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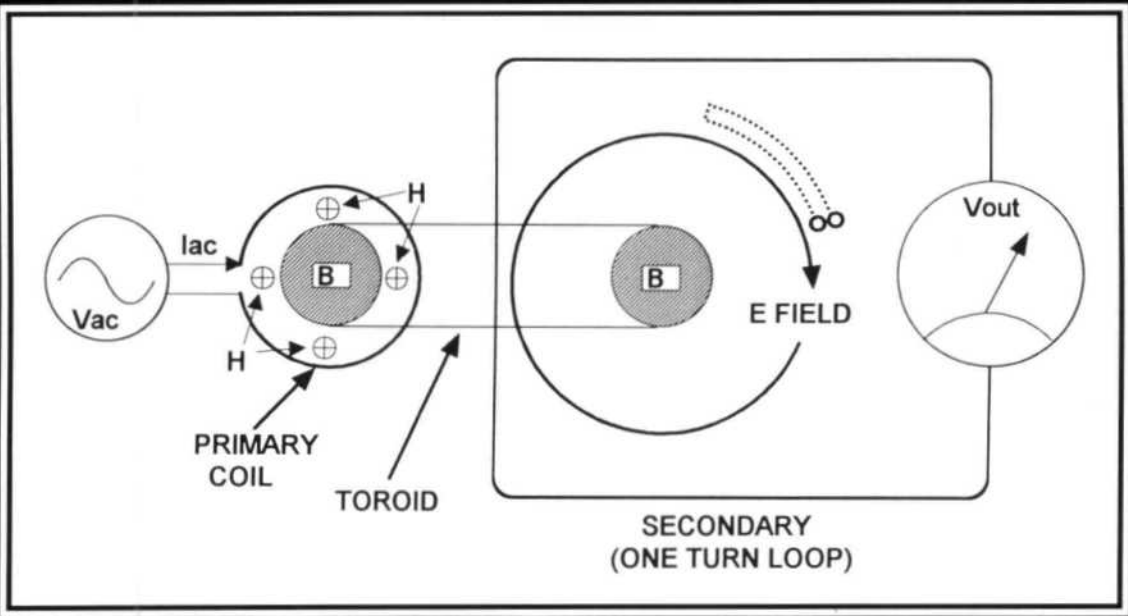
QEX

ARRL Experimenter's Exchange

May 1995



Transformers for Directional Couplers

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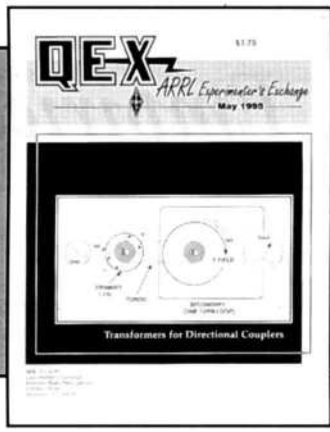

About the Cover

WØIYH explains all about how to design the transformers for your directional coupler.

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Purpose of *QEX*:

1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio field
3) support efforts to advance the state of the Amateur Radio art

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and correspondence for publication in *QEX* should be marked: Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Empirically Speaking

HF Automation Arrives

Just as this issue of *QEX* was going to press, the FCC released their Report and Order in PR Docket 94-59, concerning HF digital communication. In this document, the FCC takes long-awaited action regarding automatic control of digital stations at HF.

This subject has long been of interest to operators of digital message-forwarding stations. Previously, the only legal unattended HF forwarding was performed under the auspices of an STA issued by the FCC to ARRL in 1986, and renewed periodically thereafter. That STA expires on July 1, 1995, when the new rules go into effect.

One change ordered by the FCC will allow digital (that is, RTTY and data, *not* Morse) stations to operate under automatic control in small, defined segments of the amateur HF bands. This new rule should make it possible to continue and expand the HF components of message-forwarding networks that operate entirely without operator intervention. These are the networks that have been relying on the STA for so many years. It should come as a relief to all operators within these networks to be rid of the extra paper shuffling imposed by STA requirements.

The "other shoe" in the Report and Order is a rule that will allow digital stations to operate elsewhere on HF (where RTTY and data are permitted) under automatic control, but only if they transmit solely as a result of interrogation by a station under local or remote—not automatic—control. And these automatic stations' signals can occupy no more than a 500-Hz bandwidth. This provision is intended to serve the growing number of "server," or "mailbox" stations, using APLINK and other systems, that respond to connections initiated by the users. In this case, there must be at least one pair of ears listening to the channel: those of the initiating user. Hopefully, this will be sufficient to keep interference between these systems and other manual operations to a minimum. The 500-Hz bandwidth restriction also serves to minimize the potential for interference to other operations.

The road we've traveled to get to this change in the rules has been long and sometimes rocky. Strong opinions abound. Many in the "traditional" RTTY and Morse communities think that no automatic operation of any kind can usefully coexist with manual operations. Some who use APLINK can't see any reason for automatic-operation segments, but think the new manual-interrogation rules are fine. And some packeteers think only fully automatic operation—no control operator at either end of the link—is of any use. What the FCC has done is to craft rules that try to form the best compromise between these very different points of view.

Now it's time to lay aside the arguments and strive to make the new rules work for everyone. That means following the "prime directive" of HF operating: always try to interfere as little as possible. We can never eliminate all interference; the nature of HF propagation and nonchannelized operation make that impossible. But we can *minimize* interference. Proper choice of frequency, by conforming to band plans, even though they don't have the force of law; carefully listening on *and around* your operating frequency to ensure you aren't interfering with others; and, for automatically controlled stations, being sensitive to the complaints of operators who experience interference from your station—all of these are necessary if we are to share the spectrum. Manual operations and automatic operations both serve to further Amateur Radio. They do so best when they work in concert rather than in competition.

This Month in *QEX*

In the March issue of *QEX*, William E. Sabin, WØIYH, explained the inner workings of transformer-type directional couplers. This month, he explains "Designing the Toroid Transformer for the Directional Coupler."

Switching the antenna between the transmitter and receiver is, even today, commonly done with a slow relay because electronic switching designs often rely on expensive, or hard-to-get, PIN diodes. In "Electronic Antenna Switching," Wes Hayward, W7ZOI, shows how to use inexpensive, available devices for this purpose.

In a reprint from the February 1995 issue of *Radio Communication*, A. J. Harwood, G4HHZ, describes a novel "Graphic Method for Calculating Z" on a transmission line.

Finally, Zack Lau, KH6CP/1, discusses Yagi antenna design at VHF and above in this month's "RF" column.—*KE3Z, email: jbloom@arrl.org (Internet)*

Electronic Antenna Switching

Not only is electronic T/R switching faster than using relays, it can be less expensive, too!

By Wes Hayward, W7ZOI

Until recently, the typical amateur station switched the antenna between the receiver input and the transmitter output with a relay. The common T/R, or transmit-receive relay was relatively slow, switching in 10 or 20 milliseconds. Some specialists demanded faster operation, the traditional example being the CW enthusiast who operates "full break-in," or QSK. An effective QSK system operates fast enough that received signals can be heard between dots at code speeds of 30 words per minute, or even faster.

The demand for QSK capability has increased in recent years, largely the result of interest in digital modes that are faster and more automated than CW, the original digital mode. This article explores some methods for nonmechanical switching. Not only are the methods simple, but they are probably less expensive than a good relay.

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The Basic Requirements

The fundamental T/R element, shown in Fig 1A, is a single-pole double-throw switch. This complication may or may not be needed; the simplified SPST topology of Fig 1B is often adequate. The system requirements are:

- There should be a low-loss connection between the transmitter and the antenna when the transmitter is operating.
- The switch should attenuate signals from the transmitter so that there is no damage to the receiver input. Minor receiver overload may be tolerable.
- Antenna signals must *not* be severely attenuated in reaching the receiver input. A few dB loss is generally not a problem on the lower HF bands, but would be intolerable at VHF.
- The transmit path should not introduce distortion to a transmitted SSB signal; requirements are relaxed for CW and most other digital modes.
- Distortion in the receive path should be low enough that there is no compromise in receiver performance.

T/R Switching for QRP Rigs

A very common electronic T/R system example is shown in Fig 2. A 50-Ω antenna will usually reflect a similar load to the power amplifier collector, causing the receiver tap point to *see* 50 Ω so long as the amplifier is not functional. The low impedance drives a small capacitor with a typical reactance of 500 Ω. During receive periods, the diodes do not conduct, allowing the signal to reach the receiver. The

500-Ω inductor resonates with the capacitor to form a moderately low-loss signal path.

