the impedance goes *down* as we sweep the frequency up a bit, we can try adding 0.375 λ of cable.

This method should be used on the antenna cable as well, if it measures a low impedance. If the station uses a pair of stubs, the same method may be used while leaving the coupling line between stubs in place. Before replacing the stub, it is a good idea to make the measurements again.

Thanks to N3RR, whose questions prompted me to look into this issue, and to NØAX for editing help.

Jim Brown, K9YC / k9yc@arrl.net

Optimizing the Location of Stubs for Harmonic Suppression

George Cutsogeorge, W2VJN, earlier this year submitted a very thought-provoking article (see "Optimizing the Performance of Harmonic Attenuation Stubs," January/February 2015 *NCJ*) on the placement of stubs for harmonic suppression in the feed lines of monoband antennas. The basic concept is solid, but I differ with him on the implementation. What we agree on is 1) a resonant antenna that presents a load in the range of 25 Ω to 100 Ω on its operating frequency will look like a very high impedance on the second harmonic; 2) the SWR at the second harmonic is typically quite high; 3) a stub will provide the greatest suppression if placed at a point along the line where the impedance is at a maximum at the frequency where suppression is desired; and 4) a stub that is poorly located along the line may provide little if any harmonic suppression.

There's another important factor: Interaction between the stub and the amplifier's output network. Modern power amplifiers generate rather strong harmonics that must be filtered by a powerful output network; the second harmonic may be only 6 dB down from the fundamental, so this network must suppress the second harmonic by at least 40 dB. The effectiveness of any passive network depends upon the load impedance. With or without the stub, the transmission line can transform the antenna's impedance at the harmonic to a value that reduces the suppression that network provides by 25 dB! When we place a stub on a line, we establish a new standing wave pattern between the stub and the amplifier. Thus, the second part of our problem is to adjust the length of the line between stub and amplifier, so that the standing wave pattern optimizes the effectiveness of the output network.

A stub works to suppress a harmonic by placing a short across the line at the harmonic frequency. Placing it at a low-impedance point provides very little suppression, because we're placing a short in parallel with a short. If the length of coax between stub and amplifier degrades the suppression that the output network was providing, the total suppression (output network + stub) may be no better than it would be without the stub. Indeed, George began pursuing this problem, because he was building stubs for stations owners who reported that they weren't working well.

Process Overview

I won't go into the details of George's suggested procedure. Read the *NCJ* article if you're interested. Here's what I recommend:

1. Measure the complex impedance of the feed line in the shack (with the antenna connected) at the frequency of the second harmonic.
2. Use N6BV's *TLW* software (comes with *The ARRL Antenna Book*), AC6LA's free *Zplots Excel* spreadsheet or his *TLDetails* windows program to find the impedance peaks on the line, or use AE6TY's free *SimSmith* software. It's a Smith Chart program that runs in Java.
3. Break the line at one of those peaks, insert a coax T, and add the stub. If you want a second stub (for greater attenuation), add it 0.25 λ closer (on the harmonic band) to the transmitter.
4. Make the line between the stub and the output of your rig or power amp a length that preserves the harmonic suppression of the rig or the power amp.

The Details

There are several methods for each step, depending on the available measurement and software tools. Our example is a 40 meter dipole with a stub to suppress the second harmonic.

Step 1: If you have only a single-frequency antenna analyzer that reads complex impedance (R ±jX), disconnect the coax from the rig, connect it to the analyzer, and measure it at the frequency of the harmonic (choose 14.175 MHz, the middle of 20 meters). It *is* important to enter the sign of the reactance. Some analyzers don't read the sign, providing a procedure for finding it, but AC6LA[1] and VK1OD[2] have observed that it can give an erroneous result when measuring the impedance of a transmission line feeding an antenna. If you have a vector impedance analyzer or vector network analyzer, measure the impedance over the limits of the harmonic band (20 meters in our example).

For a simple antenna such as a dipole, you can model it using *NEC* and compute the SWR at the harmonic. Put the cursor at the center of the band and read the impedance. To make use of this data, you must know the length of the line. For the measurement method above, you do *not* need to know the length of the line, although if you do, and if you know line loss, you can compute suppression all the way to the antenna.

Step 2 — Using *TLW*: Choose the type of coax you are using (if multiple types are in the run to the antenna, choose the type of the section of the line that you measured). Enter the measured impedance value, checking the *Input* box (because you measured at the input end of the line). Enter the measurement frequency (14.175 MHz), and the length of the line of the line if you know it. If you don't, use any length greater than 1 λ. (If your impedance is from *NEC*, check the *Load* box, and you *will* need to know the length of the line.) Now, go to the Graph section, select Voltage/Current, and click on Graph. This gives us a graph of voltage and current peaks. Place your choke at the location of the most convenient voltage peak. *Caution! Don't use the* NEC *approach with* TLW*, unless you know the length of the line to good accuracy, and you are using the same type of coax for the entire run.*

If you can't put the stub within the length of the existing line (the voltage peaks may be at inaccessible locations), then note the location of the voltage peak closest to the transmitter and the spacing (feet and inches) between peaks (0.5 λ), add enough coax to extend the line to the next peak, and put the stub there. The added coax should have the same impedance but can have a different velocity factor (VF). If the VF is different, change to the new coax type in *TLW*, select the *Load* box, Click on *Graph*, and read the distance to the next voltage peak. (This last step is important; *TLW* won't redraw the graph automatically, so without clicking on *Graph* you'll be looking at the earlier version of the problem.)

Step 2 — Using *TLDetails*: Download the software from **http://ac6la.com**. Set the line type, enter the frequency, enter the measured R and (signed) X at the station end of the line, choose "At Input". Set the initial line length to 0. Use the length spin button to increase the length until the blue dot (the "At Load" marker) is at the right side of the Smith chart. That's the length toward the antenna from the station end of the line where the first stub should be placed. Follow the same procedure as with *TLW*, if you want to add coax before the stub.

Step 2 — Using *Zplots*: Download the software from **http://ac6la.com** and load it into *Excel* (it will not run in other spreadsheet programs), see the instructions at **http://ac6la.com/zplots1.html#GenerateFF** to *Generate Data for a Transmission Line at fixed frequency*.

Step 2 — Using *SimSmith*: If your measurement is a single data point, enter it in the *Load* block. If you have made a sweep measurement, export the data from your measuring instrument in Touchstone format (it's a plain text file with a .s1p extension), and import it into *SimSmith* as a "Load File." (*SimSmith* author AE6TY advises that the only form of .s1p file that older versions of *SimSmith* imports is the "S" format. Version 11.5 and later will also accept "Z" format.)

Add a transmission line to the model, and choose the coax type you're using. Looking at the Smith Chart display, vary the length of the transmission line you just added until the antenna (or the single data point) is centered along the horizontal line at the right side of the chart. This is a high-impedance point on the line; it repeats every 0.5 λ. Place your stub at one of them. A negative length in *SimSmith* moves along the line toward the antenna (you must break the feed line at that point and insert the stub), a positive number moves away from the antenna (add coax to the line, and put the stub there).

As with *TLW*, if you add coax to the system, you must add coax of the same impedance, but it can have a different VF. If it does, change the coax in the model to match what you're using.

Step 3 — Building and installing the stub: Install a connector on one end of a piece of coax, cutting the coax about 10 percent longer than the computed value. Strip one-half inch or so at the far end, and short the shield to the center conductor. Connect the stub to an impedance-measuring device and vary the frequency until you see an impedance near 0 Ω with X also near 0 Ω. It should be below your desired frequency by about 10 percent. Use the actual percentage below your target frequency to tell you how much to cut, and cut about one-half as much. Repeat until the stub is at the desired frequency. Carefully solder the shorted end and weatherproof it. Add a coax T to the line where the stub will go. Connect the antenna side of the feed line to one side of the T. Using a coax barrel, connect the transmitter side to the male connector, and connect the stub to the other side of the T.

In general, it's best to add the stub in the *existing* length of line without making the line longer; the longer line will increase the loss slightly. Stubs are also effective at peaks near the antenna, so if that is convenient, consider it. You may also want to add a second stub (see below), which is another good reason for putting the first stub closer to the antenna.

Step 3a — Placing a Second Stub: Building the stub from high-quality coax, such as Belden 8237, 8238, 8213, or Times LMR400, should provide about 30 dB of suppression at the second harmonic. Another 30 dB of attenuation is possible by adding a second stub. The first stub disturbs the line by placing a short across it on 20 meters, so it changes the 20 meter standing wave pattern between it and the transmitter. Because impedance minima and maxima are 90° apart, the next maximum will be 90° closer to the transmitter. Consider this when deciding where to put the first stub; depending on how much of the line you have access to, you may be able to position *both* stubs without making the line longer.

There are, of course, alternative methods for measuring feed-line impedance and finding voltage maxima. Use the methods that work for you and with the resources available to you. *SimSmith* has the added advantage of allowing you to compute the attenuation of the stub over the range of frequencies where you measured Z. It can also compute the attenuation of a second stub. To do that, you must know the feed-line length and its attenuation, subtract it from your measured data (negative feed line length), then add back the same feed line to the stub (and the second stub, if used).

Sometimes it isn't practical to find the high-voltage point. The effectiveness of the stub nearest the antenna will depend on luck — anything from a few decibels to as much as 36 dB (because we haven't found the null). But the second stub *will* be good for 30 dB, because we *know* it's 90° from a short (the first stub).

Make sure you know your coax. Sometimes we splice runs of coax together to reach the antenna, and they may not all be the same type. We might use hard line for the long run to a tower, with RG-8 from the end of the hard line to the shack. You need to know the type(s) of coax only within the section of the run between where you take your measurement and where you place the stub, and between the stub and the amplifier. For good accuracy, you need good data on the coax for that part of the run. Other coax in the run to the antenna contributes to establishing the locations of the nulls, but you don't need to know about it except to compute loss (including attenuation at the harmonic).

Step 4 Discussion — Amplifier Output Networks: In vacuum tube amplifiers, these networks function both as very good low pass filters and as impedance-matching networks. Two common filter configurations are in common use. Looking at the schematic starting from the tube or transistors, the pi network has a shunt capacitor, followed by a series inductor, and another shunt capacitor. The second common network in vacuum tube amplifiers is the pi-L, which starts with a pi section followed by a second series inductor. Solid state amps, which use output devices having relatively low output impedance, often utilize multi-stage elliptical filters, with shunt C as the last element, and may do little or no impedance matching. Pi networks and elliptical filters provide the greatest attenuation into loads of 50 Ω or greater; pi-L networks work best into loads of 50 Ω or less. Since our stub causes line impedance (at the second harmonic) to vary from a near short to a near open every 90°, it's possible to give the amp its optimum load simply by using the right line length between the stub and the amplifier. "Some Q&A About Coax and Stubs for Your HF Station" at **http://k9yc.com/Coax-Stubs.pdf** includes a table listing the output network configurations of many popular power amplifiers.

Step 4 — Optimizing Line Length Between Amp Output and Stub: You don't need to be very precise about the length of this line; getting within about 35° of optimum loses only about 1 dB of harmonic suppression, 45° only about 3 dB. If we knew the actual values of L and C in the amplifier's output network, it's possible to model it in *TLW* or *SimSmith* and get very precise, but that information is rarely available. You can get reasonably close (within that ±45° range of ideal) by using this rule.

♦ If the last element, nearest the output terminal, is a series inductor, make the coax between the amp and the stub as short as possible or some even multiple of 0.5 λ at the harmonic.

♦ If the last output component is a capacitor, make the coax to the stub some odd multiple of 0.25 λ at the harmonic. If using *two* stubs, the coax length here is that from the power amp to the first stub it sees.

Note that this rule does not yield optimum results for all cases, because that output network may be adjustable to match your particular antenna at the operating frequency. But it does get you in the ballpark for most reasonably well-matched antennas, and that's likely to be close enough that you won't lose much of the harmonic suppression of the output stage.

As an experiment, I inserted a voltage probe at the output of my Ten-Tec Titan (pi-L) and Elecraft KPA500 (elliptical) and measured 2nd harmonic suppression as I varied the length line to the first stub, and compared it to a 50 Ω dummy load. For both amplifiers, following the rule of thumb resulted in suppression that was at least as good as the dummy load, and both worst-case and best-case suppression occurred from 10° to 30° one side or the other from the rule of thumb.

Going Beyond the Second Harmonic

If a stub is intended to suppress the *third* harmonic rather than the second, design and measurement should be at that third harmonic frequency. Likewise, if the stub is intended to suppress the fourth harmonic as well as the second, measure and attempt to optimize at *both* harmonics.

How about optimizing for more than one harmonic? Consider, for example, the stub for our 40 meter antenna. Although 20 meters is usually the most critical, we usually would like it also to suppress the 4th harmonic on 10 meters. The impedance at the antenna is likely to be fairly high on 10 meters, but the phase, and, thus, the posi-

Modeled Results For Two-Stub Filters

How much harmonic suppression can we get from a well-placed pair of stubs? Figures 1 and 2 were modeled in *SimSmith* for a 40 meter dipole with stubs made from RG-8X and stubs made with RG-8, respectively. The upper curve is the attenuation between the amplifier and the nearest stub, the next curve is the total attenuation to the second stub, and the lower curve is the total attenuation to the antenna. These stubs are tuned for maximum attenuation in the CW band segment, because most 40 meter phone segment harmonics fall outside the 20 meter phone band.

Note that the RG-8X stubs offer less attenuation but flatter response. This is because the smaller-diameter coax has more resistance, is a less-ideal short circuit, and thus has a lower Q. The model assumes that the spacing between the amplifier and the nearest stub has been adjusted so that it does not degrade harmonic suppression in the output stage. The line is 100 feet long, and the loss in the line at the second harmonic without the stubs is 9.1 dB — because of the severe mismatch at the harmonic.

The lower graphs make the same comparison for an antenna that is resonant on both 40 and 20 meters. The stubs provide a bit less attenuation — the line is matched at the second harmonic, so there is no impedance peak — and because the loss in the line at the second harmonic without the stubs is only 0.65 dB (no loss to mismatch).

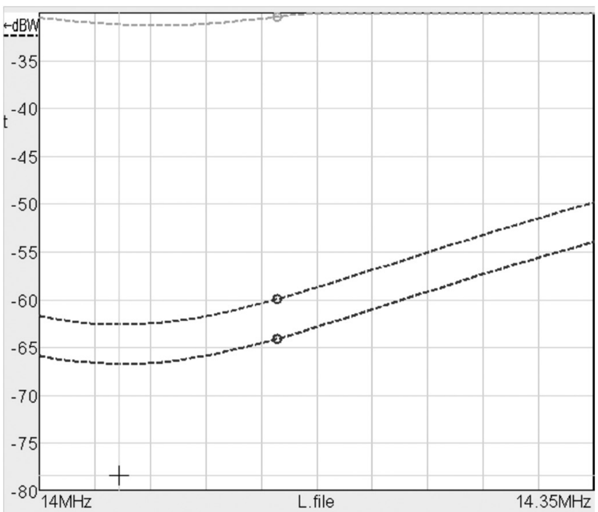


Figure 1 — Two RG-8X stubs on a 40 meter dipole.