Switching action occurs when the two back-to-back diodes are turned on by the transmitter radio frequency energy. With each diode conducting on alternating RF half cycles, the 500-Ω capacitor is essentially grounded. It then becomes part of the low-pass transmitter network. This must, in concept, be taken into account during the design of the low-pass matching network, although it can usually be ignored. This "standard" T/R switch was popularized by W7EL in his "Optimized" transceiver.[1]

[1]Notes appear on page 7.

The maximum current capability of the diodes limits the transmit power that can be used with this scheme. Standard 1N4152 switching diodes can conduct a maximum current of about 100 mA. They are adequate for 5-W rigs in a 50-Ω system, or up to about 25 W in a 12-Ω environment. Higher powers can be accommodated with the methods presented below.

There are several performance parameters that we might measure (or calculate) in this simple circuit. The first is receive-path insertion loss. This is usually under 1 dB with typical (X=500 Ω) components, a small price (at HF) for the convenience of electronic T/R.

Next we ask how well the receiver is protected. Specifically, how much transmitter power is available at the receiver antenna terminal? My measurements show an available power of −10 dBm, at least 20 dB below the level that would cause damage, but still a very "loud" signal. Effective receiver muting circuitry is needed for smooth station operation. The power available at the receiver does not depend strongly on the transmitter power being used.

The T/R circuit is in the receiver signal path and should be evaluated for IMD performance. Less than ideal performance might compromise receiver dynamic range. This is rarely of concern in low-power portable stations used on quiet bands. It can become a major concern to a 40-meter operator in Europe or even on the east coast of North America.[2]

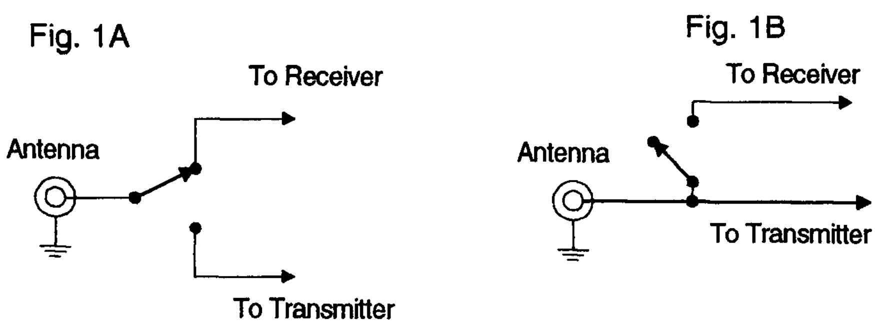

Fig 1—Basic SPDT antenna switch is shown at A. The configuration in B is suitable if a direct connection between transmitter and antenna does not cause problems during receive periods.

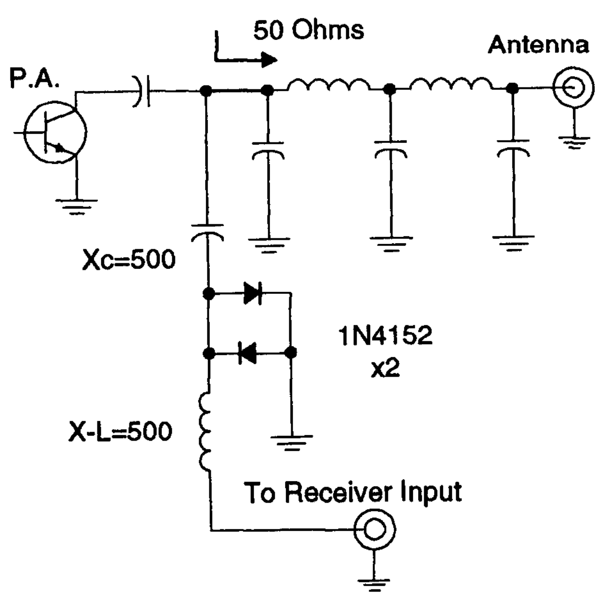

Fig 2—Simple T/R system for use with QRP transmitters. This scheme is suitable for powers up to 25 W or more. See text.

A 20-meter version of the circuit of Fig 2 was built and evaluated for insertion loss and 3rd-order input intercept. This band was picked because measurement equipment was available. The T/R series-tuned circuit used a small trimmer capacitor (22 pF) and a 5-µH inductor wound on a T50-2 toroid. The measured input intercept was −3 dBm. This is low enough to compromise the performance of *modern* receivers, including many direct conversion circuits using diode-ring-mixer front ends.

Each diode was replaced with a series connection of two diodes, increasing the measured input intercept to +7 dBm. The same circuit with three diodes in each leg produced an $IP3_{in}$ of +13.5 dBm. This performance approaches that needed for use with wide-dynamic-range receivers.

A 2-W transmitter using this T/R circuit was evaluated for receiver protection with multiple diodes. The available power at the receiver antenna terminal was −10 dBm with one diode per leg, −4 dBm with two diodes per leg, and −1 dBm with three diodes per leg. While adding diodes reduces the protection, the power available to the receiver is still well below the damage level.

The QRP scheme can be extended to higher powers with a single switch element (Fig 3) where the back-to-back diodes are replaced by a shunt PIN diode. The bias must now be controlled with the key or VOX. A bipolar transistor has also been used, although I've seen no performance reports. A suitable reverse bias is required during receive periods when a single diode is used. This is required owing to the capacitance of the PIN diodes with zero bias. PIN diodes will be discussed more below.

Shown in Fig 4A is a scheme that I have used at the 100-W level with an amplifier operating from a 12-V supply. This scheme is still safe at the 100-W level because the L-C circuit in the T/R system was changed to keep the peak current at or below 100 mA in the 1N4152 diodes. Generally, with a N:1 turns ratio transformer from a single transistor power amplifier, the peak diode current will be

$$I_{peak} = \frac{N \cdot C \cdot F \cdot V_{cc}}{159.2}$$

where C is in pF, V_{cc} is in volts, F is in MHz, and I is in mA. N is doubled if a push-pull amplifier is used. In the example, the transformer is 1 center-tapped turn to 4 turns, so N becomes 8. At 7 MHz, a 17-pF capacitor with V_{cc}=12 yields a peak diode current of 72 mA, still a safe level with the 1N4152. The inductor used in this instance was 30 µH, the series combination of two 15 µH RF chokes. The Q for these chokes was about 50, which produced an insertion loss of about 2 dB; this level was acceptable at 7 MHz.

Part B of Fig 4 shows another refinement. The single L-C circuit and diode pair has been doubled, but with capacitors that are now half the original value. This reduces the current in each diode by two, allowing the transmitter power to increase by four.

Toward Higher Power T/R Systems

Although the systems presented above can be extended to higher power, they have problems. First, they are single-band designs. Second, the scheme only works well with power amplifiers that are biased off during receive periods. An amplifier biased on will generate noise, with much of it available to the receiver input where it can mask signals. It is often desirable to bias high power amplifiers on, even when used in digital modes, owing to improved stability.[3]

The answer to the T/R problem has been available for many years in the form of high-power PIN diodes. A PIN diode is a structure with the usual P and N semiconductor regions found in most junction diodes, but with an intervening region of *intrinsic* silicon. The intrinsic material has no doping that would cause one conductor type to dominate over another. The PIN diode has the useful property that it can conduct an RF current that is much larger than the dc current that biases the diode on. The intrinsic region also serves the purpose of increasing the breakdown voltage of the structure when it is reverse biased. PIN diodes, specified for RF applications and suitable for high power T/R switching, are available from Microsemi Corp of Watertown, MA, and are discussed in the literature.[4]

Although I have used inexpensive high-voltage rectifier diodes as PIN diode substitutes for RF switching applications, I had never tried to use them to switch large RF currents.[5] The project was "on the list," but still untried. Then I had one of those experiences that are becoming all too rare in Amateur Radio, an on-the-air contact that evolved into a meaningful technical exchange. A CQ on 20-m CW produced a reply from Jim, K5CX. Jim's full break-in, high-power, solid-state transmitter included diode T/R switching *with inexpensive diodes*! The information exchanged in the contact and in later correspondence provided the basis for the antenna switch to be described.

A test circuit is required to evaluate diodes; this is shown in Fig 5. RF chokes and high-voltage ceramic capacitors isolate biasing circuitry from the RF switch. The components shown are suitable for 7- to 14-MHz operation. The diode to be tested is placed in the circuit, bias is attached, and RF is applied. The "on" performance is first

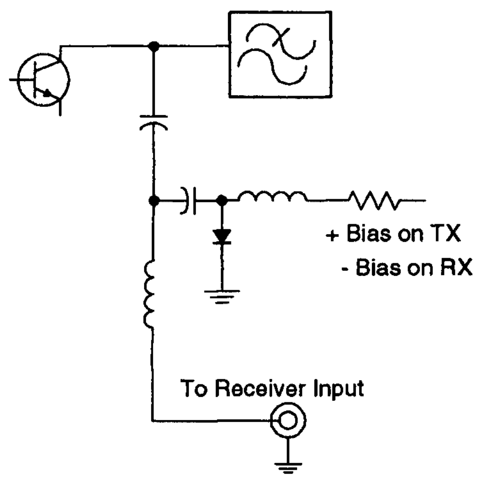

Fig 3—A single shunt diode serves the T/R function if the bias is controlled from external circuitry.

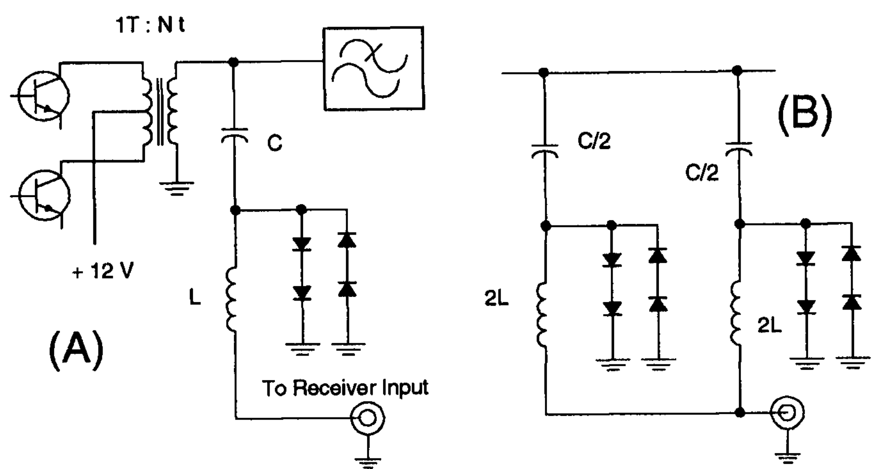

Fig 4—Some variations of the resonant T/R system are shown. A transformer coupled amplifier is used at A. The system in B uses extra parallel reactances, reducing the current in the diodes.

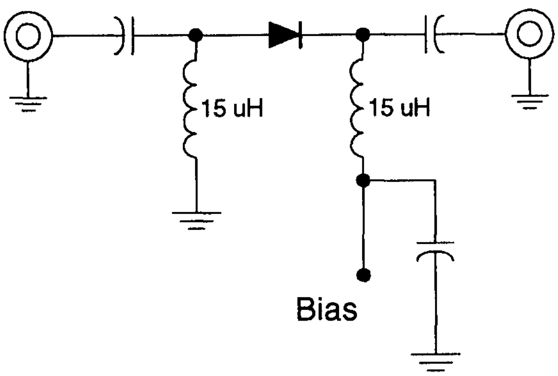

Fig 5—Simple SPST PIN switch. The inductors are molded chokes. All capacitors are 0.02 µF, 1 kV ceramic. There is no difference between input and output for this circuit.

evaluated with an available QRP transmitter. The switch is bypassed and the transmitter output is measured with an oscilloscope and a suitable 50-Ω termination. The RF voltage is measured with as much resolution as possible, for we are looking for small changes. After this calibration, the switch is inserted in the coax line to the termination and the measurement is repeated. The switch performance can be studied as a function of the forward dc current applied to the diode.

After the initial low-power measurements, the process can be repeated with higher power. A high-power termination is required for this measurement.

The next measurement is an evalu-

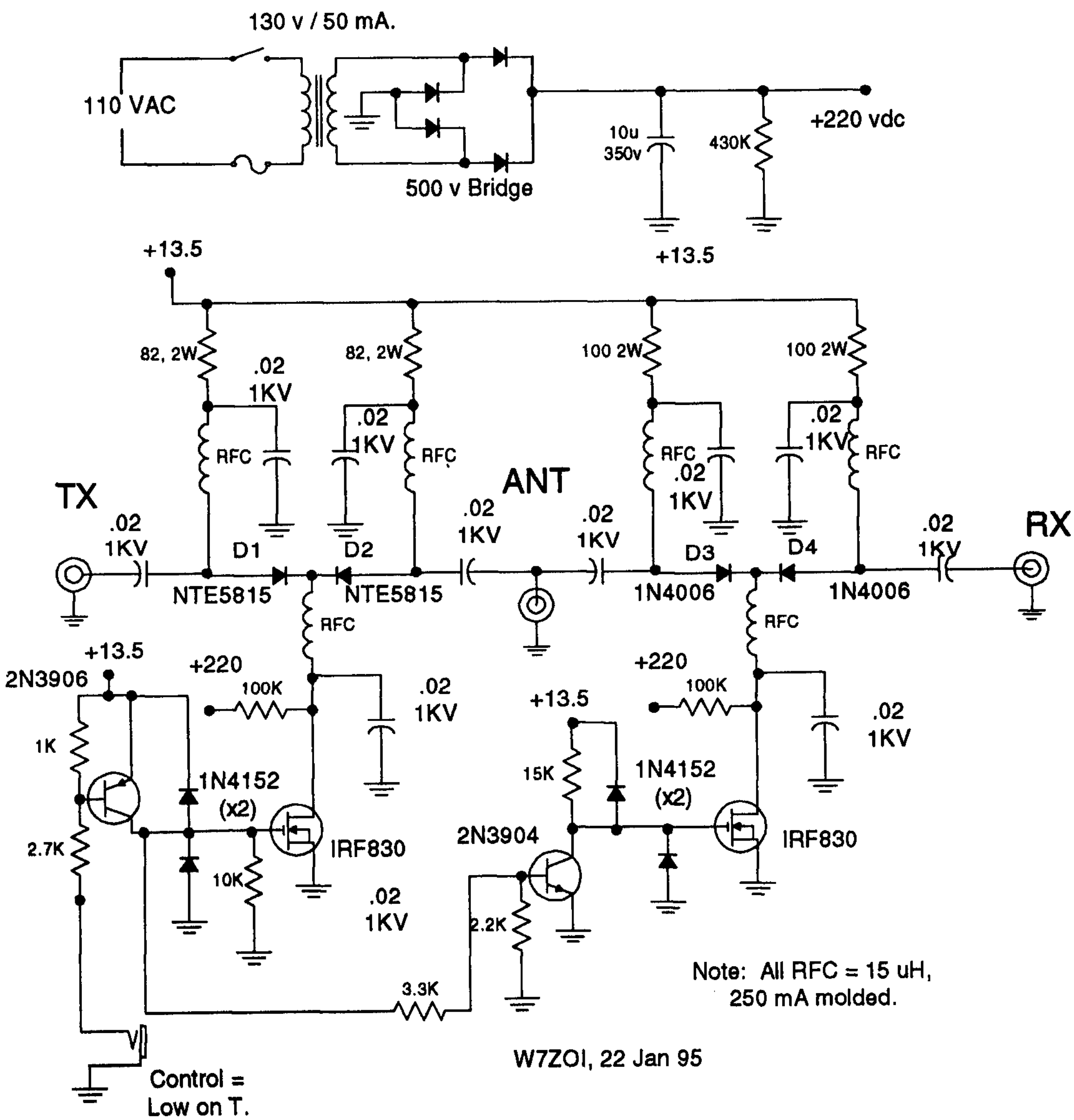

Fig 6—A T/R system suitable for 100-W HF transmitters.

ation of the "off" performance. This is done at low signal levels. A signal generator can be used to provide RF from a well defined source impedance. Alternatively, the QRP transmitter can be used with an attenuator (approximately 10 dB) in its output. I measured output voltage either with a spectrum analyzer or with a 50-Ω terminated HP3400A ac voltmeter.

The switching diodes are characterized by a small resistance when operating with forward bias and by a capacitance when operating with reverse bias. I was not able to measure any forward resistance in any of the diodes I tested; the losses were too low, even at the 100-W level. Also, no temperature rise could be detected in the diodes, even with 100 W flowing.

The series capacitance is related to the insertion loss by

$$C(farad) = \frac{1}{4 \cdot \pi \cdot F \cdot R} \cdot \sqrt{\frac{P}{1-P}}$$

where F is frequency in Hz, R is the source and termination resistance, usually 50 Ω, and P is the power ratio,

$$P = 10^{-IL(dB)/10}$$

P is less than unity.

The diode type recommended by K5CX was the Motorola 6A6. This is a power rectifier specified for I=6 A and reverse voltage of 600 V. The DigiKey catalog lists this diode and several similar devices as manufactured by Diodes, Inc. Jim commented that virtually any of these diodes will provide the needed low on resistance so long as the dc bias current is adequate. I've found that a current of 100 to 200 mA is more than enough for 7-MHz operation at the 100-W level in a 50-Ω system.