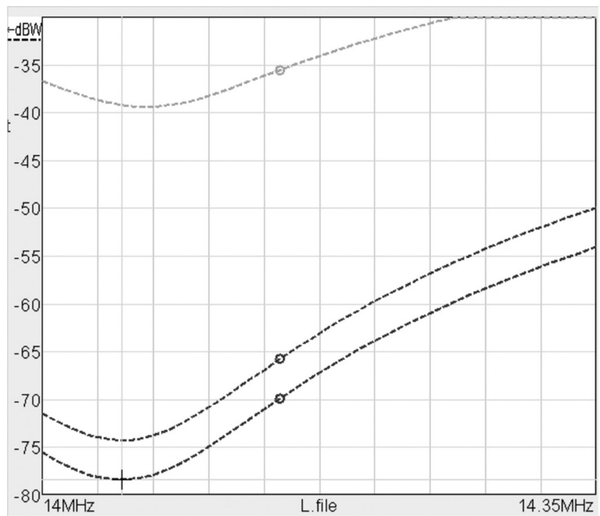


Figure 2 — Two RG-8 stubs on a 40 meter dipole.

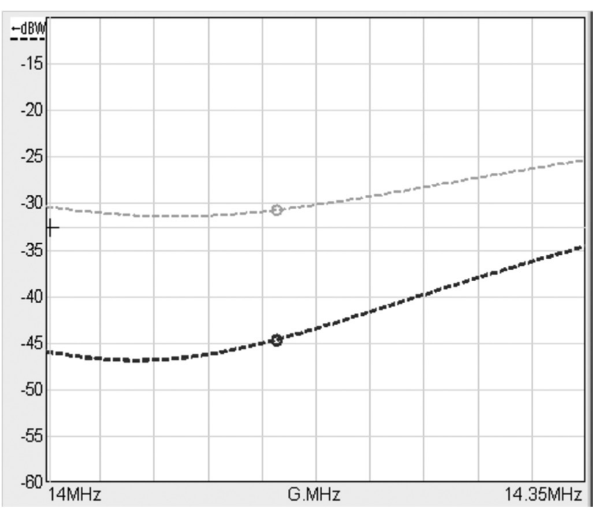


Figure 3 — Two RG-8X stubs when 20 meters is 50 Ω.

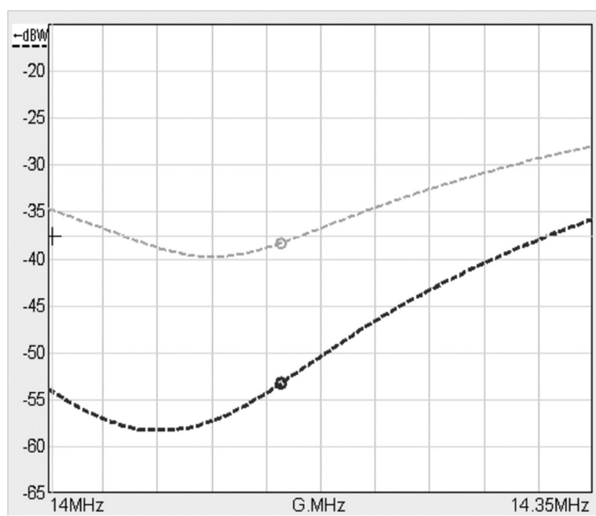


Figure 4 — Two RG-8 stubs when 20 meters is 50 Ω.

tion along the line of the impedance peaks may be different, and the VF will be slightly higher. With optimal placement of the stub for 20 meters, there is a good chance of being in the ballpark, but not necessarily optimum, on 10 meters. The good news is that you'll usually need a lot less suppression on higher-order harmonics.

What if we're using two stubs to increase suppression on 20 meters? We placed these 90° apart on 20 meters, which makes them 180° apart on 10 meters, and makes the second stub relatively ineffective. Shifting its position by only 20° on 20 meters would sacrifice a few decibels of attenuation on 20 meters while greatly improving attenuation on 10 meters. Whether to do this depends entirely on how much attenuation is needed on the two bands in question.

We still must consider the length of line between the stub and the power amp. What worked on 20 meters could turn out to be poor for 10 meters. The same tactic of adding or subtracting 20° or so of line on 20 meters would sacrifice a few decibels of attenuation on 20 meters but significantly improve attenuation on 10 meters.

Optimizing the Position of a Receive Stub

In receive mode, the antenna is the source, and the receiver input is the load. Receive stubs typically operate at the fundamental frequency of the harmonically related band below where we're operating. These are designed to prevent that transmitter from overloading the receiver. Just as the output impedance of an amplifier will vary with frequency, so can the input impedance of a receiver's input stage. To place this stub, disconnect the antenna from the receiver and measure the impedance of the receiver input at that lower frequency. Sticking with our 40 meter dipole example, the process becomes:

1. Measure the receiver input impedance, either at the receiver input, or at the end of a length of coax connected to it.

2. Enter the measured impedance in *TLW, TLD, ZPlots*, or *SimSmith* as the *Load*, make the line length at least 0.5 λ on the lower band (80 meters in our example), and *Graph* the Voltage/Current standing waves. In *SimSmith*, coax going toward the antenna should be entered as negative length, because the antenna is now the *Source*, and the receiver is the *Load*.

3. As before, find a convenient voltage maximum in *TLW, TLD, or ZPlots*, or a point near the right end of the horizontal line through the center of the Smith Chart in *SimSmith*. This optimizes the stub location with respect to the receiver input. When placing a receiving stub, be careful not to change the position of any stubs already inserted for harmonic suppression.

Acknowledgements

This article benefited from extensive online discussions among NØAX, AE6TY, N6BV, AC6LA, W2VJN, and the author.

Notes

[1]**www.ac6la.com/zpapndx3.html**
[2]**http://owenduffy.net/blog/?p-2436**

The following material was extracted from earlier editions. Figure and Equation sequence references are from the 21st edition of *The ARRL Antenna Book*

Smith Chart Calculations

The Smith Chart is a sophisticated graphic tool for solving transmission line problems. One of the simpler applications is to determine the feed-point impedance of an antenna, based on an impedance measurement at the input of a random length of transmission line. By using the Smith Chart, the impedance measurement can be made with the antenna in place atop a tower or mast, and there is no need to cut the line to an exact multiple of half wavelengths. The Smith Chart may be used for other purposes, too, such as the design of impedance-matching networks. These matching networks can take on any of several forms, such as L and pi networks, a stub matching system, a series-section match, and more. With a knowledge of the Smith Chart, the amateur can eliminate much "cut and try" work.

Named after its inventor, Phillip H. Smith, the Smith Chart was originally described in *Electronics* for January 1939. Smith Charts may be obtained at most university book stores. Smith Charts are also available from ARRL HQ. (See the caption for Fig 3.)

The input impedance, or the impedance seen when "looking into" a length of line, is dependent upon the SWR, the length of the line, and the Z_0 of the line. The SWR, in turn, is dependent upon the load which terminates the line. There are complex mathematical relationships which may be used to calculate the various values of impedances, voltages, currents, and SWR values that exist in the operation of a particular transmission line. These equations can be solved with a personal computer and suitable software, or the parameters may be determined with the Smith Chart. Even if a computer is used, a fundamental knowledge of the Smith Chart will promote a better understanding of the problem being solved. And such an understanding might lead to a quicker or simpler solution than otherwise. If the terminating impedance is known, it is a simple matter to determine the input impedance of the line for any length by means of the chart. Conversely, as indicated above, with a given line length and a known (or measured) input impedance, the load impedance may be determined by means of the chart—a convenient method of remotely determining an antenna impedance, for example.

Although its appearance may at first seem somewhat formidable, the Smith Chart is really nothing more than a specialized type of graph. Consider it as having curved, rather than rectangular, coordinate lines. The coordinate system consists simply of two families of circles—the resistance family, and the reactance family. The resistance circles, **Fig 1**, are centered on the resistance axis (the only straight line on the chart), and are tangent to the outer circle at the right of the chart. Each circle is assigned a value of resistance, which is indicated at the point where the circle crosses the resistance axis. All points along any one circle have the same resistance value.

The values assigned to these circles vary from zero at the left of the chart to infinity at the right, and actually represent a *ratio* with respect to the impedance value assigned to the center point of the chart, indicated 1.0. This center point is called prime center. If prime center is assigned a value of 100 Ω, then 200 Ω resistance is represented by the 2.0 circle, 50 Ω by the 0.5 circle, 20 Ω by the 0.2 circle, and so on. If, instead, a value of 50 is assigned to prime center, the 2.0 circle now represents 100 Ω, the 0.5 circle 25 Ω, and the 0.2 circle 10 Ω. In each case, it may be seen that the value on the chart is determined by dividing the actual resistance by the number

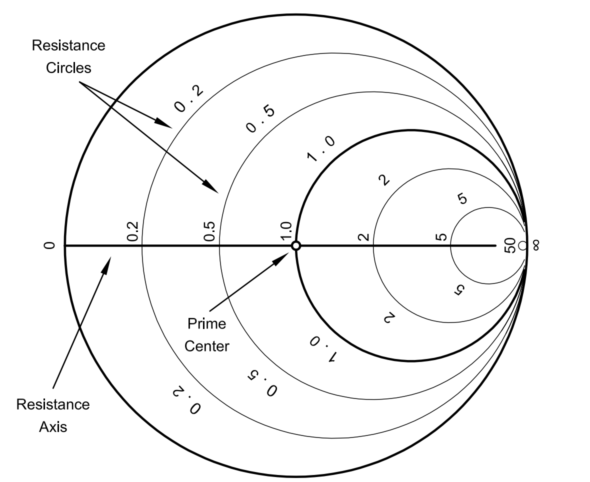


Fig 1—Resistance circles of the Smith Chart coordinate system.

assigned to prime center. This process is called normalizing.

Conversely, values from the chart are converted back to actual resistance values by multiplying the chart value times the value assigned to prime center. This feature permits the use of the Smith Chart for any impedance values, and therefore with any type of uniform transmission line, whatever its impedance may be. As mentioned above, specialized versions of the Smith Chart may be obtained with a value of 50 Ω at prime center. These are intended for use with 50-Ω lines.

Now consider the reactance circles, **Fig 2**, which appear as curved lines on the chart because only segments of the complete circles are drawn. These circles are tangent to the resistance axis, which itself is a member of the reactance family (with a radius of infinity). The centers are displaced to the top or bottom on a line tangent to the right of the chart. The large outer circle bounding the coordinate portion of the chart is the reactance axis.

Each reactance circle segment is assigned a value of reactance, indicated near the point where the circle touches the reactance axis. All points along any one segment have the same reactance value. As with the resistance circles, the values assigned to each reactance circle are normalized with respect to the value assigned to prime center. Values to the top of the resistance axis are positive (inductive), and those to the bottom of the resistance axis are negative (capacitive).

When the resistance family and the reactance family of circles are combined, the coordinate system of the Smith Chart results, as shown in **Fig 3**. Complex impedances (R + *j*X) can be plotted on this coordinate system.

IMPEDANCE PLOTTING

Suppose we have an impedance consisting of 50 Ω resistance and 100 Ω inductive reactance (Z = 50 + *j* 100). If we assign a value of 100 Ω to prime center, we normalize the above impedance by dividing each component of the impedance by 100. The normalized impedance is then 50/100 + *j* (100/100) = 0.5 + *j* 1.0. This impedance is plotted on the Smith Chart at the intersection of the 0.5 resistance circle and the +1.0 reactance circle, as indicated in Fig 3. Calculations may now be made from this plotted value.

Now say that instead of assigning 100 Ω to prime center, we assign a value of 50 Ω. With this assignment, the 50 + *j* 100 Ω impedance is plotted at the intersection of the 50/50 = 1.0 resistance circle, and the 100/50 = 2.0 positive reactance circle. This value, 1 + *j* 2, is also indicated in Fig 3. But now we have *two* points plotted in Fig 3 to represent the same impedance value, 50 + *j* 100 Ω. How can this be?

These examples show that the same impedance may be plotted at different points on the chart, depending upon the value assigned to prime center. But two plotted points cannot represent the same impedance *at the same time*! It is customary when solving transmission-line problems to assign to prime center a value equal to the characteristic impedance, or Z_0, of the line being used. This value should always be recorded at the start of calculations, to avoid possible confusion later. (In using the specialized charts with the value of 50 at prime center, it is, of course, not necessary to normalize impedances when working with 50-Ω line. The resistance and reactance values may be read directly from the chart coordinate system.)

Prime center is a point of special significance. As just mentioned, is is customary when solving problems to assign the Z_0 value of the line to this point on the chart—50 Ω for a 50-Ω line, for example. What this means is that the center point of the chart now represents 50 +*j* 0 ohms–a pure resistance equal to the characteristic impedance of the line. If this were a load on the line, we recognize from transmission-line theory that it represents a perfect match, with no reflected

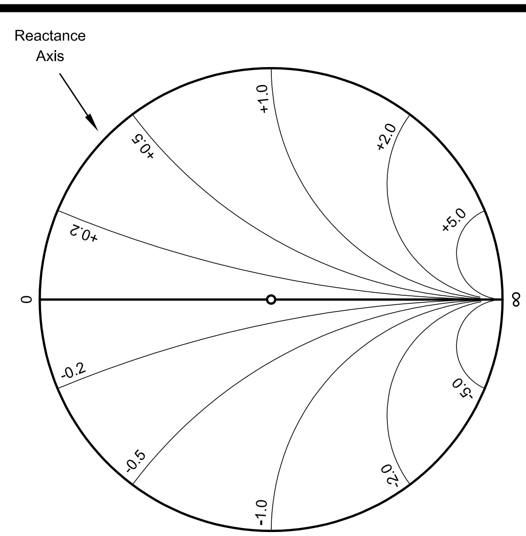


Fig 2—Reactance circles (segments) of the Smith Chart coordinate system.

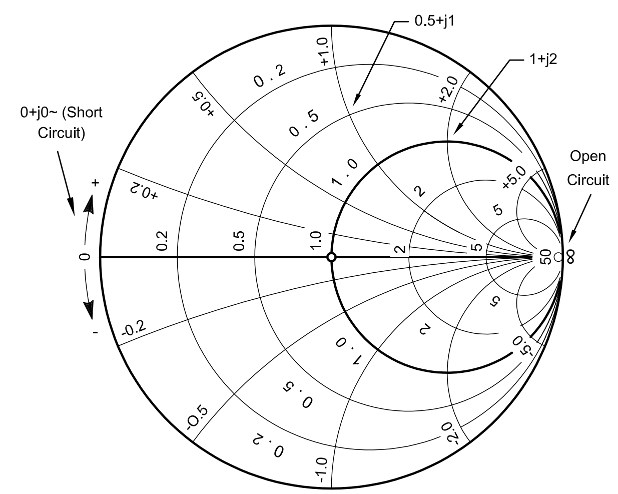


Fig 3—The complete coordinate system of the Smith Chart. For simplicity, only a few divisions are shown for the resistance and reactance values. Various types of Smith Chart forms are available from ARRL HQ. At the time of this writing, five 8 1/2 × 11 inch Smith Chart forms are available for $2.

power and with a 1.0 to 1 SWR. Thus, prime center also represents the 1.0 SWR circle (with a radius of zero). SWR circles are also discussed in a later section.