The diodes I evaluated included a so-called "equivalent" device, the NTE5815, an industrial rectifier specified at 600 V and 6 A. The 1N4006 diode was also investigated, a device specified at 1 A and 800 V. The NTE5815 appears to be like the 6A6, although it is difficult to tell without more detailed specifications. The best procedure is to test whatever diode is to be used.

The NTE5815 had a capacitance of over 30 pF with a reverse bias of 80 V, with little change in going up to 200 V. The 1N4006, however, had C=3.6 pF. Owing to the small size of the 1N4006 when compared with the monstrous NTE5815, the 1N4006 is only used in the receive path. (In spite of these recommendations, I've encountered no diode failures with 100 W through the 1N4006 diodes I tested. They were surplus without a well-defined manufacturer.)

A friend, K6OLG, sent a few Motorola 1N4007 diodes. These devices, rated at 1 A and 600 V, had a measured C of 2.1 pF at −80 V.

A T/R System for 100-W Transmitters

The final result of these experiments is shown in Fig 6. Owing to the high series capacitance of the NTE5815 diodes, a configuration using two series-connected diodes per switch leg was chosen for the final circuit. All of the components were found in the junk box except for the diodes, and they were inexpensive. High-voltage HEX FETs were used for the dc control. The 1N4152 diodes are included to protect the HEXFETs against transient gate voltages. A junk box power transformer provided a high reverse bias for diodes in the "off" position. If a suitable replacement cannot be found, a pair of back-to-back identical filament transformers could be used. The reverse bias of 220 V is more than enough to keep the diode capacitances low. A forward current of about 150 mA is used to bias the "on" diodes, with the current being "stolen" from the transmitter power supply.

The T/R switch of Fig 6 is normally in the receive mode. It is switched to the TX mode when the control line input is short circuited to ground. The signal to realize this comes from timing circuitry built into the exciter that drives the power amplifier. The same control signal formerly activated a relay.

Several measurements were done with the finished switch. The system provides an isolation of 56 dB between the TX and RX ports when the ANT port is terminated in 50 Ω. If greater isolation was required, a shunt switch at the receive port could be added; this additional diode would be forward biased during transmit periods. The 56-dB value is entirely adequate for the present application, a 100-W transmitter at 7 MHz.

The next measurement was an evaluation of intermodulation distortion in the receive path. No intercept number could be established, for the only IMD observed was attributed to the instrumentation used for the measurement. The measurement did put a lower limit on the switch intercept at +40 dBm. This is beyond the performance of even the best receivers in routine amateur application. It would have been interesting to evaluate the 3rd-order IMD in the transmit path. This was not possible owing to inadequacies in measurement equipment.

RX port isolation will degrade at higher frequencies, encouraging the builder to include the shunt switch mentioned earlier. The RF chokes used are suitable for 7 MHz, but not much lower. The builder interested in the 80- and 160-meter bands should increase the inductance accordingly. L values that become too large may compromise the higher bands through self-resonance effects, so measurements are definitely in order.

Acknowledgments

As mentioned, the basic work that led to the higher power circuits presented was all performed by Jim Miles, K5CX. Many thanks to Jim for sharing his results with us. Thanks also go to Bill Carver, K6OLG/7, for his interests in the circuitry and the methods.

Notes

[1]Lewallen, Roy, W7EL, "An Optimized QRP Transceiver," *QST*, August, 1980.

[2]Personal correspondence, Dave Newkirk, ARRL.

[3]Dye, Norm and Granberg, Helge, *Radio Frequency Transistors: Principles and Practical Applications*, Butterworth-Heinemann, Boston, 1993.

[4]Doherty and Joos, "PIN Diodes Offer High-Power HF-Band Switching," *Microwaves and RF*, December, 1993.

[5]Hayward, Wes, W7ZOI, "Beyond the Dipper," *QST*, May, 1986. A 1N647 was used for band switching in the signal generator described.

Designing the Toroid Transformer for the Directional Coupler

Getting the transformer right is one key to a successful directional coupler design.

By William E. Sabin, WØIYH

This article is a follow-up to a previous *QEX* article which was an in-depth analysis of the two-transformer directional coupler, as illustrated in Fig 1.[1] When the diodes are present, this design is commonly used as a combination SWR meter and RF wattmeter. It is also used as an ALC detector and as a protection device in RF power amplifiers.[2,3] The diode outputs contain amplitude information but no phase information (this is lost in the rectification process). When designed without diodes it is used in network analyzers and many other applications where both amplitude and phase measurements are required. And it is often used to inject a signal into or sample a signal from a circuit for various purposes in a way that creates negligible disturbance to the normal functions of the circuit (Note 3). This contrasts with the behavior of certain other types of bridge circuits.

[1]Notes appear on page 12.

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In this brief article we will discuss some of the properties of the toroidal transformers that are used in these couplers and give a fairly detailed explanation of how they work.

The Transformer

The two transformers of Fig 1 are *conventional* transformers, not *transmission line* transformers. The oscillating magnetic flux in the core is responsible for the power transfer. There are some things that we need to understand about transformers, especially a toroidal type such as shown in Fig 2, where there may be just a single thin wire that passes through the center of the hole but does not wrap tightly around the core. How can this be an efficient transformer? The following discussion describes what happens and adheres exactly to the standard teachings in textbooks on electromagnetics, in this case Notes 4 and 5. I hope that this review will be interesting to those who may not have thought much recently about this subject.

1. In Fig 2 the ac current flowing through the primary winding on the left produces a magnetizing force H parallel to the outside surface of the core in accordance with Ampere's Law.
2. There is a *boundary relation* that says that this parallel magnetizing force is *continuous* across the boundary of the core. It has the same value inside the core that it has just outside the core. Incidentally, we assume that there is no *permanent* magnetization of the core.

3. The core is assumed to have a high value of permeability μ. This means that the magnetic flux density $B = \mu H$ inside the core is large. The total flux in the core is the flux density B multiplied by the cross-sectional area A of the core.
4. In this ideal example there is *no* flux density B *outside* the core except for a tiny amount in the air space between the coil and the core that is due to the H that is found there in Fig 2. B is virtually *all* confined to the interior of the core, especially in view of the high value of μ.
5. The secondary winding at the right passes through the center of the hole and connects to an ac voltmeter that has an infinite resistance (no current flows). The wire does not touch the core but is spaced away from it as shown. The wire and its voltmeter constitute a complete loop around the core in the manner shown.
6. We emphasize that *none* of the flux in the core comes into contact with the wire. In particular, there is *no* magnetic flux in the hole of this perfect toroid (if the voltmeter current is zero as we have initially assumed; see later discussion).
7. According to one of J. C. Maxwell's equations, based on Faraday's law of induction, the voltage read by the meter is equal to the *rate of change of the flux inside the core*. This action occurs despite the fact that the flux inside the core *never* comes into contact with the wire itself. The only thing that is required is a completed loop that surrounds the flux that is inside the core. The loop can be of any reasonable size.
8. The explanation for this voltage is that, according to Maxwell, the oscillating flux in the core establishes an oscillating *electric* field E that encircles the *outside* of the core as shown in Fig 2. *If* (note the word "if") the wire is present the E field *induces* a voltage V across the wire terminals. For N turns of wire, the voltage is multiplied by N.
9. We should be more precise about the word "induce." A second boundary relation says that if an electric field E is parallel (tangential) to the boundary between two different media, the field is the same on both sides of the boundary. If a free-space E exists next to the secondary wire in Fig 2 it also exists inside the wire (Note 5).
10. Why does the loop have to *enclose* the magnetic flux? Consider the dotted-line open-circuit loop in Fig 2, which does not enclose the flux. The voltage induced in the inner wire and the voltage induced in the outer wire cancel each other at the terminals.
11. Return to the main loop. At a greater distance from the core the E field is still present but it is weaker. In Fig 2 we can see that this E field is most intense in the hole of the toroid and weakens as we move along the wire away from the hole. But the total voltage induced in the wire loop is the same, regardless of the length of the path. Nearly all of the voltage occurs close to the toroid.
12. In practice there may be a small amount of leakage flux in the neighborhood of the core, and some of it may come into contact with the wire, but it is not nearly enough to account for the almost perfect coupling that exists between primary and secondary.
13. The effect due to Faraday/Maxwell also affects the primary winding. The same E field induces a "back" voltage in the primary that opposes the generator voltage and reduces the primary current. This is the well known *counter-emf.*
14. Ampere's Law also operates at the secondary if a current is allowed to flow in the voltmeter. This current sets up an H field around this wire, in the hole of the toroid, that induces a flux into the core. This flux *opposes* the increase or decrease of the flux that was created by the generator in Fig 2. This is Lenz's law.

We see that a careful study of the Maxwell equation derived from

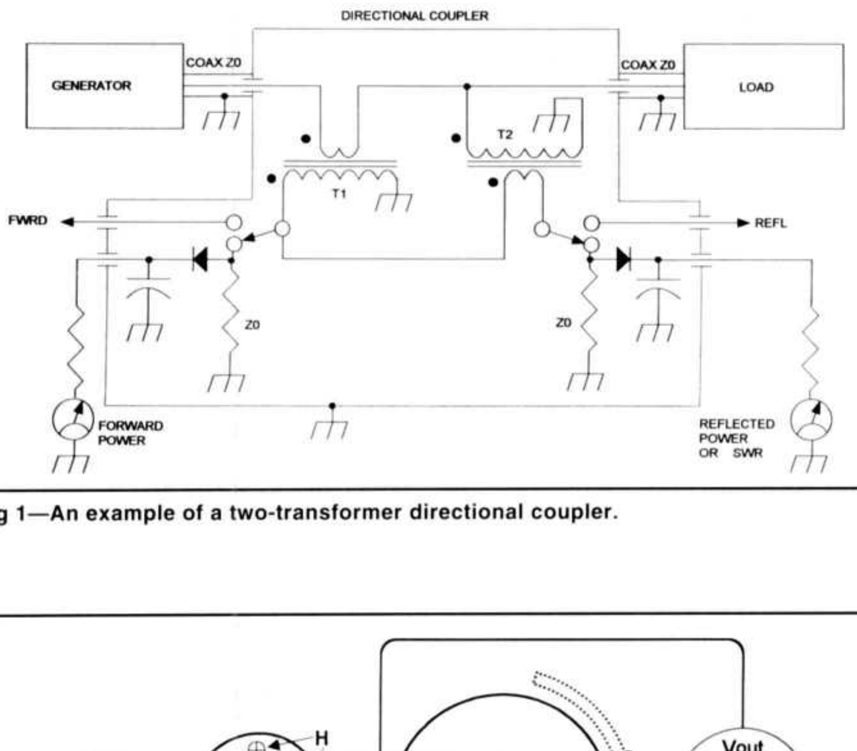

Fig 1—An example of a two-transformer directional coupler.

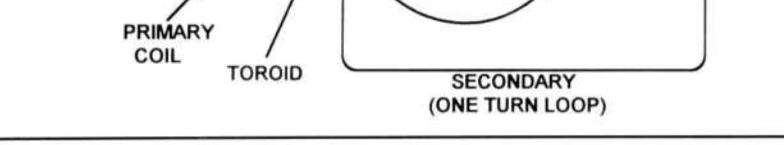

Fig 2—The transformer action of a toroid. The perspective is from the side of the toroid, with the shaded areas representing the cross section of the toroid core.

Faraday's law confirms that there is *no* requirement for the secondary wire to be in contact with the flux, only that the path of the E field and the conductors *enclose* the flux. All conventional transformers operate on the principles outlined in these steps. And that is why the wire through the center that appears to be ineffectual as a transformer winding is, in fact, very effective indeed. The author has measured coefficients of coupling of at least 0.98 for transformers like those in Fig 1.

Also, note that the straight wire through the hole can be a one-turn primary winding just as easily as a secondary winding. There is a principle of reciprocity that guarantees this. The H field that surrounds this wire couples into the core very efficiently. But as the core, especially the center hole, gets larger there can be an increase in leakage effects, which is to be avoided in either of the situations that we have described.

The Faraday Shield

Fig 3 illustrates another aspect of the toroid transformer design. At high frequencies, capacitive coupling between the toroid coil winding and the center wire can induce undesired voltages (in either direction) which are not in the same phase as the desired couplings caused by the H field (caused by current in the center lead) and the E field (caused by flux change in the toroid). These can cause errors in the operation of the directional coupler. The shield, grounded at one end, prevents this capacitive coupling. And yet, the normal operation of the E field coupling from the toroid to the center lead and H field coupling from center lead to toroid are not impaired. The reasons are as follows:

1. Let the toroid winding be the primary, as in Fig 2. The E field created by the toroid, as shown in Fig 2, induces a voltage along the length of the shield. This voltage can be (and has been) measured with an RF voltmeter from ground to the open end of the shield. Because only one end is grounded, no current flows along the length of this shield.
2. The second boundary relation mentioned in Sec 9, above, also operates in this situation (the component of an *electric* field that is tangential to the surface of one medium is continuous across the boundary of an adjacent medium). An exception *can* occur if the second medium is a conductor. The shield material in this case is a very good conductor, but because it is open-ended as shown in Fig 3 and not conducting any current, it qualifies as a "non-conductor." That is, the E field generated by the toroid exists unimpaired on the outside of the shield, in the shield material, inside the shield and in the dielectric. In other words, the shield is *transparent* to the E field. But if both ends of the shield are grounded the current flow on the shield creates an H field whose E field cancels the E field that is generated by the toroid, effectively "shorting out" the E field.
3. Now assume that the center conductor is acting as a primary winding. The current in the center conductor generates an H field that is not interrupted by the Faraday shield, again because there is no current flow in the shield. The boundary relation mentioned in part 2 of the previous section applies (the tangential H field is continuous across a boundary). Compare this with the usual coaxial cable situation, where the H field generated by the current flow in the center conductor and the H field caused by the *equal-valued* current in the braid, which flows in the *opposite* direction, cancel each other on the *outside* of the coax. Again, the Faraday shield is transparent.

The Faraday shield makes a difference at the upper end of the frequency range and extends this range significantly. Because the coax braid is interrupted by the Faraday shield, the metal case of the coupler must provide the coax return path. The characteristic impedance Z_0 of this case is seldom the same as the Z_0 of the coax. Microstrip techniques are often used at the higher frequencies to fix this problem.

The interrelationships between voltage, current, magnetic flux and electric fields, how they *create* each other as described briefly in these two sections and spelled out succinctly and precisely by Ampere's Law and Faraday's Law (and also how these two laws *interact* with each other), together with a few boundary relations, are the keys to understanding a wide variety of electromagnetic devices and principles.

Transformer Design

The details of the transformer design have been, in the past, to some extent a matter of experimentation, but there are several engineering guidelines that will help a great deal. We will discuss some of these considerations.

In Fig 1, the high impedance winding of T2 is connected across the OUTPUT terminal and encounters the full output voltage level. Normally (meaning at a low SWR) T1's high-impedance winding has only $1/N$ as much voltage and is much less vulnerable. Nevertheless, we make T1 and T2 identical, to get maximum symmetry.

The toroid core's flux density (in gauss) in T2 is given by the formula:

$$B_{MAX} = \frac{V_{PK} \cdot 10^2}{4.44 \cdot A_e \cdot N \cdot F}$$

where B_{MAX} is peak flux density, V_{PK} is the peak value of the coupler's ac output voltage, A_e is the cross-section area in cm^2 and F is in MHz.[6] This formula is derived from the Maxwell/Faraday

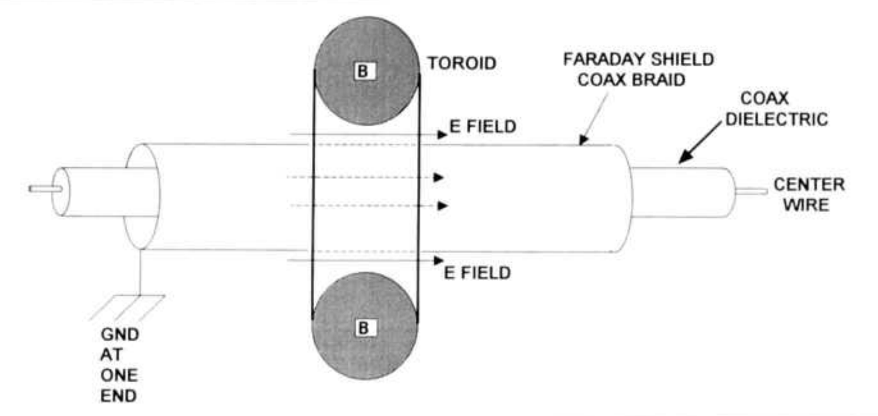

Fig 3—The Faraday shield.

equation mentioned previously.

Amidon Associates suggests the maximum values of B_{MAX} shown in Table 1 to avoid overheating of *either* iron *or* ferrite cores at high power and at HF frequencies. These numbers suggest the increasing losses and heating at the higher frequencies due to hysteresis, eddy currents and dielectric losses. If the primary is a one-turn, straight-through wire, then for a desired value of N (the coupling factor) the core cross section is chosen to satisfy the values of Table 1 at the maximum value of V_{PK}. Note that the formula does not require the diameter (path length) of the core. If necessary, we use two or three stacked standard-size cores to get a good value of cross-section A_e. We can also increase the coupling factor, N. This reduces the volts-per-turn and therefore reduces the flux density.