Short and Open Circuits

On the subject of plotting impedances, two special cases deserve consideration. These are short circuits and open circuits. A true short circuit has zero resistance and zero reactance, or $0 + j\,0$). This impedance is plotted at the left of the chart, at the intersection of the resistance and the reactance axes. By contrast, an open circuit has infinite resistance, and therefore is plotted at the right of the chart, at the intersection of the resistance and reactance axes. These two special cases are sometimes used in matching stubs, described later.

Standing-Wave-Ratio Circles

Members of a third family of circles, which are not printed on the chart but which are added during the process of solving problems, are standing-wave-ratio or SWR circles. See **Fig 4**. This family is centered on prime center, and appears as concentric circles inside the reactance axis. During calculations, one or more of these circles may be added with a drawing compass. Each circle represents a value of SWR, with every point on a given circle representing the same SWR. The SWR value for a given circle may be determined directly from the chart coordinate system, by reading the resistance value where the SWR circle crosses the resistance axis to the right of prime center. (The reading where the circle crosses the resistance axis to the left of prime center indicates the inverse ratio.)

Consider the situation where a load mismatch in a length of line causes a 3-to-1 SWR ratio to exist. If we temporarily disregard line losses, we may state that the SWR remains constant throughout the entire length of this line. This is represented on the Smith Chart by drawing a 3:1 constant SWR circle (a circle with a radius of 3 on the resistance axis), as in **Fig 5**. The design of the chart is such that any impedance encountered *anywhere* along the length of this mismatched line will fall on the SWR circle. The impedances may be read from the coordinate system merely by the progressing around the SWR circle by an amount corresponding to the length of the line involved.

This brings into use the wavelength scales, which appear in Fig 5 near the perimeter of the Smith Chart. These scales are calibrated in terms of portions of an electrical wavelength along a transmission line. Both scales start from 0 at the left of the chart. One scale, running counterclockwise, starts at the generator or input end of the line and progresses toward the load. The other scale starts at the load and proceeds toward the generator in a clockwise direction. The complete circle around the edge of the chart represents $^1/_2\ \lambda$. Progressing once around the perimeter of these scales corresponds to progressing along a transmission line for $^1/_2\ \lambda$. Because impedances repeat themselves every $^1/_2\ \lambda$ along a piece of line, the chart may be used for any length of line by disregarding or subtracting from the line's total length an integral, or whole number, of half wavelengths.

Also shown in Fig 5 is a means of transferring the

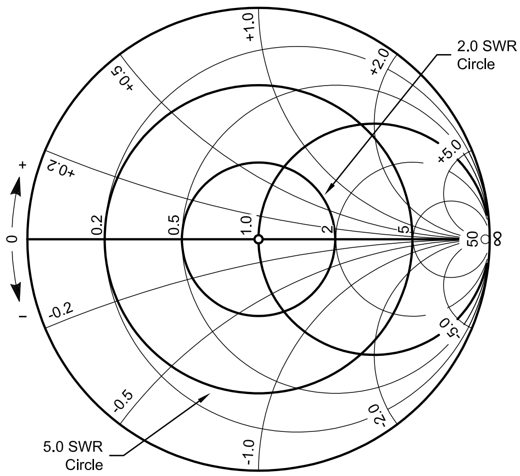


Fig 4—Smith Chart with SWR circles added.

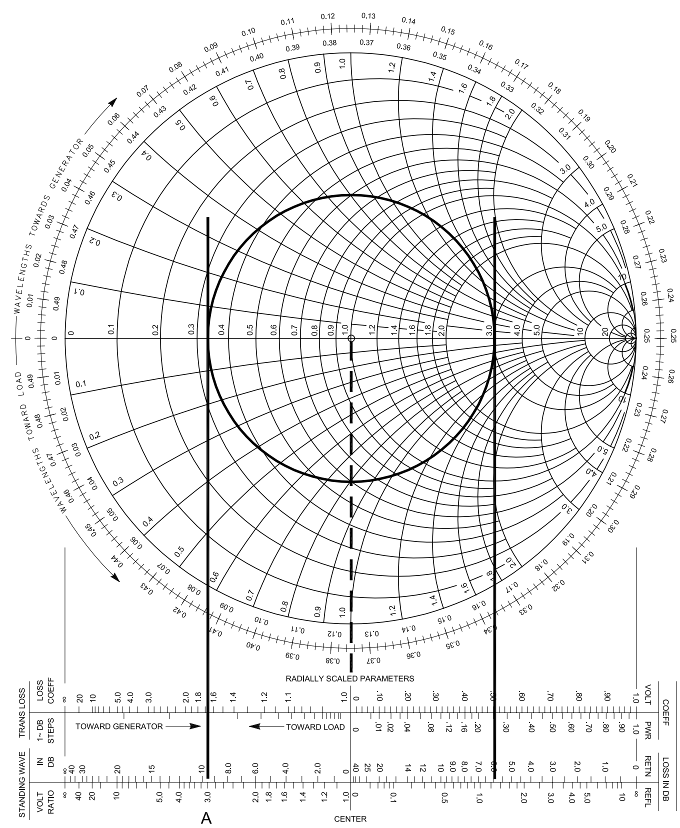


Fig 5—Example discussed in text.

radius of the SWR circle to the external scales of the chart, by drawing lines tangent to the circle. Another simple way to obtain information from these external scales is to transfer the radius of the SWR circle to the external scale with a drawing compass. Place the point of a drawing compass at the center or 0 line, and inscribe a short arc across the appropriate scale. It will be noted that when this is done in Fig 5, the external STANDING-WAVE VOLTAGE-RATIO scale indicates the SWR to be 3.0 (at A)—our condition for initially drawing the circle on the chart (and the same as the SWR reading on the resistance axis).

SOLVING PROBLEMS WITH THE SMITH CHART

Suppose we have a transmission line with a characteristic impedance of 50 Ω and an electrical length of 0.3 λ. Also, suppose we terminate this line with an impedance having a resistive component of 25 Ω and an inductive reactance of 25 Ω (Z = 25 + *j* 25). What is the input impedance to the line?

The characteristic impedance of the line is 50 Ω, so we begin by assigning this value to prime center. Because the line is not terminated in its characteristic impedance, we know that standing waves will exist on the line, and that, therefore, the input impedance to the line will not be exactly 50 Ω. We proceed as follows. First, normalize the load impedance by dividing both the resistive and reactive components by 50 (Z_0 of the line being used). The normalized impedance in this case is 0.5 + *j* 0.5. This is plotted on the chart at the intersection of the 0.5 resistance and the +0.5 reactance circles, as in **Fig 6**. Then draw a constant SWR circle passing through this point. Transfer the radius of this circle to the external scales with the drawing compass. From the external STANDING-WAVE VOLTAGE-RATIO scale, it may be seen (at A) that the voltage ratio of 2.62 exists for this radius, indicating that our line is operating with an SWR of 2.62 to 1. This figure is converted to decibels in the adjacent scale, where 8.4 dB may be read (at B), indicating that the ratio of the voltage maximum to the voltage minimum along the line is 8.4 dB. (This is mathematically equivalent to 20 times the log of the SWR value.)

Next, with a straightedge, draw a radial line from prime center through the plotted point to intersect the wavelengths scale. At this intersection, point C in Fig 6, read a value from the wavelengths scale. Because we are starting from the load, we use the TOWARD GENERATOR or outermost calibration, and read 0.088 λ.

To obtain the line input impedance, we merely find the point on the SWR circle that is 0.3 λ toward the generator from the plotted load impedance. This is accomplished by adding 0.3 (the length of the line in wavelengths) to the reference or starting point, 0.088; 0.3 + 0.088 = 0.388. Locate 0.388 on the TOWARD GENERATOR scale (at D). Draw a second radial line from this point to prime center. The intersection of the new radial line with the SWR circle represents the normalized line input impedance, in this case 0.6 – *j* 0.66.

To find the unnormalized line impedance, multiply by 50, the value assigned to prime center. The resulting value is 30 – *j* 33, or 30 Ω resistance and 33 Ω capacitive reactance. *This* is the impedance that a transmitter must match if such a system were a combination of antenna and transmission line. This is also the impedance that would be measured on an impedance bridge if the measurement were taken at the line input.

In addition to the line input impedance and the SWR, the chart reveals several other operating characteristics of the above system of line and load, if a closer look is desired. For example, the voltage reflection coefficient, both magnitude and phase angle, for this particular load is given. The phase angle is read under the radial line drawn through the plot of the load impedance, where the line intersects the ANGLE OF REFLECTION COEFFICIENT scale. This scale is not included in Fig 6, but will be found on the Smith Chart just inside the wavelengths scales. In this example, the reading is 116.6 degrees. This indicates the angle by which the reflected voltage wave leads the incident wave at the load. It will be noted that angles on the bottom half, or capacitive-reactance half,

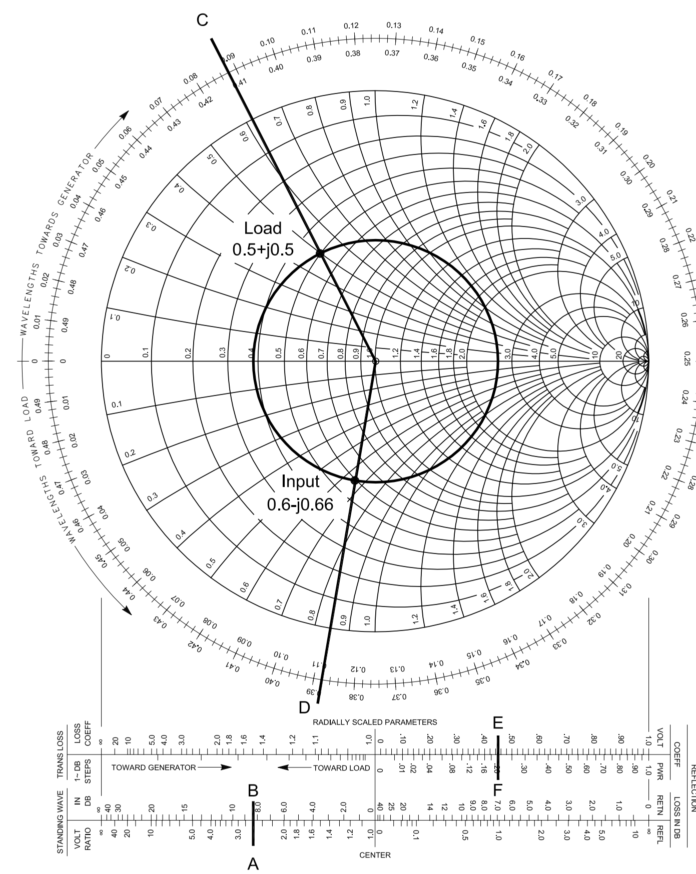


Fig 6—Example discussed in text.

of the chart are negative angles, a "negative" lead indicating that the reflected voltage wave actually lags the incident wave.

The magnitude of the voltage-reflection-coefficient may be read from the external REFLECTION COEFFICIENT VOLTAGE scale, and is seen to be approximately 0.45 (at E) for this example. This means that 45 percent of the incident voltage is reflected. Adjacent to this scale on the POWER calibration, it is noted (at F) that the power reflection coefficient is 0.20, indicating that 20 percent of the incident power is reflected. (The amount of reflected power is proportional to the square of the reflected voltage.)

ADMITTANCE COORDINATES

Quite often it is desirable to convert impedance information to admittance data—conductance and susceptance. Working with admittances greatly simplifies determining the resultant when two complex impedances are connected in parallel, as in stub matching. The conductance values may be added directly, as may be the susceptance values, to arrive at the overall admittance for the parallel combination. This admittance may then be converted back to impedance data, if desired.

On the Smith Chart, the necessary conversion may be made very simply. The equivalent admittance of a plotted impedance value lies diametrically opposite the impedance point on the chart. In other words, an impedance plot and its corresponding admittance plot will lie on a straight line that passes through prime center, and each point will be the same distance from prime center (on the same SWR circle). In the above example, where the normalized line input impedance is 0.6 – *j* 0.66, the equivalent admittance lies at the intersection of the SWR circle and the extension of the straight line passing from point D though prime center. Although not shown in Fig 6, the normalized admittance value may be read as 0.76 + *j* 0.84 if the line starting at D is extended.

In making impedance-admittance conversions, remember that capacitance is considered to be a positive susceptance and inductance a negative susceptance. This corresponds to the scale identification printed on the chart. The admittance in siemens is determined by dividing the normalized values by the Z_0 of the line. For this example the admittance is 0.76/50 + *j* 0.84/50 = 0.0152 + *j* 0.0168 siemen. Of course admittance coordinates may be converted to impedance coordinates just as easily—by locating the point on the Smith Chart that is diametrically opposite that representing the admittance coordinates, on the same SWR circle.

DETERMINING ANTENNA IMPEDANCES

To determine an antenna impedance from the Smith Chart, the procedure is similar to the previous example. The electrical length of the feed line must be known and the impedance value at the input end of the line must be determined through measurement, such as with an impedance-measuring or a good quality noise bridge. In this case, the antenna is connected to the far end of the line and becomes the load for the line. Whether the antenna is intended purely for transmission of energy, or purely for reception makes no difference; the antenna is still the terminating or load impedance on the line as far as these measurements are concerned. The input or generator end of the line is that end connected to the device for measurement of the impedance. In this type of problem, the measured impedance is plotted on the chart, and the TOWARD LOAD wavelengths scale is used in conjunction with the electrical line length to determine the actual antenna impedance.