Consider an example: 1000 W (worst case, steady carrier) into 50 Ω equates to 224 V_{rms}=316 V_{pk}. Let N=40, F=3.5 to 28 MHz and use the B_{MAX} value for each amateur band from Table 1. Solve the equation for A_e=0.45 cm^2 at 7 MHz (worst case, 3.5 to 28 MHz). A T-80-2 (μ=10) core has an A_e=0.242 cm^2, so two of these cores will fit the Amidon guidelines for core heating. These cores, wound with #22 wire, fit nicely over an RG-213 or RG-8 braid (Faraday shield) with a teflon tape wrap. The coil has a measured Q of 250 at 2.5 MHz and a measured inductance of 19 μH (418 Ω at 3.5 MHz). For CW (50% duty cycle) or SSB at 1500-W PEP this transformer would be more than adequate. Also, the flux density is far below the saturation level of type-2 powdered-iron material and should have no IMD problems. Consultation with an Amidon applications engineer confirms the reasoning in this example.

An exception to this high power example can occur if the OUTPUT terminal is not properly loaded, in which case the RF voltage across T2 (open-circuit load), or across T1 (short-circuit load), can become excessive. This is an unfortunate characteristic of this type of coupler that must be avoided. Fast acting high-reflected-power detection, derived from the REFLECTED terminal, or operator caution, is advisable to prevent damage to the coupler. This problem is especially serious if the output impedance of the coax that is connected to the INPUT terminal is very high (current source) or very low (voltage source). Initial tuneup of the load at greatly reduced power is always a safe procedure. But if the output impedance of the coax and the power source are close to Z_0 then the voltage rises across T1 and T2 are limited to much less destructive levels. Some RF power amplifiers use negative feedback to adjust their dynamic output resistance to the range of the Z_0 value (see Note 3, Chapter 14). The flux density levels mentioned in Table 1 are conservative, so voltage increases of, say, 3 times or so should be acceptable—at least for a few minutes.

Table 1—Recommended Toroid-Core Peak Flux Densities

Frequency	B_{max}
1 MHz	500 gauss
3.5 MHz	150 gauss
7 MHz	57 gauss
14 MHz	42 gauss
21 MHz	36 gauss
28 MHz	30 gauss

The insulation of the toroid windings to ground, via the Faraday shield or via the core from the "hot" end of T1 or T2 is an issue at higher power and higher SWR levels. These voltages can become large and the windings can arc over, especially for certain types of powdered-iron cores that might have a low value of resistivity. The Faraday shield must be well insulated with teflon tape. Powdered-iron cores are painted at the factory to protect the core and have a rounded and smooth (tumbled) surface. High-grade enameled wire should be used. The author has measured >2 × 10^7 Ω (the limit of my ohmmeter) across the diameter of a T-80-2 core (paint removed), so this core material and other low μ types should not have any resistivity problems. A layer of thin teflon tape on the core(s) does not seem to be crucial but might add some reassurance. A thick coat of polystyrene Q-dope over the finished coil is always a good idea. Place a 20° gap between the *start* and *finish* of the winding.

The copper loss due to the resistance of the wire is difficult to separate from other losses in the toroid transformer. For a toroid it is probably best to use the largest wire size that is reasonable. We then find that for a certain maximum power level into the correct load resistance Z_0, the voltage across T2 is constant over frequency and so the current through the winding is greatest at the lowest operating frequency, where the reactance is minimum. At higher frequencies the reactance increases linearly and, as an approximation, the resistance due to skin effect increases as the square root of frequency. So under these constraints the I^{2}R loss in the copper *tends* to become smaller at the higher frequencies. The behavior of skin effect in a coil over frequency is complicated, though, so this is only a guideline.[7] The "touchy-feely" approach at the various frequency bands is always a good idea.

The open-circuit impedance of T2 is in parallel with the load impedance Z_L at the OUTPUT terminal. T1 loads the FORWARD terminal. Figs 4 and 5 show R_P and X_P for one example, two stacked FT-37-61 (μ=125) ferrite cores with 10 turns of #28 wire. Above 50 MHz the reactance becomes capacitive. This loading degrades the directivity of the coupler a little, especially at high frequencies. These errors and others can be reduced, if we wish, by terminating the REFLECTED terminal *internally* with a test-selected Z_0 resistor. A small trimmer capacitance (1 to 5 pF) in parallel is almost always found to be helpful also. Using the formula, this pair of cores would be adequate for about 7.5 to 10 W. The terminating resistor could be 0.25 W (>10/N^2).

This procedure minimizes the transfer from the OUTPUT port to the FORWARD port when the signal is applied to the OUTPUT port. The FORWARD port can then be used to read forward or reflected signal by reversing the connection of input and output, and the directivity will be at its best. Two internally terminated couplers back to back, as shown in Note 1, Fig 3, is an excellent option for instrumentation setups.

Another option for a dual output coupler like that of Fig 1 is to compensate both the FORWARD and the REFLECTED ports. Switch the input source and the output load back and forth and fine-tune the FORWARD port, then the REFLECTED port, for a uniformly low output across the frequency range. The coupler then has pretty good directivity in both directions. This can be important in certain applications. When fine-tuning the REFLECTED port, the FORWARD port and the OUTPUT port should both be accurately terminated (and vice versa). In all situations, beware of incorrect values of coax Z_0.

In many uses of the directional coupler, only a single RF signal of low amplitude is encountered, so the cores

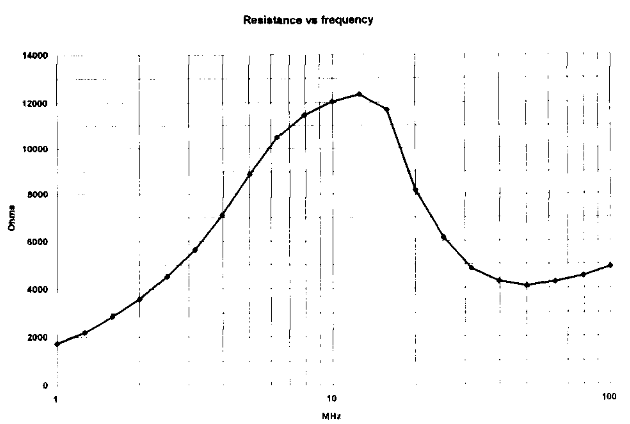

Fig 4—The resistive part of a typical toroid coil's impedance.

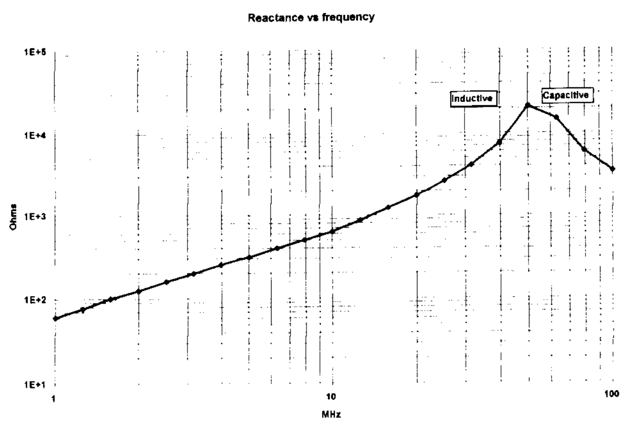

Fig 5—The inductive reactance part of a typical toroid coil's impedance.

can be very small, high-permeability ferrites, and a very large frequency range can be achieved. In signal-processing applications where multiple signals are present, usually at higher power levels, intermodulation distortion, heating and core saturation are problems that increase the needed size of the core and the length of the wire on the winding. This affects the high frequency response. Therefore, the proper way to design the transformer is to customize it for a specific frequency range. Trying to accomplish too much in one coupler can be a frustrating experience. At higher frequencies a microstrip layout is very desirable but does not seem necessary below 200 MHz *if* all dimensions are made as small as possible. Large values of N (too much wire) can affect the high frequency response. Ten turns (for a 20-dB coupler) has been a good compromise for low power, 1.0 to 150-MHz couplers, in my experience.

Another source of error involves the diameter of the core and therefore the length of the winding. The E field in the hole (Fig 1) is the result of contributions from each turn of the coil. These contributions add vectorially. At high frequency there is a small phase delay from one end of the coil to the other. A kind of transmission-line (or delay-line) effect, if you will. This causes a phase error in the net E field and therefore a degradation of directivity. The compensation techniques try to reduce this error.