For example, assume we have a measured input impedance to a 50-Ω line of 70 – *j* 25 Ω. The line is 2.35 λ long, and is terminated in an antenna. What is the antenna feed impedance? Normalize the input impedance with respect to 50 Ω, which comes out 1.4 – *j* 0.5, and plot this value on the chart. See **Fig 7**. Draw a constant SWR circle through the point, and transfer the radius to the external scales. The SWR of 1.7 may be read from the VOLTAGE RATIO scale (at A). Now draw a radial line from prime center through this plotted point to the wavelengths scale, and read a reference value (at B). For this

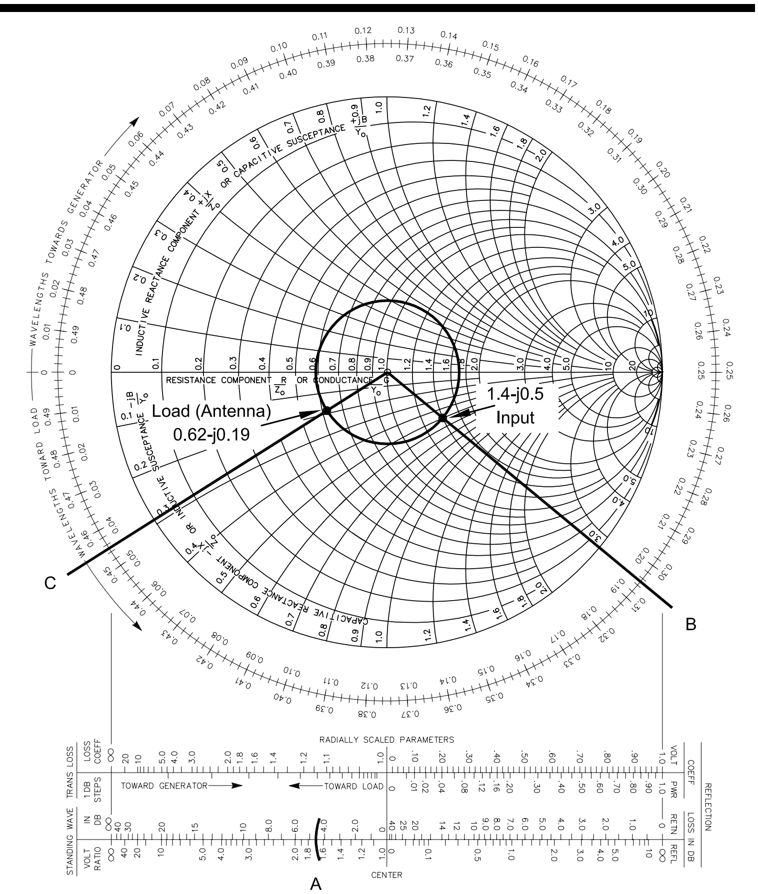


Fig 7—Example discussed in text.

case the value is 0.195, on the TOWARD LOAD scale. Remember, we are starting at the generator end of the transmission line.

To locate the load impedance on the SWR circle, add the line length, 2.35 λ, to the reference value from the wavelengths scale; 2.35 + 0.195 = 2.545. Locate the new value on the TOWARD LOAD scale. But because the calibrations extend only from 0 to 0.5, we must first subtract a number of half wavelengths from this value and use only the remaining value. In this situation, the largest integral number of half wavelengths that can be subtracted with a positive result is 5, or 2.5 λ. Thus, 2.545 – 2.5 = 0.045. Locate the 0.045 value on the TOWARD LOAD scale (at C). Draw a radial line from this value to prime center. Now, the coordinates at the intersection of the second radial line and the SWR circle represent the load impedance. To read this value closely, some interpolation between the printed coordinate lines must be made, and the value of $0.62 - j\,0.19$ is read. Multiplying by 50, we get the actual load or antenna impedance as $31 - j\,9.5$ Ω, or 31 Ω resistance with 9.5 Ω capacitive reactance.

Problems may be entered on the chart in yet another manner. Suppose we have a length of 50-Ω line feeding a base-loaded resonant vertical ground-plane antenna which is shorter than $^1/_4$ λ. Further, suppose we have an SWR monitor in the line, and that it indicates an SWR of 1.7 to 1. The line is known to be 0.95 λ long. We want to know both the input and the antenna impedances.

From the information available, we have no impedances to enter into the chart. We may, however, draw a circle representing the 1.7 SWR. We also know, from the definition of resonance, that the antenna presents a purely resistive load to the line, that is, no reactive component. Thus, the antenna impedance must lie on the resistance axis. If we were to draw such an SWR circle and observe the chart with only the circle drawn, we would see two points which satisfy the resonance requirement for the load. These points are $0.59 + j\,0$ and $1.7 + j\,0$. Multiplying by 50, we see that these values represent 29.5 and 85 Ω resistance. This may sound familiar, because, when a line is terminated in a pure resistance, the SWR in the line equals Z_R/Z_0 or Z_0/Z_R, where Z_R=load resistance and Z_0=line impedance.

If we consider antenna fundamentals, we know that the theoretical impedance of a $^1/_4$-λ ground-plane antenna is approximately 36 Ω. We therefore can quite logically discard the 85-Ω impedance figure in favor of the 29.5-Ω value. This is then taken as the load impedance value for the Smith Chart calculations. To find the line input impedance, we subtract 0.5 λ from the line length, 0.95, and find 0.45 λ on the TOWARD GENERATOR scale. (The wavelength-scale starting point in this case is 0.) The line input impedance is found to be $0.63 - j\,0.20$, or $31.5 - j\,10$ Ω.

DETERMINATION OF LINE LENGTH

In the example problems given so far in this chapter, the line length has conveniently been stated in wavelengths. The electrical length of a piece of line depends upon its physical length, the radio frequency under consideration, and the velocity of propagation in the line. If an impedance-measurement bridge is capable of quite reliable readings at high SWR values, the line length may be determined through line input-impedance measurements with short- or open-circuit line terminations. Information on the procedure is given later in this chapter. A more direct method is to measure the physical length of the line and calculate its electrical length from

$$N = \frac{f\,L}{984\ VF} \qquad \text{(Eq 1)}$$

where

N = number of electrical wavelengths in the line
L = line length in feet
f = frequency, MHz
VF = velocity or propagation factor of the line

The velocity factor may be obtained from transmission-line data tables.

LINE-LOSS CONSIDERATIONS WITH THE SMITH CHART

The example Smith Chart problems presented in the previous section ignored attenuation, or line losses. Quite frequently it is not even necessary to consider losses when making calculations; any difference in readings obtained are often imperceptible on the chart. However, when the line losses become appreciable, such as for high-loss lines, long lines, or at VHF and UHF, loss considerations may become significant in making Smith Chart calculations. This involves only one simple step, in addition to the procedures previously presented.

Because of line losses, the SWR does not remain constant throughout the length of the line. As a result, there is a decrease in SWR as one progresses away from the load. To truly present this situation on the Smith Chart, instead of drawing a constant SWR circle, it would be necessary to draw a spiral inward and clockwise from the load impedance toward the generator, as shown in **Fig 8**. The rate at which the curve spirals toward prime center is related to the attenuation in the line. Rather than drawing spiral curves, a simpler method is used in solving line-loss problems, by means of the external scale TRANSMISSION LOSS 1-DB STEPS. This scale may be seen in **Fig 9**. Because this is only a relative scale, the decibel steps are not numbered.

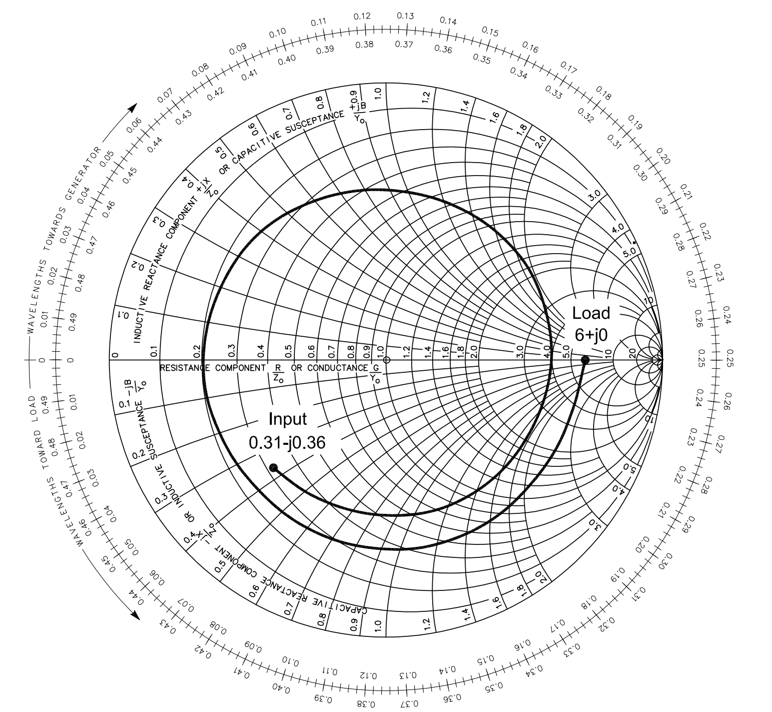


Fig 8—This spiral is the actual "SWR circle" when line losses are taken into account. It is based on calculations for a 16-ft length of RG-174 coax feeding a resonant 28-MHz 300-Ω antenna (50-Ω coax, velocity factor = 66%, attenuation = 6.2 dB per 100 ft). The SWR at the load is 6:1, while it is 3.6:1 at the line input. When solving problems involving attenuation, two constant SWR circles are drawn instead of a spiral, one for the line input SWR and one for the load SWR.

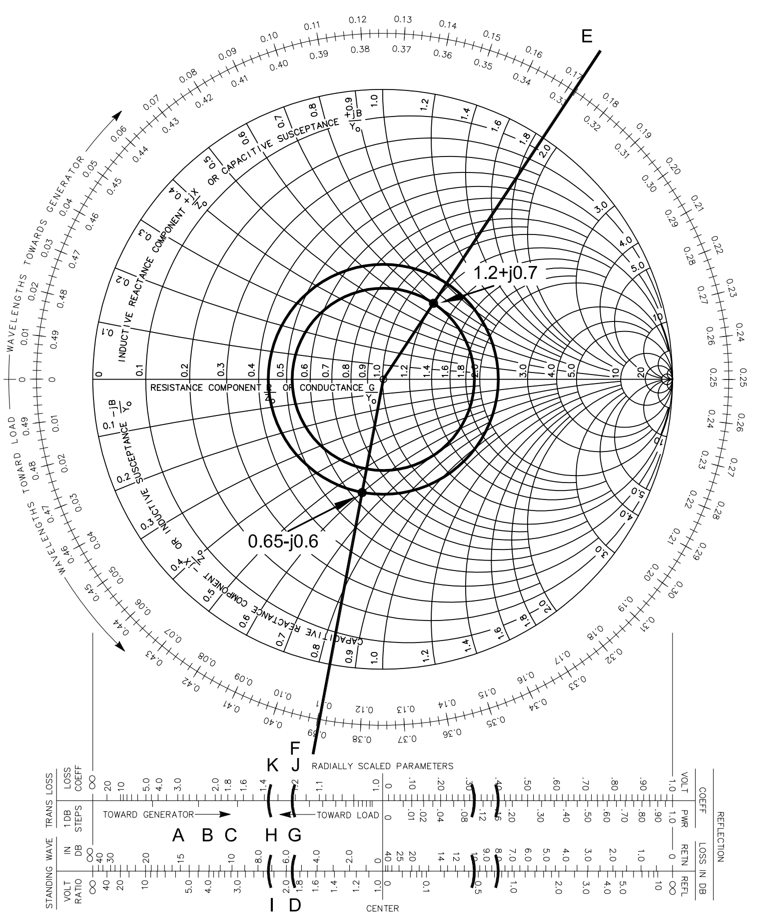


Fig 9—Example of Smith Chart calculations taking line losses into account.

If we start at the left end of this external scale and proceed in the direction indicated TOWARD GENERATOR, the first dB step is seen to occur at a radius from center corresponding to an SWR of about 9 (at A); the second dB step falls at an SWR of about 4.5 (at B), the third at 3.0 (at C), and so forth, until the 15th dB step falls at an SWR of about 1.05 to 1. This means that a line terminated in a short or open circuit (infinite SWR), and having an attenuation of 15 dB, would exhibit an SWR of only 1.05 at its input. It will be noted that the dB steps near the right end of the scale are very close together, and a line attenuation of 1 or 2 dB in this area will have only slight effect on the SWR. But near the left end of the scale, corresponding to high SWR values, a 1 or 2 dB loss has considerable effect on the SWR.

Using a Second SWR Circle

In solving a problem using line-loss information, it is necessary only to modify the radius of the SWR circle by an amount indicated on the TRANSMISSION-LOSS 1-DB STEPS scale. This is accomplished by drawing a second SWR circle, either smaller or larger than the first, depending on whether you are working toward the load or toward the generator.

For example, assume that we have a 50-Ω line that is 0.282 λ long, with 1-dB inherent attenuation. The line input impedance is measured as 60 + *j* 35 Ω. We desire to know the SWR at the input and at the load, and the load impedance. As before, we normalize the 60 + *j* 35-Ω impedance, plot it on the chart, and draw a constant SWR circle and a radial line through the point. In this case, the normalized impedance is 1.2 + *j* 0.7. From Fig 9, the SWR at the line input is seen to be 1.9 (at D), and the radial line is seen to cross the TOWARD LOAD scale, first subtract 0.500, and locate 0.110 (at F); then draw a radial line from this point to prime center.

To account for line losses, transfer the radius of the SWR circle to the external 1-DB STEPS scale. This radius crosses the external scale at G, the fifth decibel mark from the left. Since the line loss was given as 1 dB, we strike a new radius (at H), one "tick mark" to the left (toward load) on the same scale. (This will be the fourth decibel tick mark from the left of the scale.) Now transfer this new radius back to the main chart, and scribe a new SWR circle of this radius. This new radius represents the SWR at the load, and is read as 2.3 on the external VOLTAGE RATIO scale. At the intersection of the new circle and the load radial line, we read 0.65 – *j* 0.6. This is the normalized load impedance. Multiplying by 50, we obtain the actual load impedance as 32.5 – *j* 30 Ω. The SWR in this problem was seen to increase from 1.9 at the line input to 2.3 (at I) at the load, with the 1-dB line loss taken into consideration.

In the example above, values were chosen to fall conveniently on or very near the "tick marks" on the 1-dB scale. Actually, it is a simple matter to interpolate between these marks when making a radius correction. When this is necessary, the relative distance between marks for each decibel step should be maintained while counting off the proper number of steps.

Adjacent to the 1-DB STEPS scale lies a LOSS COEFFICIENT scale. This scale provides a factor by which the matched-line loss in decibels should be multiplied to account for the increased losses in the line when standing waves are present. These added losses do not affect the SWR or impedance calculations; they are merely the additional dielectric and copper losses caused by the higher voltages and currents in the presence of standing waves. For the above example, from Fig 9, the loss coefficient at the input end is seen to be 1.21 (at J), and 1.39 (at K) at the load. As a good approximation, the loss coefficient may be averaged over the length of line under consideration; in this case, the average is 1.3. This means that the total losses in the line are 1.3 times the matched loss of the line (1 dB), or 1.3 dB.

Smith Chart Procedure Summary

To summarize briefly, any calculations made on the Smith Chart are performed in four basic steps, although not necessarily in the order listed.

1) Normalize and plot a line input (or load) impedance, and construct a constant SWR circle.
2) Apply the line length to the wavelengths scales.
3) Determine attenuation or loss, if required, by means of a second SWR circle.
4) Read normalized load (or input) impedance, and convert to impedance in ohms.

The Smith Chart may be used for many types of problems other than those presented as examples here. The transformer action of a length of line—to transform a high impedance (with perhaps high reactance) to a purely resistive impedance of low value—was not mentioned. This is known as "tuning the line," for which the chart is very helpful, eliminating the need for "cut and try" procedures. The chart may also be used to calculate lengths for shorted or open matching stubs in a system, described later in this chapter. In fact, in any application where a transmission line is not perfectly matched, the Smith Chart can be of value.

ATTENUATION AND Z_0 FROM IMPEDANCE MEASUREMENTS

If an impedance bridge is available to make accurate measurements in the presence of very high SWR values, the attenuation, characteristic impedance and velocity factor of any random length of coaxial transmission line can be determined. This section was written by Jerry Hall, K1TD.