When building and adjusting a coupler for best directivity, a spectrum analyzer with a tracking generator is a valuable resource. If it is not available (the usual case) a low-cost signal source and a receiver that tunes the desired frequency range can be used. The process is slower, but for a limited frequency range, not difficult. A wideband oscilloscope may be available. The main problems occur at the low frequency end (not enough core) and at the high end (poor symmetry) where phasing errors cause most of the problems (recall that the coupler is a phase-sensitive device).

Computer simulations using the *ARRL Radio Designer* (ARD) program show that capacitive coupling of only a fraction of a pF from the toroid coil to the through-wire causes significant loss of directivity at the upper frequencies. The Faraday shield appears to be well justified, as confirmed experimentally.

When fine-tuning for maximum directivity, a high quality termination is needed for the main signal path. This should match the Z_0 of the coax, which can be 50 Ω, or 52 Ω, etc. A good goal for directivity is 35 to 40 dB. Often, small changes in the positions of the transformers on the through-wires will pick up a few dB at the high frequency end. Building the coupler in an enclosed aluminum box with coax connectors has produced consistently better directivity, in my experience, because it does a good job of diverting ground-path input-output currents away from sensitive areas. It also helps to complete the coaxial cable return path. A microstrip ground plane is also helpful. The final adjustment for directivity should be made with the case closed up. A small access hole to the trimmer cap is desirable.

If a high level of directivity is not obtained, a perfect load will not produce a very low output at the REFLECTED terminal. A transmatch or other tunable load might end up in a slightly mistuned condition in order to minimize the REFLECTED output. The same can occur if the coax doesn't have the right Z_0.

Conclusion

The author appreciates the helpful suggestions on electromagnetics by Daniel M. Mitchell, WØPWN, Rockwell Co, Cedar Rapids, Iowa, and for discussions about cores, the Applications Department, Amidon Associates. For some additional material on this subject, please see Chapter 22 of the 1995 edition of the *ARRL Handbook.*

Notes

[1]Sabin, W. E., WØIYH, "The Lumped-Element Directional Coupler", *QEX*, March 1995.
[2]Blocksome, R., KØDAS, "An HF 50-W Linear Amplifier," *ARRL Handbook*, 1995 edition, pp 17.93-17.97.
[3]Sabin, W. E., WØIYH, Schoenike, E. O. and others, *Single-Sideband Systems and Circuits*, McGraw-Hill, 1987, Chapters 12, 13 and 16. Also second edition, 1995.
[4]Kraus, J., W8JK, *Electromagnetics*, 4th edition, McGraw-Hill, 1992, Chapter 10.
[5]Harnwell, G. P., *Principles of Electricity and Electromagnetism*, McGraw-Hill, 1938, p 313.
[6]*Iron-Powder and Ferrite Coil Forms Catalog, 1992,* Amidon Associates, 3122 Alpine Avenue, Santa Ana, CA 92704, 714-850-4660.
[7]Terman, F. E., *Radio Engineering*, McGraw-Hill, 1937, pp 35-41.

Graphic Method for Calculating Z

Reprinted from February 1995 Radio Communication.

By A. J. Harwood, G4HHZ

This article arose out of a simple inquiry, which upon reflection (and *there's* an apt phrase when writing about transmission lines) proved quite difficult to answer without being too mathematical whilst still giving an adequate explanation.

The question was prompted by a discussion on the measurement of impedance using a noise bridge and was: "If a complex impedance (consisting of both a resistive and reactive component) is measured at the end of a transmission line, how can the standing wave ratio on the line and the impedance at the other points on the line be calculated?" The quick answer is that you simply plot the impedance on a Smith Chart.

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The Circle Chart

As the inquirer was neither familiar with, nor had, a Smith Chart, I showed how to use a different graphical solution, which also gives a lot of information of use in allied areas such as the design of impedance bridges and aerial tuning units. What this article aims to do is to explain how this method, a simple graphical transmission line calculator known as the Cartesian Circle Diagram, can be constructed with a ruler, compasses and protractor and used in conjunction with a pocket calculator to solve such problems.

Although an understanding of the mathematics is not essential to solving transmission line problems it is helpful to understand exactly what is meant by impedance, perhaps by reading G4FZH's recent article explaining how impedance changes at different points along a transmission line.[1] A full explanation of the latter was given by G3HRH, see Note 2, and is summarized in the "Why the Impedance Varies Along a Transmission Line" sidebar.

It is the usual practice when dealing with transmission lines to work in terms of the normalized impedance, that is, all impedances are divided by the characteristic impedance, Z_0, of the line. For instance a resistance of 75 Ω connected across a 50-Ω line has a normalized value of 75/50 or 1.5. This article considers the case of series connected impedances, and those quoted will be normalized.

One of the easiest ways of understanding a subject that can be expressed as a complicated mathematical expression is by plotting it as a graph, and this is particularly true when applied to the transmission line problem. Taking a lossless line that is

[1]Notes appear on page 16.

over a half wavelength long and terminated in a normalized impedance, R_t in series with X_t, the resistive and reactive components R and X of the impedance arising as the measurement point is moved along the line are plotted. Resistance is on the horizontal scale, (1 representing the characteristic impedance of the line) and the reactance is on the vertical. By convention, inductive reactance is positive and capacitive reactance is negative. This results in a circle as shown in Fig 1. The complete circle represents the change of impedance occurring over a half wavelength of line for the particular load impedance. Moving around the circle in an anticlockwise direction corresponds to moving from the measurement point towards the load and vice versa. Of particular interest are the points which are purely resistive since these are equal in value to the SWR(S) and its reciprocal ($1/S$) and are a quarter wavelength apart. As an example, for an SWR of 2 the circle passes through 2 and 0.5 on the line of zero reactance and so has a diameter of 1.5 with its center at the point (2+0.5)/2=1.25. In general terms the circle will have a diameter of (S–1/S) with the center at (S+1/S)/2. We are well on the way to answering the original question! To find the complete answer we need to consider Fig 2.

The circle passing through the point R, X also passes through the point R, –X and two other points where the reactance is +X and –X, but the resistance has a different value, r. Point r, –X lies at the other end of a diameter of the circle to R, X. R, r and X are connected by the expression:

$$Rr = 1 + X^2 \text{ or } r = \frac{1+X^2}{R} \qquad \text{Eq 1}$$

so as R and X are known, r can be calculated. The center point of the circle is at R_o on the zero reactance line. Here the reactance is at its maximum value, X_{max}. R_o has the average value of R and r, so can also be calculated since:

$$R_c = \frac{R+r}{2} = \frac{R+\frac{1+X^2}{R}}{2} = \frac{R^2+X^2+1}{2R} \qquad \text{Eq 2}$$

The circle can now be drawn with center at R_o, 0 and a radius from this point to R, X enabling S to be found.

X_{max} can also be calculated since at this point the values of R and r coincide, both being R_o, hence:

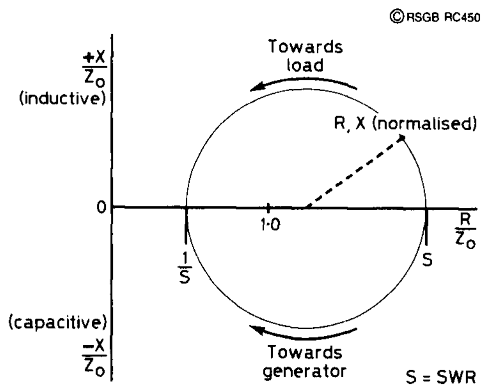


Fig 1—Plotting all values of impedance along a line results in a circle. If the impedance is normalized the circle passes the points *S* and *1/S* on the *R* axis. From *S* to *1/S* corresponds to a quarter wavelength on the transmission line.

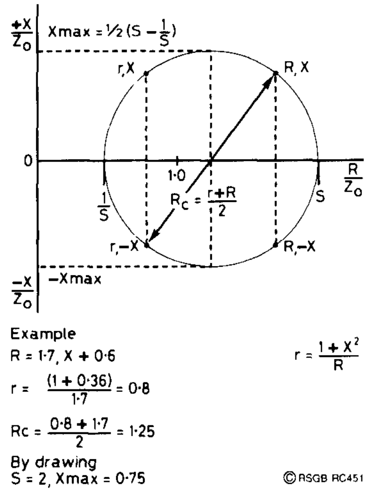


Fig 2—The SWR can be found from the values of *R*, *X* by calculating $r = 1 + X^2/R$ and $R_c = R = r/2$. The SWR circle can then be drawn.