Homemade impedance bridges and noise bridges will seldom offer the degree of accuracy required to use this technique, but sometimes laboratory bridges can be found as industrial surplus at a reasonable price. It may also be possible for an amateur to borrow a laboratory type of bridge for the purpose of making some weekend measurements. Making these determinations is not difficult, but the procedure is not commonly known among amateurs. One equation treating complex numbers is used, but the math can be handled with a calculator supporting trig functions. Full details are given in the paragraphs that follow.

For each frequency of interest, two measurements are required to determine the line impedance. Just one measurement is used to determine the line attenuation and velocity factor. As an example, assume we have a 100-foot length of unidentified line with foamed dielectric, and wish to know its characteristics. We make our measurements at 7.15 MHz. The procedure is as follows.

1) Terminate the line in an open circuit. The best "open circuit" is one that minimizes the capacitance between the center conductor and the shield. If the cable has a PL-259 connector, unscrew the shell and slide it back down the coax for a few inches. If the jacket and insulation have been removed from the end, fold the braid back along the outside of the line, away from the center conductor.
2) Measure and record the impedance at the input end of the line. If the bridge measures admittance, convert the measured values to resistance and reactance. Label the values as $R_{oc} + j\ X_{oc}$. For our example, assume we measure 85 + *j*179 Ω. (If the reactance term is capacitive, record it as negative.)
3) Now terminate the line in a short circuit. If a connector exists at the far end of the line, a simple short is a mating connector with a very short piece of heavy wire soldered between the center pin and the body. If the coax has no connector, removing the jacket and center insulation from a half inch or so at the end will allow you to tightly twist the braid around the center conductor. A small clamp or alligator clip around the outer braid at the twist will keep it tight.
4) Again measure and record the impedance at the input end of the line. This time label the values as $R_{sc} \pm j\,X$. Assume the measured value now is 4.8 – *j* 11.2 Ω.

This completes the measurements. Now we reach for the calculator.

As amateurs we normally assume that the characteristic impedance of a line is purely resistive, but it can (and does) have a small capacitive reactance component. Thus, the Z_0 of a line actually consists of $R_0 + j X_0$. The basic equation for calculating the characteristic impedance is

$$Z_0 = \sqrt{Z_{oc} \times Z_{sc}} \qquad \text{(Eq 2)}$$

where

$Z_{oc} = R_{oc} + jX_{oc}$

$Z_{sc} = R_{sc} + jX_{sc}$

From Eq 2 the following working equation may be derived.

$$Z_0 = \sqrt{\left(R_{oc}R_{sc} - X_{oc}X_{sc}\right) + j\left(R_{oc}X_{sc} + R_{sc}X_{oc}\right)} \qquad \text{(Eq 3)}$$

The expression under the radical sign in Eq 3 is in the form of $R + j X$. By substituting the values from our example into Eq 3, the R term becomes $85 \times 4.8 - 179 \times (-11.2) = 2412.8$, and the X term becomes $85 \times (-11.2) + 4.8 \times 179 = -92.8$. So far, we have determined that

$$Z_0 = \sqrt{2412.8 - j92.8\Omega}$$

The quantity under the radical sign is in rectangular form. Extracting the square root of a complex term is handled easily if it is in polar form, a vector value and its angle. The vector value is simply the square root of the sum of the squares, which in this case is

$$\sqrt{2412.8^2 + 92.8^2} = \sqrt{2414.58}$$

The tangent of the vector angle we are seeking is the value of the reactance term divided by the value of the resistance term. For our example this is arctan $-92.8/2412.8$ = arctan -0.03846. The angle is thus found to be $-2.20°$. From all of this we have determined that

$$Z_0 = \sqrt{2414.58 \angle -2.20°}$$

Extracting the square root is now simply a matter of finding the square root of the vector value, and taking half the angle. (The angle is treated mathematically as an exponent.)

Our result for this example is $Z_0 = 49.1 \angle -1.1°$. The small negative angle may be ignored, and we now know that we have coax with a nominal 50-Ω impedance. (Departures of as much as 6 to 8% from the nominal value are not uncommon.) If the negative angle is large, or if the angle is positive, you should recheck your calculations and perhaps even recheck the original measurements. You can get an idea of the validity of the measurements by normalizing the measured values to the calculated impedance and plotting them on a Smith Chart as shown in **Fig 10** for this example. Ideally, the two points should be diametrically opposite, but in practice they will be not quite 180° apart and not quite the same distance from prime center. Careful measurements will yield plotted points that are close to ideal. Significant departures from the ideal indicates sloppy measurements, or perhaps an impedance bridge that is not up to the task.

Determining Line Attenuation

The short circuit measurement may be used to determine the line attenuation. This reading is more reliable than the open circuit measurement because a good short circuit is a short, while a good open circuit is hard to find. (It is impossible to escape some amount of capacitance between conductors with an "open" circuit, and that capacitance presents a path for current to flow at the RF measurement frequency.)

Use the Smith Chart and the 1-DB STEPS external scale to find line attenuation. First normalize the short circuit impedance reading to the calculated Z_0, and plot this point on the chart. See Fig 10. For our example, the normalized impedance is $4.8/49.1 - j\ 11.2\ /\ 49.1$ or $0.098 - j\ 0.228$. After plotting the point, transfer the radius to the 1-DB STEPS scale. This is shown at A of Fig 10.

Remember from discussions earlier in this document

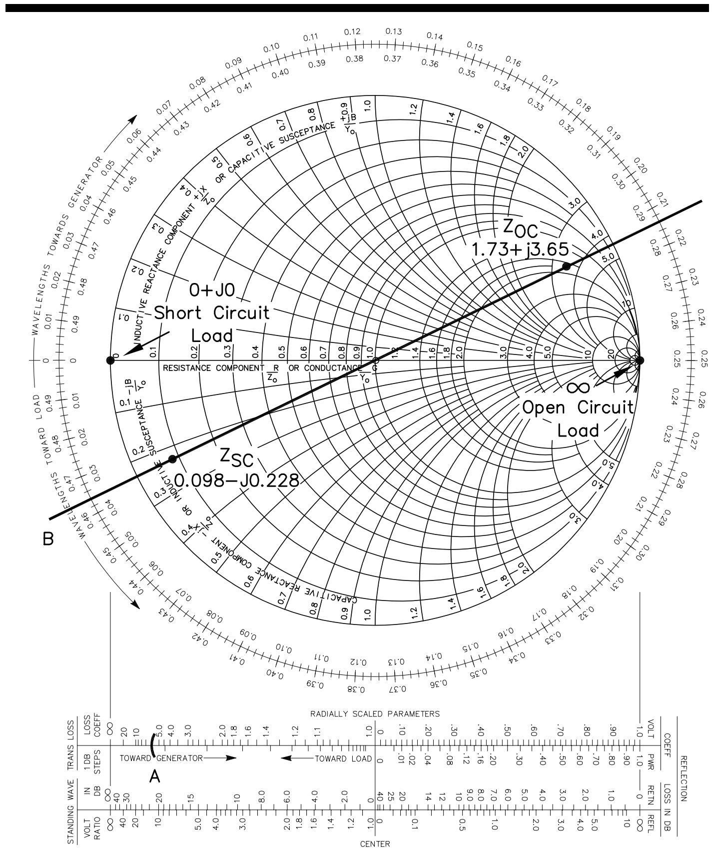


Fig 10—Determining the line loss and velocity factor with the Smith Chart from input measurements taken with open-circuit and short-circuit terminations.

that the impedance for plotting a short circuit is 0 + *j* 0, at the left edge of the chart on the resistance axis. On the 1-DB STEPS scale this is also at the left edge. The total attenuation in the line is represented by the number of dB steps from the left edge to the radius mark we have just transferred. For this example it is 0.8 dB. Some estimation may be required in interpolating between the 1-dB step marks.

Determining Velocity Factor

The velocity factor is determined by using the TOWARD GENERATOR wavelength scale of the Smith Chart. With a straightedge, draw a line from prime center through the point representing the short-circuit reading, until it intersects the wavelengths scale. In Fig 10 this point is labeled B. Consider that during our measurement, the short circuit was the load at the end of the line. Imagine a spiral curve progressing from 0 + *j* 0 clockwise and inward to our plotted measurement point. The wavelength scale, at B, indicates this line length is 0.464 λ. By rearranging the terms of Eq 1 given early in this chapter, we arrive at an equation for calculating the velocity factor.

$$VF = \frac{f\,L}{984N} \qquad \text{(Eq 4)}$$

where

VF = velocity factor
L = line length, feet
f = frequency, MHz
N = number of electrical wavelengths in the line

Inserting the example values into Eq 4 yields VF = 100 × 7.15/(984 × 0.464) = 1.566, or 156.6%. Of course, this value is an impossible number—the velocity factor in coax cannot be greater than 100%. But remember, the Smith Chart can be used for lengths greater than $^1/_2$ λ. Therefore, that 0.464 value could rightly be 0.964, 1.464, 1.964, and so on. When using 0.964 λ, Eq 4 yields a velocity factor of 0.753, or 75.3%. Trying successively greater values for the wavelength results in velocity factors of 49.6 and 37.0%. Because the cable we measured had foamed dielectric, 75.3% is the probable velocity factor. This corresponds to an electrical length of 0.964 λ. Therefore, we have determined from the measurements and calculations that our unmarked coax has a nominal 50-Ω impedance, an attenuation of 0.8 dB per hundred feet at 7.15 MHz, and a velocity factor of 75.3%.

It is difficult to use this procedure with short lengths of coax, just a few feet. The reason is that the SWR at the line input is too high to permit accurate measurements with most impedance bridges. In the example above, the SWR at the line input is approximately 12:1.

The procedure described above may also be used for determining the characteristics of balanced lines. However, impedance bridges are generally unbalanced devices, and the procedure for measuring a balanced impedance accurately with an unbalanced bridge is complicated.

LINES AS CIRCUIT ELEMENTS

Transmission-line sections may also be used as circuit elements. For example, it is possible to substitute transmission lines of the proper length and termination for coils or capacitors in ordinary circuits. While there is seldom a practical need for that application, lines are frequently used in antenna systems in place of lumped components to tune or resonate elements. Probably the most common use of such a line is in the hairpin match, where a short section of stiff open-wire line acts as a lumped inductor.

The equivalent "lumped" value for any "inductor" or "capacitor" may be determined with the aid of the Smith Chart. Line losses may be taken into account if desired, as explained earlier. See **Fig 11**. Remember that the top half of the Smith Chart coordinate system is used for impedances containing inductive reactances, and the bottom half for capacitive reactances. For example, a section of 600-Ω line $^3/_{16}$-λ long (0.1875 λ) and short-circuited at the far end is represented by ℓ1, drawn around a portion of the perimeter of the chart. The "load" is a short-circuit, 0 + *j* 0 Ω, and the TOWARD GENERATOR wavelengths scale is used for marking off the line length. At A in

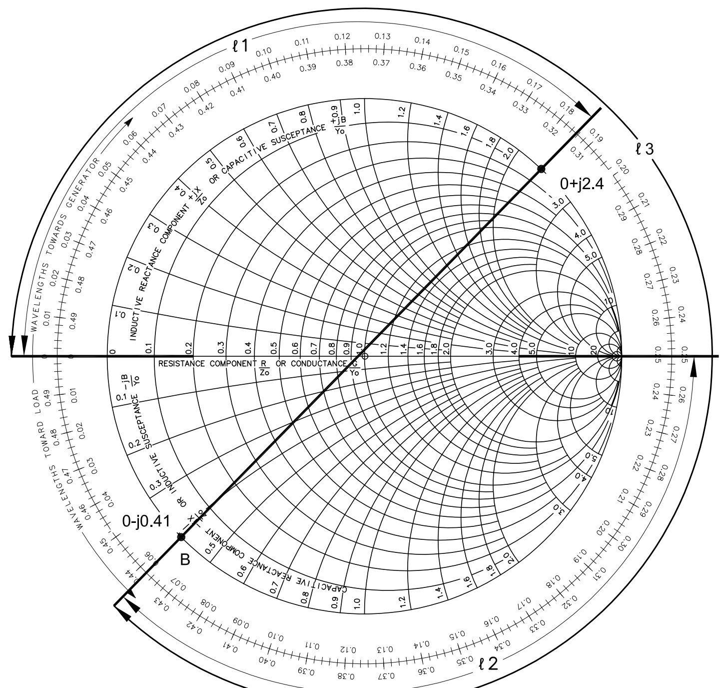


Fig 11—Smith Chart determination of input impedances for short- and open-circuited line sections, disregarding line losses.

Fig 11 may be read the normalized impedance as seen looking into the length of line, $0 + j\,2.4$. The reactance is therefore inductive, equal to $600 \times 2.4 = 1440\ \Omega$. The same line when open-circuited (termination impedance = ∞, the point at the right of the chart) is represented by $\ell 2$ in Fig 11. At B the normalized line-input impedance may be read as $0 - j\,0.41$; the reactance in this case is capacitive, $600 \times 0.41 = 246\ \Omega$. (Line losses are disregarded in these examples.) From Fig 11 it is easy to visualize that if $\ell 1$ were to be extended by $^1/_4\ \lambda$, the total length represented by $\ell 3$, the line-input impedance would be identical to that obtained in the case represented by $\ell 2$ alone. In the case of $\ell 2$, the line is open-circuited at the far end, but in the case of $\ell 3$ the line is terminated in a short. The added section of line for $\ell 3$ provides the "transformer action" for which the $^1/_4$-λ line is noted.

The equivalent inductance and capacitance as determined above can be found by substituting these values in the equations relating inductance and capacitance to reactance, or by using the various charts and calculators available. The frequency corresponding to the line length in degrees must be used, of course. In this example, if the frequency is 14 MHz the equivalent inductance and capacitance in the two cases are 16.4 µH and 46.2 pF, respectively. Note that when the line length is 45° (0.125 λ), the reactance in either case is numerically equal to the characteristic impedance of the line. In using the Smith Chart it should be kept in mind that the electrical length of a line section depends on the frequency and velocity of propagation, as well as on the actual physical length.

At lengths of line that are exact multiples of $^1/_4\ \lambda$, such lines have the properties of resonant circuits. At lengths where the input reactance passes through zero at the left of the Smith Chart, the line acts as a series-resonant circuit. At lengths for which the reactances theoretically pass from "positive" to "negative" infinity at the right of the Smith Chart, the line simulates a parallel-resonant circuit.

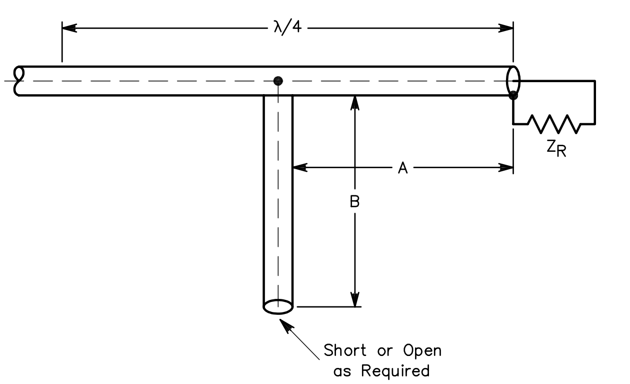


Fig 12—The method of stub matching a mismatched load on coaxial lines.

Designing Stub Matches with the Smith Chart

The design of stub matches is covered in detail in Chapter 26. Equations are presented there to calculate the electrical lengths of the main line and the stub, based on a purely resistive load and on the stub being the same type of line as the main line. The Smith Chart may also be used to determine these lengths, without the requirements that the load be purely resistive and that the line types be identical.

Fig 12 shows the stub matching arrangement in coaxial line. As an example, suppose that the load is an antenna, a close-spaced array fed with a 52-Ω line. Further suppose that the SWR has been measured as 3.1:1. From this information, a constant SWR circle may be drawn on the Smith Chart. Its radius is such that it intersects the right portion of the resistance axis at the SWR value, 3.1, as shown at point B in **Fig 13**.

Since the stub of Fig 12 is connected in parallel with the transmission line, determining the design of the matching arrangement is simplified if Smith Chart values are dealt with as admittances, rather than impedances. (An admittance is simply the reciprocal of the associated impedance. Plotted on the Smith Chart, the two associated points are on the same SWR circle, but diametrically opposite each other.) Using admittances leaves less chance for errors

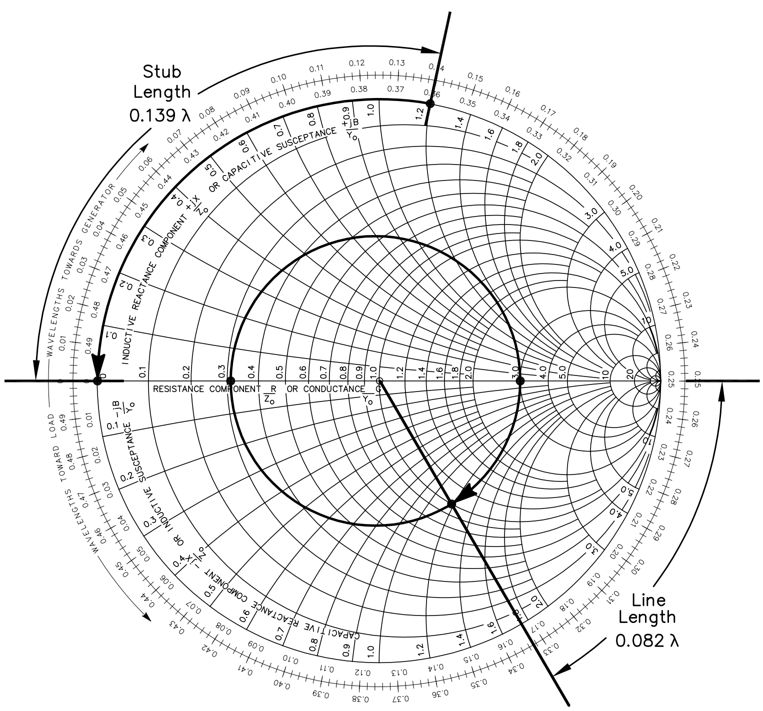


Fig 13—Smith Chart method of determining the dimensions for stub matching.

in making calculations, by eliminating the need for making series-equivalent to parallel-equivalent circuit conversions and back, or else for using complicated equations for determining the resultant value of two complex impedances connected in parallel.

A complex impedance, Z, is equal to R + j X. The equivalent admittance, Y, is equal to G – j B, where G is the conductive component and B the susceptance. (Inductance is taken as negative susceptance, and capacitance as positive.) Conductance and susceptance values are plotted and handled on the Smith Chart in the same manner as resistance and reactance.

Assuming that the close-spaced array of our example has been resonated at the operating frequency, it will present a purely resistive termination for the load end of the 52-Ω line. It is known that the impedance of the antenna equals Z_0/SWR = 52/3.1 = 16.8 Ω. (We can logically discard the possibility that the antenna impedance is SWR × Z_0, or 0.06 Ω.) If this 16.8-Ω value were to be plotted as an impedance on the Smith Chart, it would first be normalized (16.8/52 = 0.32) and then plotted as 0.32 + j 0. Although not necessary for the solution of this example, this value is plotted at point A in Fig 13. What is necessary is a plot of the admittance for the antenna as a load. This is the reciprocal of the impedance; 1/16.8 Ω equals 0.060 siemen. To plot this point it is first normalized by multiplying the conductance and susceptance values by the Z_0 of the line. Thus, (0.060 + j 0) × 52 = 3.1 + j 0. This admittance value is shown plotted at point B in Fig 13. It may be seen that points A and B are diametrically opposite each other on the chart. Actually, for the solution of this example, it wasn't necessary to compute the values for either point A or point B as in the above paragraph, for they were both determined from the known SWR value of 3.1. As may be seen in Fig 13, the points are located on the constant SWR circle which was already drawn, at the two places where it intersects the resistance axis. The plotted value for point A, 0.32, is simply the reciprocal of the value for point B, 3.1. However, an understanding of the relationship between impedance and admittance is easier to gain with simple examples such as this.

In stub matching, the stub is to be connected at a point in the line where the conductive component equals the Z_0 of the line. Point B represents the admittance of the load, which is the antenna. Various admittances will be encountered along the line, when moving in a direction indicated by the TOWARD GENERATOR wavelengths scale, but all admittance plots must fall on the constant SWR circle. Moving clockwise around the SWR circle from point B, it is seen that the line input conductance will be 1.0 (normalized Z_0 of the line) at point C, 0.082 λ toward the transmitter from the antenna. Thus, the stub should be connected at this location on the line.

The normalized admittance at point C, the point representing the location of the stub, is 1 – j 1.2 siemens, having an inductive susceptance component. A capacitive susceptance having a normalized value of + j 1.2 siemens is required across the line at the point of stub connection, to cancel the inductance. This capacitance is to be obtained from the stub section itself; the problem now is to determine its type of termination (open or shorted), and how long the stub should be. This is done by first plotting the susceptance required for cancellation, 0 + j 1.2, on the chart (point D in Fig 13). This point represents the input admittance as seen looking into the stub. The "load" or termination for the stub section is found by moving in the TOWARD LOAD direction around the chart, and will appear at the closest point on the resistance/conductance axis, either at the left or the right of the chart. Moving counterclockwise from point D, this is located at E, at the left of the chart, 0.139 λ away. From this we know the required stub length. The "load" at the far end of the stub, as represented on the Smith Chart, has a normalized admittance of 0 + j 0 siemen, which is equivalent to an open circuit.

When the stub, having an input admittance of 0 + j 1.2 siemens, is connected in parallel with the line at a point 0.082 λ from the load, where the line input admittance is 1.0 – j 1.2, the resultant admittance is the sum of the individual admittances. The conductance components are added directly, as are the susceptance components. In this case, 1.0 – j 1.2 + j 1.2 = 1.0 + j 0 siemen. Thus, the line from the point of stub connection to the transmitter will be terminated in a load which offers a perfect match. When determining the physical line lengths for stub matching, it is important to remember that the velocity factor for the type of line in use must be considered.

MATCHING WITH LUMPED CONSTANTS

It was pointed out earlier that the purpose of a matching stub is to cancel the reactive component of line impedance at the point of connection. In other words, the stub is simply a reactance of the proper kind and value shunted across the line. It does not matter what physical shape this reactance takes. It can be a section of transmission line or a "lumped" inductance or capacitance, as desired. In the above example with the Smith Chart solution, a capacitive reactance was required. A capacitor having the same value of reactance can be used just as well. There are cases where, from an installation standpoint, it may be considerably more convenient to connect a capacitor in place of a stub. This is particularly true when open-wire feeders are used. If a variable capacitor is used, it becomes possible to adjust the capacitance to the exact value required.

The proper value of reactance may be determined from Smith Chart information. In the previous example, the required susceptance, normalized, was +j 1.2 siemens. This is converted into actual siemens by dividing by the line Z_0; 1.2/52 = 0.023 siemen, capacitance. The required capacitive reactance is the reciprocal of this latter value, 1/0.023 = 43.5 Ω. If the frequency is 14.2 MHz, for instance, 43.5 Ω corresponds to a capacitance of 258 pF. A 325-pF variable capacitor connected across the line 0.082 λ from the antenna terminals would provide ample adjustment range. The RMS voltage across the capacitor is

$$E = \sqrt{P \times Z_0}$$

For 500 W, for example, E = the square root of 500 × 52 = 161 V. The peak voltage is 1.41 times the RMS value, or 227 V.

The Series-Section Transformer

The series-section transformer can be designed graphically with the aid of a Smith Chart. This information is based on a *QST* article by Frank A. Regier, OD5CG. Using the Smith Chart to design a series-section match requires the use of the chart in its less familiar off-center mode. This mode is described in the next two paragraphs.

Fig 14 shows the Smith Chart used in its familiar centered mode, with all impedances normalized to that of the transmission line, in this case 75 Ω, and all constant SWR circles concentric with the normalized value r = 1 at the chart center. An actual impedance is recovered by multiplying a chart reading by the normalizing impedance of 75 Ω. If the actual (unnormalized) impedances represented by a constant SWR circle in Fig 14 are instead divided by a normalizing impedance of 300 Ω, a different picture results. A Smith Chart shows all possible impedances, and so a closed path such as a constant SWR circle in Fig 14 must again be represented by a closed path. In fact, it can be shown that the path remains a circle, but that the constant SWR circles are no longer concentric. **Fig 15** shows the circles that result when the impedances along a mismatched 75-Ω line are normalized by dividing by 300 Ω instead of 75. The constant SWR circles still surround the point corresponding to the characteristic impedance of the line (r = 0.25) but are no longer concentric with it. Note that the normalized impedances read from corresponding points on Figs 14 and 15 are different but that the actual, unnormalized, impedances are exactly the same.

An Example

Now turn to the example shown in **Fig 16**. A complex load of $Z_L = 600 + j\ 900\ \Omega$ is to be fed with 300-Ω line, and a 75-Ω series section is to be used. These characteristic impedances agree with those used in Fig 15, and thus Fig 15 can be used to find the impedance variation along the 75-Ω series section. In particular, the constant SWR circle which passes through the Fig 15 chart center, SWR = 4 in this case, passes through all the impedances (normalized to 300 Ω) which the 75-Ω series section is able to match to the 300-Ω main line. The length $\ell 1$ of 300-Ω line has the job of transforming the load impedance to some impedance on this matching circle.

Fig 17 shows the whole process more clearly, with all impedances normalized to 300 Ω. Here the normalized load impedance $Z_L = 2 + j\ 3$ is shown at R, and the matching circle appears centered on the resistance axis and passing through the points r = 1 and $r = n^2 = (75/300)^2 = 0.0625$. A constant SWR circle is drawn from R to an intersection with the matching circle at Q or Q′ and the corresponding length $\ell 1$ (or $\ell 1'$) can be read directly from the Smith Chart. The clockwise distance around the matching circle represents the length of the matching line, from either Q′ to P or from Q to P. Because in this example the distance QP is the shorter of the two for the matching section, we choose the length $\ell 1$ as shown. By using values from the TOWARD GENERATOR scale, this length is found as 0.045 – 0.213, and adding 0.5 to obtain a positive result yields a value of 0.332 λ.

Although the impedance locus from Q to P is shown in Fig 17, the length $\ell 2$ cannot be determined directly from this chart. This is because the matching circle is not concentric with the chart center, so the wavelength scales do not apply to this circle. This problem is overcome by forming **Fig 18**, which is the same as Fig 17 except that all normalized impedances have been divided by n = 0.25, resulting in a Smith Chart normalized to 75 Ω instead of 300. The matching circle and the chart center are now concentric, and the series-section length $\ell 2$, the distance between Q and P, can be taken directly from the chart. By again using the TOWARD GENERATOR scale, this length is found as 0.250 – 0.148 = 0.102 λ.

In fact it is not necessary to construct the entire impedance locus shown in Fig 18. It is sufficient to plot Z_Q/n (Z_Q is read from Fig 17) and $Z_p/n = 1/n$, connect them by a circular arc centered on the chart center, and to determine the arc length $\ell 2$ from the Smith Chart.

Procedure Summary

The steps necessary to design a series-section transformer by means of the Smith Chart can now be listed:

1) Normalize all impedances by dividing by the characteristic impedance of the main line.
2) On a Smith Chart, plot the normalized load impe-dance Z_L at R and construct the matching circle so that its center is on the resistance axis and it passes through the points r = 1 and $r = n^2$.
3) Construct a constant SWR circle centered on the chart center through point R. This circle should intersect the matching circle at two points. One of these points, normally the one resulting in the shorter clockwise distance along the matching circle to the chart center, is chosen as point Q, and the clockwise distance from R to Q is read from the chart and taken to be $\ell 1$.
4) Read the impedance Z_Q from the chart, calculate Z_Q/n and plot it as point Q on a second Smith Chart. Also plot r = 1/n as point P.
5) On this second chart construct a circular arc, centered on the chart center, clockwise from Q to P. The length of this arc, read from the chart, represents $\ell 2$. The design of the transformer is now complete, and the necessary physical line lengths may be determined.

The Smith Chart construction shows that two design solutions are usually possible, corresponding to the two intersections of the constant SWR circle (for the load) and the matching circle. These two values correspond to positive and negative values of the square-root radical in the equation for a mathematical solution of the problem. It may happen, however, that the load circle misses the matching circle completely, in which case no solution is possible. The cure is to enlarge the matching circle by choosing a series section whose impedance departs more from that of the main line.

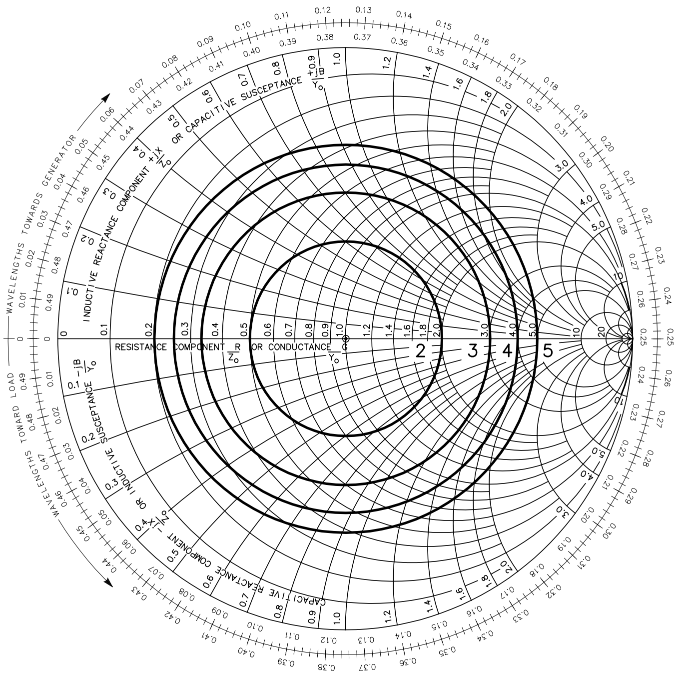


Fig 14—Constant SWR circles for SWR = 2, 3, 4 and 5, showing impedance variation along 75-Ω line, normalized to 75 Ω. The actual impedance is obtained by multiplying the chart reading by 75 Ω.

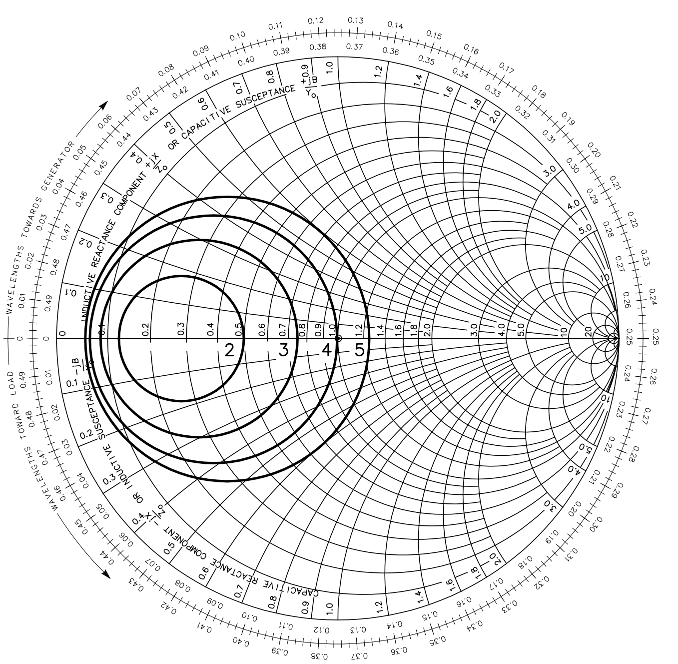


Fig 15—Paths of constant SWR for SWR = 2, 3, 4 and 5, showing impedance variation along 75-Ω line, normalized to 300 Ω. Normalized impedances differ from those in Fig 14, but actual impedances are obtained by multiplying chart readings by 300 Ω and are the same as those corresponding in Fig 14. Paths remain circles but are no longer concentric. One, the matching circle, SWR = 4 in this case, passes through the chart center and is thus the locus of all impedances which can be matched to a 300-Ω line.

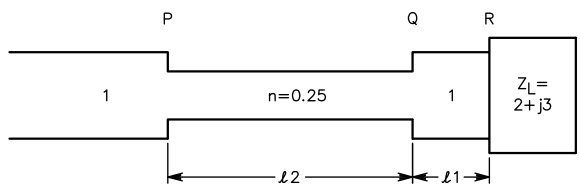


Fig 16—Example for solution by Smith Chart. All impedances are normalized to 300 Ω.

A final possibility is that, rather than intersecting the matching circle, the load circle is tangent to it. There is then but one solution—that of the 1/4-λ transformer.

BIBLIOGRAPHY

Source material and more extended discussion of topics covered in this chapter can be found in the references given below and in the textbooks listed at the end of the Antenna Fundamentals chapter of *The ARRL Antenna Book.*

W. N. Caron, *Antenna Impedance Matching* (Newington: ARRL, 1989).

C. MacKeand, "The Smith Chart in BASIC," *QST*, Nov 1984, pp 28-31.

M. W. Maxwell, *Reflections—Transmission Lines and Antennas* (Newington: ARRL, 1990).

F. A. Regier, Series-Section Transmission-Line Impedance Matching," *QST*, Jul 1978, pp 14-16.

P. H. Smith, *Electronic Applications of the Smith Chart*, reprint ed. (Malabar, FL: Krieger Pub Co, Inc, 1983).

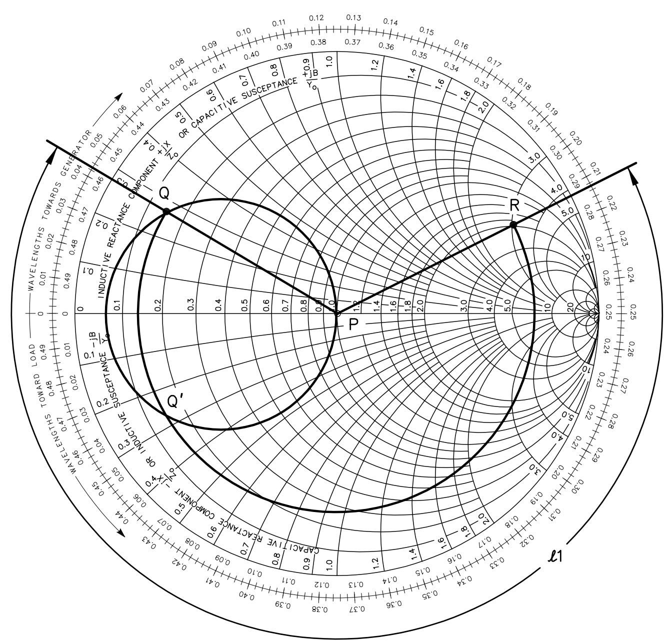


Fig 17—Smith Chart representation of the example shown in Fig 16. The impedance locus always takes a clockwise direction from the load to the generator. This path is first along the constant SWR circle from the load at R to an intersection with the matching circle at Q or Q′, and then along the matching circle to the chart center at P. Length ℓ1 can be determined directly from the chart, and in this example is 0.332 λ.

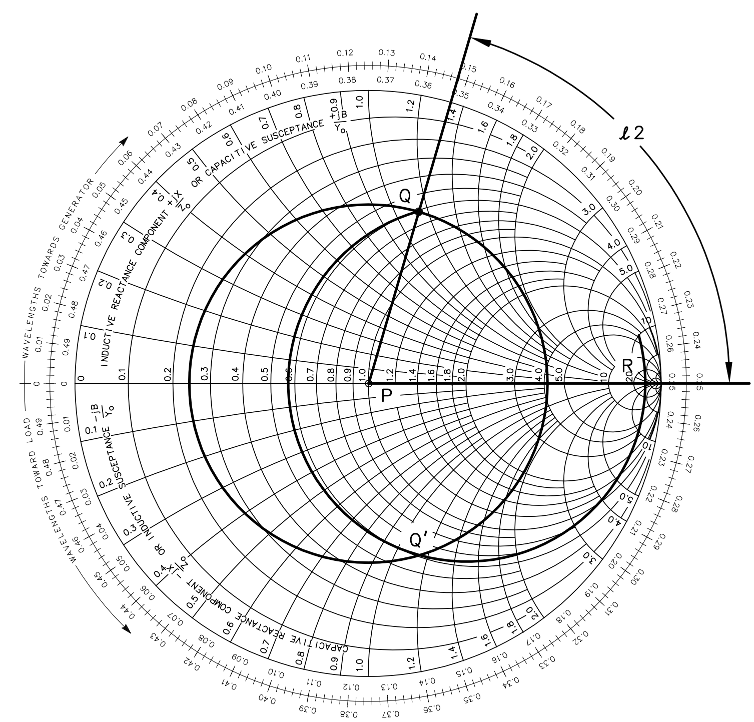


Fig 18—The same impedance locus as shown in Fig 17 except normalized to 75 Ω instead of 300. The matching circle is now concentric with the chart center, and ℓ2 can be determined directly from the chart, 0.102 λ in this case.

Transmission Lines in Digital Circuits

The performance of digital logic families covers a wide range of signal transition times. The signal rise and fall times are most important when considering how to construct a circuit. The operating frequency of a circuit is not the primary consideration. A circuit that uses high-speed logic yet runs only at a few kHz can be difficult to tame if long point-to-point wiring is used.

If the path between two points has a delay of more than 1/6 the logic family rise time, some form of transmission line should be considered. We know that waves propagate at 300 million meters per second in air and at 0.66 times as fast in common coax cable. So, in about 5 ns a wave will travel 1 meter or in 1 ns it will travel 0.2 m.

Consider a logic family which has 2 ns rise and fall time. Using the rule mentioned above, if the path length exceeds 0.066 meter or about 2.6 inches we need to use a transmission line. Another way to look at it is the approximate equivalent analog bandwidth is:

$$BW = \frac{0.35}{\text{rise time}} = \frac{0.35}{2\ \text{ns}} = 175\ \text{MHz}$$

If we were building an analog circuit that operated at 175 MHz, we would have to keep the wire lengths down to a fraction of an inch. So, even if our logic's clock is running at a few kHz, we still need to use these short wire lengths. But, suppose we are building a non-trivial circuit that has a number of gates and other digital blocks to interconnect. In order to reach several ICs from the clock's source, we will need to run wires over several inches in length.

It is possible to build a high speed circuit in breadboard style if small coax cable is used for interconnections. However, if a PC board is designed with *microstrip* transmission line interconnections, success is more likely. **Figure 1** shows the way microstrip transmission line is made. Typical dimensions are shown for 1/16-inch thick FR4 material and 50-Ω line.

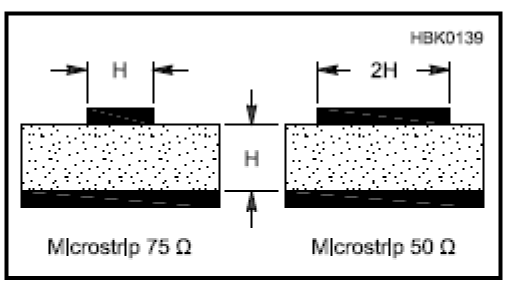


Figure 1 — Microstrip transmission lines. The approximate geometries to produce 75 Ω (A) and 50 Ω (B) microstrip lines with FR-4 PC board material are shown. This technique is used at UHF and microwave frequencies.

There are several ways that the actual circuit can be configured to assure that the desired signal reaches the receiving device input. To avoid multiple reflections that would distort the signals and possibly cause false triggering, the line should either be matched at the load end or the source end. We know that matching at the load end will completely absorb the signal so that there are no reflections. However, the signal level will be reduced because of voltage division with the source impedance in the sending gate. The dc levels will also shift because of the load, reducing the logic noise immunity. A considerable amount of power can be dissipated in the load, which might overload the source gate particularly if 50-Ω line is in use. It is possible to put a capacitor in series with the load resistor, but only if the waveform duty cycle is near 50%. If not, the dc average voltage will reduce the noise immunity of the receiving gate.

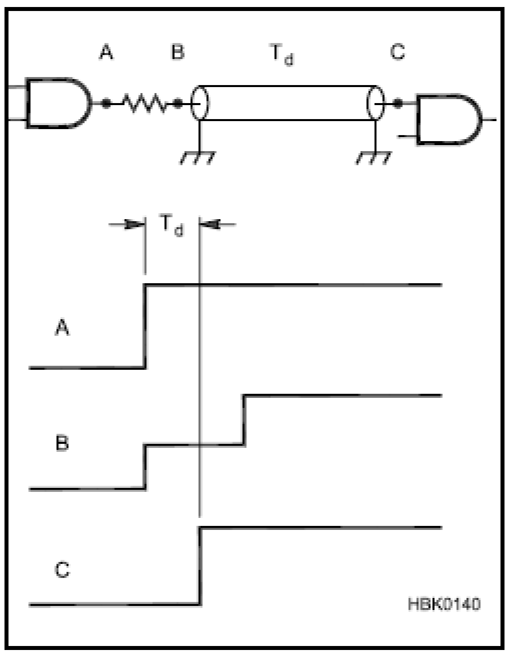


Figure 2 — Reflections in a transmission line cause stepping in the leading edge of a digital pulse at the generator (A). By adding a resistor in series with the gate output, a step is generated at the input to the transmission line (B), but a full-voltage step is created at the high input impedance of the receiving gate.

A better method is to match the transmission line at the signal source as in **Figure 2**. A resistor is added in series with the output of the sending gate, raising the gate output impedance to match Z_0 of the transmission line, which is connected to the resistor at B and has an arbitrary length T_D. No load is required on the receiving end of the transmission line, which is assumed to be connected to a gate with an input impedance much higher than Z_0.

When the sending output goes high at point A, generating the leading edge of a pulse with voltage V, the load on the output resistor is equal to the Z_0 of the transmission line so a voltage divider is formed and the voltage at point B initially goes to 1/2 V. The pulse travels down the transmission line and after T_D it completely reflects off the open end at the receiving gate. The voltage at C reaches V since the direct signal and the reflected signal add together. When the reflected edge of the pulse returns to point B after a round trip time of $2T_D$, the voltage level at B increases to V.

The receiving end can be terminated in Z_0 if a pair of resistors, each equal to $2 \times Z_0$, are connected from the positive power supply to ground at point C. The transmission line and gate input are connected to the resistor junction. The main problem with this method is the steady-state current required by the resistors. Some logic gates may not have adequate current output to drive this load.

Tuned (Resonant) Networks

(excerpted from Chapter 14 of the ARRL Handbook, 2009 and previous editions)

There is a large class of LC networks that utilize resonance at a single frequency to transform impedances over a narrow band. In many applications the circuitry that the network connects to has internal reactances, inductive or capacitive, combined with resistance. We want to absorb these reactances, if possible, to become an integral part of the network design. By looking at the various available network possibilities we can identify those that will do this at one or both ends of the network. Some networks must operate between two different values of resistance, others can also operate between equal resistances. As mentioned before, nearly all networks also allow a choice of selectivity, or Q, where Q is (approximately) the resonant frequency divided by the 3-dB bandwidth.

As a simple example that illustrates the method, consider the generator and load of **Fig 14.58A**. We want to absorb the 20 pF and the 0.1 µH into the network. We use the formulas to calculate L and C for a 500 Ω to 50 Ω L-network, then subtract 20 pF from C and 0.1 µH from L. As a second iteration we can improve the design by considering the resistance of the L that we just found. Suppose it is 2 Ω. We can recalculate new values L′ and C′ for a network from 500 Ω to 52 Ω, as shown in Fig 14.58B.

Further iterations are possible but usually trivial. More complicated networks and more difficult problems can use a computer to expedite absorbing process. Always try to absorb an inductance into a network L and a capacitance into a network C in order to minimize spurious LC resonances and undesired frequency responses. Inductors and capacitors can be combined in series or in parallel as shown in the example. Fig 14.58C shows useful formulas to convert series to parallel and vice versa to help with the designs.

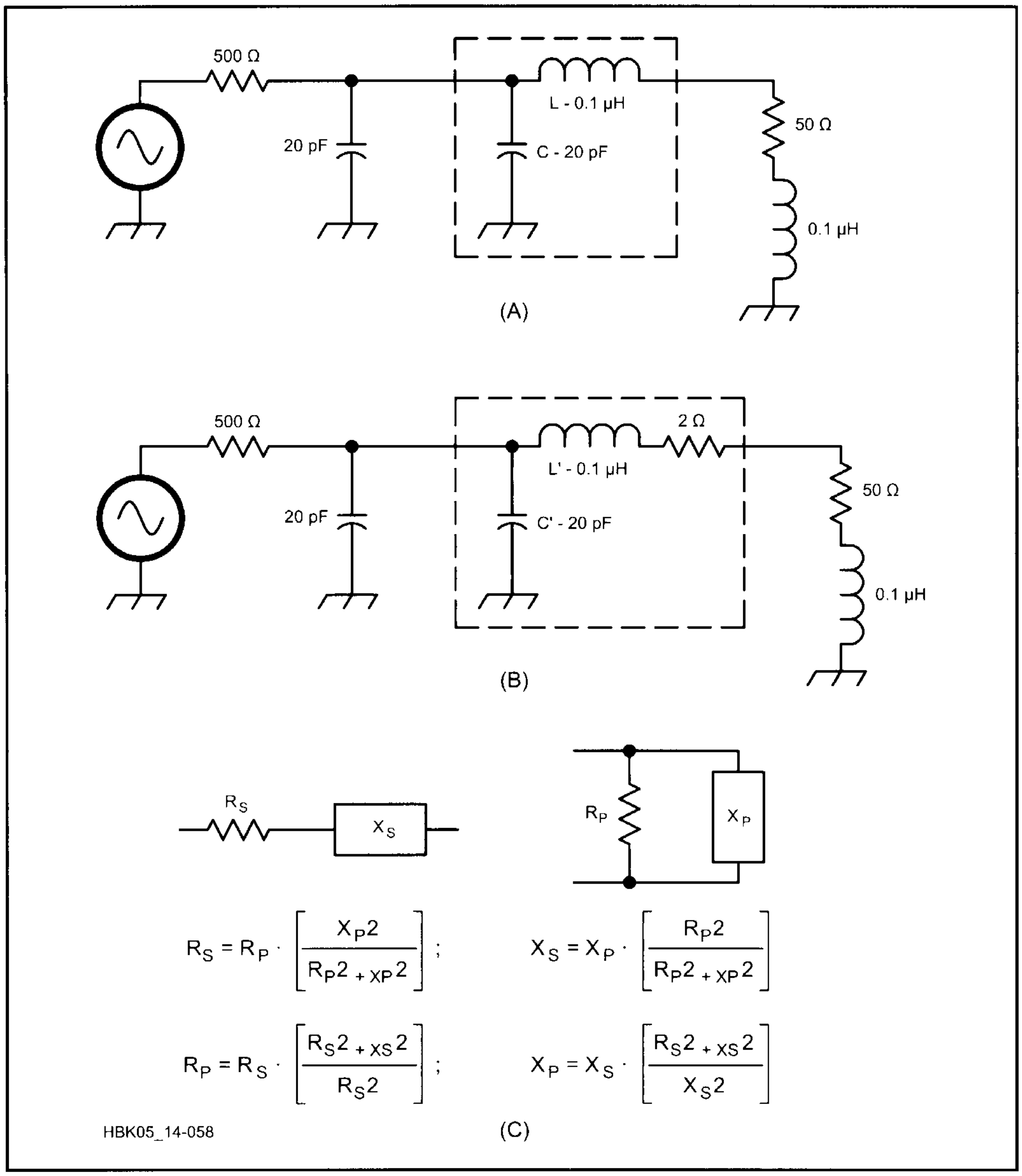
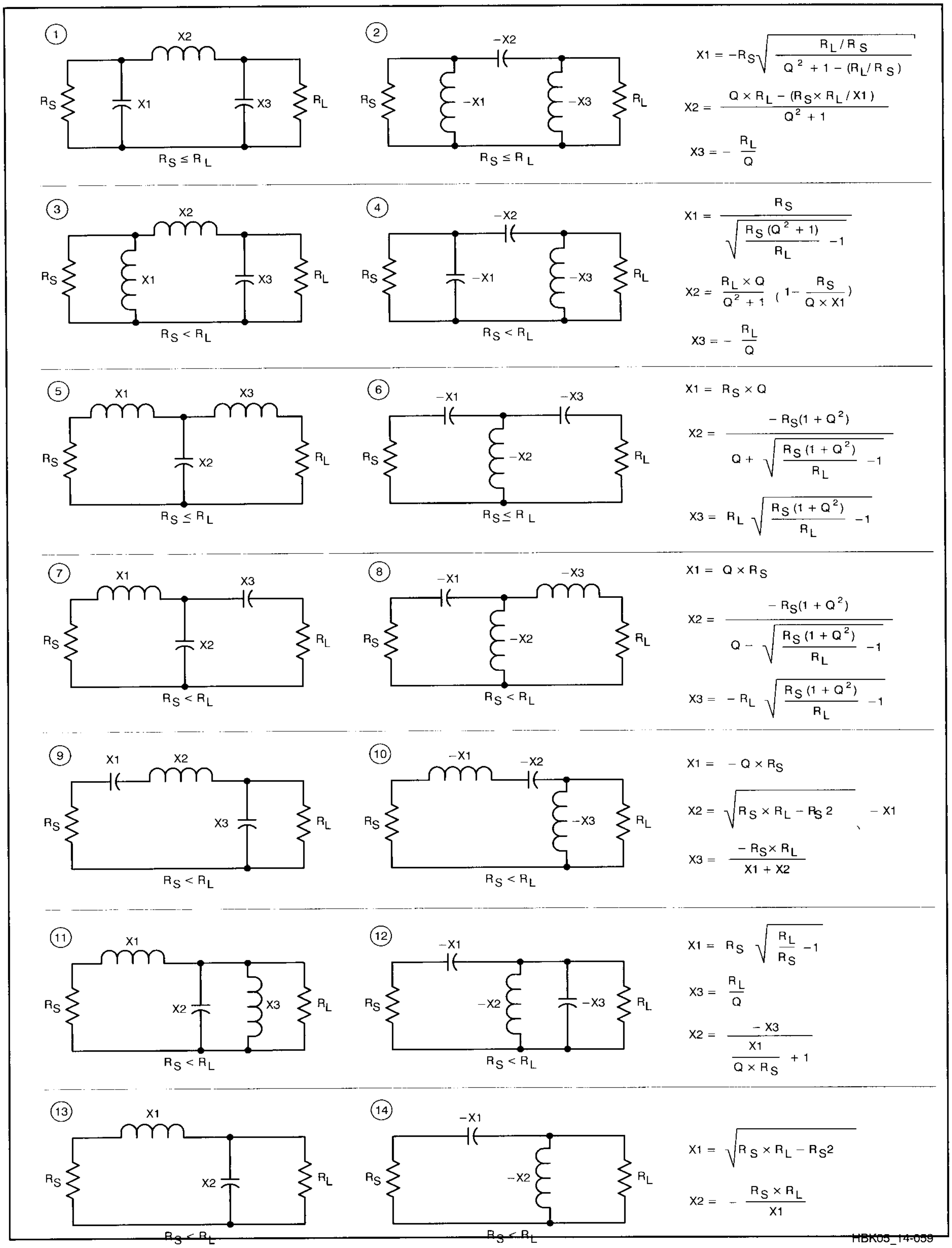


Fig 14.58 — At A, impedance transformation, first iteration. At B, second iteration compensates L and C values for coil resistance. At C, series-parallel conversions.

A set of 14 simple resonant networks, and their equations, is presented in **Fig 14.59**. Note that in these diagrams RS is the low impedance side and RL is the high impedance side and that the X values are calculated in the top-down order given. The program *MATCH.EXE* can perform the calculations.



Networks 1 and 2 ($R_S \le R_L$):

$$X1 = -R_S\sqrt{\frac{R_L/R_S}{Q^2 + 1 - (R_L/R_S)}}$$

$$X2 = \frac{Q \times R_L - (R_S \times R_L / X1)}{Q^2 + 1}$$

$$X3 = -\frac{R_L}{Q}$$

Networks 3 and 4 ($R_S < R_L$):

$$X1 = \frac{R_S}{\sqrt{\frac{R_S(Q^2+1)}{R_L} - 1}}$$

$$X2 = \frac{R_L \times Q}{Q^2 + 1}\left(1 - \frac{R_S}{Q \times X1}\right)$$

$$X3 = -\frac{R_L}{Q}$$

Networks 5 and 6 ($R_S \le R_L$):

$$X1 = R_S \times Q$$

$$X2 = \frac{-R_S(1+Q^2)}{Q + \sqrt{\frac{R_S(1+Q^2)}{R_L} - 1}}$$

$$X3 = R_L\sqrt{\frac{R_S(1+Q^2)}{R_L} - 1}$$

Networks 7 and 8 ($R_S < R_L$):

$$X1 = Q \times R_S$$

$$X2 = \frac{-R_S(1+Q^2)}{Q - \sqrt{\frac{R_S(1+Q^2)}{R_L} - 1}}$$

$$X3 = -R_L\sqrt{\frac{R_S(1+Q^2)}{R_L} - 1}$$

Networks 9 and 10 ($R_S < R_L$):

$$X1 = -Q \times R_S$$

$$X2 = \sqrt{R_S \times R_L - R_S^2} - X1$$

$$X3 = \frac{-R_S \times R_L}{X1 + X2}$$

Networks 11 and 12 ($R_S < R_L$):

$$X1 = R_S\sqrt{\frac{R_L}{R_S} - 1}$$

$$X3 = \frac{R_L}{Q}$$

$$X2 = \frac{-X3}{\frac{X1}{Q \times R_S} + 1}$$

Networks 13 and 14 ($R_S < R_L$):

$$X1 = \sqrt{R_S \times R_L - R_S^2}$$

$$X2 = -\frac{R_S \times R_L}{X1}$$

Fig 14.59 — Fourteen impedance transforming networks with their design equations (for lossless components).

Circuit simulation programs can also help a lot with special circuit-design problems and some approaches to resonant network design. It can graph the frequency response, compute insertion loss and also tune the capacitances and inductances across a frequency band. You may select the selectivity (Q) in such programs based on frequency-response requirements. The program can also be trimmed to help realize realistic or standard component values. A math program such as *Mathcad* can also make this a quick and easy process.

Using *TLW* to Design Impedance Matching Networks

By George Cutsogeorge, W2VJN

There are many devices available to amateurs that will measure the complex impedance at the end of a transmission line. For example: the MFJ 269, the LP-100A wattmeter, the AIM4170, any vector network analyzer (VNA), etc. Given an impedance in the series format R + *j*X, the software *TLW* by Dean Straw, N6BV, distributed with the *ARRL Antenna Book,* can determine the components required to match it.

Let us go through some examples to see how we can use *TLW*. We can use some of the values shown in the table for a 100-foot center fed dipole. (See the supplemental article "Multiband Operation with Open-Wire Line" for more about this type of antenna system.) Suppose we would like to use the antenna on 3.8 MHz. First of all, we should be feeding it with 450-Ω line as the losses are excessive with RG-213. After we open *TLW*, for Cable Type we select 450-ohm Window Ladder Line. Set the length to 100 feet. Next we enter the impedance value from **Table 1**, 37.5 – *j*354 Ω. Other settings are: Frequency = 3.8, Source = Normal and select Load.

Table 1
Modeled Data for a 100-ft Flat-Top Antenna

Freq (MHz)	*Antenna Impedance (Ω)*	*Input VSWR RG-213 Coax*	*Loss of 100 ft RG-213 Coax (dB)*	*Loss of 100 ft 450- Ω Line (dB)*	*Max Voltage RG-213 Coax at 1500 W*	*Max Voltage 450- Ω Line at 1500 W*
1.8	4.18 –*j* 1590	33.7	26.0	8.8	1507	10950
3.8	37.5 –*j* 354	16.7	5.7	0.5	1177	3231
7.1	447 +*j* 956	12.3	5.9	0.2	985	2001
10.1	2010 –*j* 2970	12.1	10.1	0.6	967	2911
14.1	87.6 –*j* 156	4.6	2.4	0.3	587	1747
18.1	1800 +*j* 1470	7.7	6.8	0.3	753	1600
21.1	461 –*j* 1250	4.6	3.2	0.1	585	828
24.9	155 +*j* 150	3.6	2.6	0.2	516	1328
28.4	2590 +*j* 772	6.7	9.4	0.5	703	1950

These values were computed using version 3 of the antenna modeling program, *EZNEC* (**www.eznec.com**). Antenna impedance computed using 499 segments and with the Real Ground model.

At the bottom of the screen things have been happening. We see the SWR at the line input is 16.97, the loss is 0.461 dB and the impedance at the line input is 171.02 – *j*1000.26 Ω.

Now we can look at the various impedance matching designs. There are four Network Types available. We can select them one at a time and compare some characteristics as shown in **Table 2**.

Table 2
Conditions for Different Matching Networks

	Loss	*Cap Voltage*	*Coil Current*
High-Pass L	8.9%	4039V	8.2A
Low-Pass L	5.8%	4125V	5.5A
Low-Pass Pi	33.9%	3457V	31.6A
High-Pass Tee	14.2%	5926V	8.9A

The Low-Pass L network has the lowest loss and would be preferable due to that alone. But, the capacitor voltage is very high and requires a physically large part. Maybe there is something we can do about that. Select Volt/Current and press Graph. Note that with 100 feet of transmission line we are close to a voltage maximum at the transmitter end. Also note there is about 60 feet between the current and voltage maximums. If we add some transmission line, we can move the voltage minimum to the end of the line. It looks like the voltage maximum is about 15 feet down the line, so if we add 60 – 15 = 45 feet we should have a voltage minimum at the line input. Re-plot the Volt/Current graph with 145 feet of line and verify. Now let us look at the various networks again.

Table 3
Conditions with Additional Line Added

	Loss	*Cap Voltage*	*Coil Current*
High-Pass L	0.6%	298V	5.5A
Low-Pass L	0.6%	387V	7.7A
Low-Pass Pi	0.8%	387V	7.9A
High-Pass Tee	15.1%	4237V	12.6A

What an amazing difference for a few cents worth of 450 Ω line. Note the losses are way down in the tuner and the voltages are quite low. Capacitors rated about twice the peak voltage listed voltage should be adequate. Either the Low Pass or High Pass L network will work fine.

Note that the values required in an actual antenna may vary from these calculated values due to different ground parameters or surrounding objects. These numbers are a good starting point, however.

TLW can also be used to design networks for high impedances such as plate tank circuits. Let's try an example. First, select User Defined Transmission Line. In the resulting table set Enter Attenuation to 0.000 and Enter X0 to 0.000. Then press Compute X0. This will remove some residual reactance in the software that occurs even when the transmission line length is set to zero. Close the parameter window. Now set the transmission line length to 0.

Suppose we want to design an amplifier's Pi-network tank circuit to match a 2000-Ω plate output impedance to 50 Ω at 21.1 MHz. In the Resistance/Reactance box, select Input and then input 2000 Ω resistance and 0 Ω reactance. Press Tuner , then select Low-Pass Pi Network. In the Default Values window the Pi-Network Output Capacitor, pF will be the main tuning capacitor. Input a value such as 50 pF for a starting point. Generate the tuner. In the resulting screen, note the effective Q (Eff Q) is 14.8. Most often a Q or 12 would be used. (Pi network design for amplifiers is discussed in the **RF Power Amplifiers** chapter.) To reduce the effective Q, reduce the output capacitance and rerun the example. About 40 pF gives the desired results in this case.

The same result can be had by changing from Load to Input in the opening screen. The value for Output Stray Capacitance, pF value is a reminder not to forget the strays, but this value does not change the computed circuit.