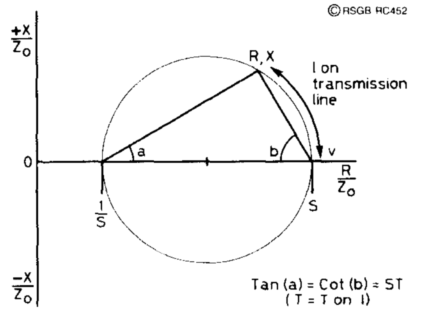


Fig 3—The distance on the transmission line between *R*, *X* and *S* can be determined from angle *a* or *b*.

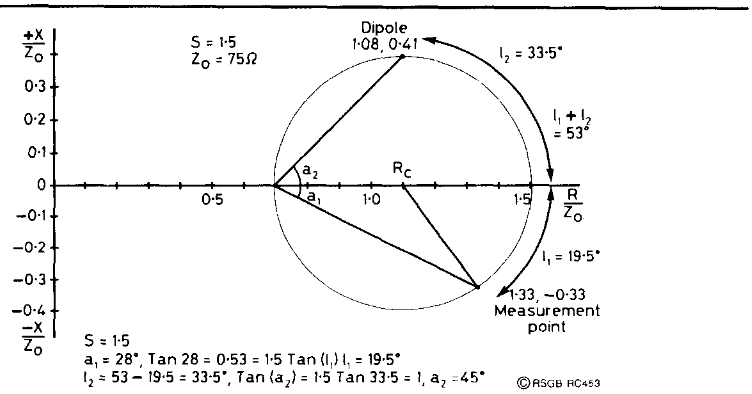


Fig 4—The circle diagram can be used to calculate the impedance at the input to a dipole from a measurement made at the transmitter end of the feeder.

$$R_o^2 = 1 + X_{max}^2 \text{ or } X_{max}^2 = R_o^2 - 1$$

$$\text{As } R_o = \frac{\left(S+\frac{1}{2}\right)}{2} \text{ and } X_{max} = \frac{\left(S-\frac{1}{2}\right)}{2S} \qquad \text{Eq 3}$$

S can also be determined simply by adding R_o to X_{max}.

To summarize, we can find a second point on the circle from Eq 1 and its center from Eq 2. The SWR can then be found either by drawing or by using Eq 3.

Practical Example

At this point, perhaps it would be useful to take an example and see how it works out in practice. Like many amateurs I use a G5RV multiband dipole and have measured the impedance on all bands at the junction of the balanced open wire feeder and the 75-Ω feeder into the shack. On 14.2 MHz the equivalent series impedance is 88.5-Ω resistance and 8.43-Ω inductive reactance, which normalized to 75 Ω is 1.18 resistive and +0.1124 reactive.

Calculating R_o (from Eq 2) gives a value of $(1.18^2+0.1124^2+1)/(2 \times 1.18)$,

Why the Impedance Varies Along a Transmission Line

Energy supplied by a generator of V_f volts and I_f amps to a lossless transmission line of characteristic impedance Z_o is transported by means of an electromagnetic wave with electric field E_f and magnetic field H_f to the load as shown in the diagram. If the load is purely resistive and equal to the line's characteristic impedance Z_i then all the energy fed to the line is dissipated in the load and the input impedance is surely resistive and equal to the load: the load is properly matched to the line. For all other values of terminating impedance a portion of the energy fed to the line is returned towards the input by a reflected wave. The magnitudes of the electric and magnetic components of the reflected wave are related to that of the forward field, E_f and H_f, by the reflection coefficient, p, which depends on the load and Z_o. If p is 10% then the reflected wave has an amplitude of 10% of the forward. At any point on the line the phases of the two fields of the reflected wave relative to those of the forward are determined by the load, the frequency and the distance from the termination.

Since the forward and reflected waves are traveling in opposite directions, at those points where the electric fields are in phase the magnetic fields are in antiphase. Here the impedance is purely resistive and at a maximum value, R_{max}, as is the voltage, V_{max}, across the line, which is $(1+p)$ times the forward voltage, V_f. The current has a minimum value, I_{min}, of $(1-p)$ times the forward current I_f. The resistance is thus $Z_o\,(1+p)/(1-p)$. Similarly, a quarter of a wavelength away the magnetic fields add and the electric fields subtract, with the voltage being at a minimum, V_{min}, of $(1-p)\,V_f$ and the current at a maximum, I_{max}, of $(1+p)I_f$. Here the resistance, R_{min}, is $Z_o(1-p)/(1+p)$. The SWR, S, is the ratio of the maximum to minimum voltage so has a value of $(1+p)/(1-p)$ and is related to the maximum and minimum resistance values and Z_o by:

$$S = \frac{1+p}{1-p} = \frac{V_{max}}{V_{min}} = \frac{R_{max}}{Z_o} = \frac{Z_o}{R_{min}}$$

so

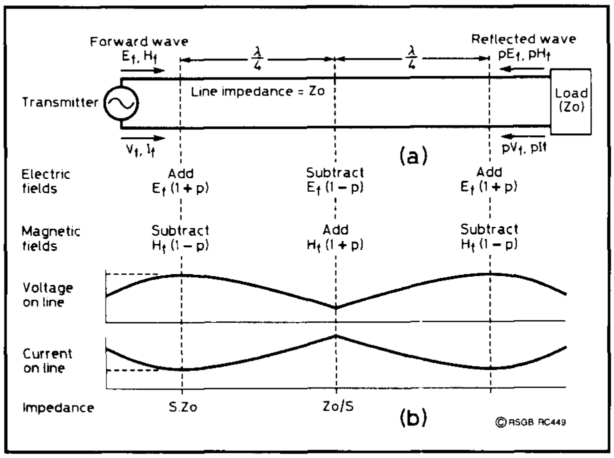
$$R_{max} = SZ_o \text{ and } R_{min} = \frac{Z_o}{S} \text{ and } R_{max} \bullet R_{min} = Z_o^{\,2}$$

At any point on the line the voltage depends on the vector sum of the forward and reflected electric fields, and the current to that of the corresponding magnetic fields. The impedance is given by the ratio of voltage to current and, for a lossless line, is given by the equation:

$$Z_s = \frac{Z_o\left(Z_r + jZ_o \tan(L)\right)}{Z_o + jZ_s \tan(L)}$$

where

Z_o is the characteristic impedance of the line
I_s is the complex input impedance



On an unmatched transmission line a portion (p) of the forward power ($V_f \times I_f$) is reflected at the load to give reflected power of $pV_f \times pI$. This causes the impedance along the line to vary and gives rise to a standing wave of SWR$(1+p)/(1-p)$.

Z_r is the terminating impedance
d is the distance from load to measuring point
L is the electrical angle corresponding to d and can be expressed either in radians when $L=2\pi d/\lambda$ or in degrees when $L=360d/\lambda$.

T will be used subsequently for the expression $\tan(L)$. For a normalized input impedance, Z_i, and terminating impedance, Z_t, the equation then becomes

$$Z_i = \frac{Z_t + jT}{1 + jZ_t T}$$

Take the case of a line terminated in a normalized impedance $Z_t = R_t + jX_t$ (ie, a resistance R_t in series with a reactance X_t). Substituting these values in the above equation and separating the real and imaginary parts shows that the input impedance Z_i consists of two components, a resistance:

$$R_i \frac{R_t\left(1+T^2\right)}{\left(1-X_tT\right)^2 + \left(TR_t\right)^2}$$

in series with a reactance:

$$X_i = \frac{(X_tT)(1-X_tT)^2 - TR_t^2}{(1-X_tT)+(TR_t)^2}$$

Plotting these values of R_1 and X_1 for different values of L results in the circle diagram.

which is 1.019; X_{max} (from Eq 3) is the square root of $(1.0192^2 - 1)$ or 0.196. Adding R_o and X_{max} gives the value of S as 1.215, which is quite a good SWR to be working with. What though of the SWR on the open-wire feeder which, for my G5RV, has a Z_o calculated from its dimension of 620 Ω? Normalizing the measured values to this gives 0.1427-Ω resistive and 0.014-Ω reactive. R_o works out at 3.576 and X_{max} as 3.433, giving an SWR of 7:1 on the open wire feeder. A high SWR does not necessarily mean an inefficient aerial system!

The circle diagram is a graph of all values of impedance existing on the transmission line for a given value of SWR and, although moving a given number of electrical degrees along the line does not correspond to moving twice the same distance around the circle (as in the case for the Smith Chart), the diagram can be used to calculate the impedance at any point once the relationship between a point on the circle and its equivalent position on the transmission line is known. For instance, it is often useful to know the impedance a quarter of a wavelength along the line from a point where the impedance is resistance R in series with reactance X. Here the values of the resistance and reactance are given by

$$\frac{R}{R^2 + X^2} \text{ and } \frac{X}{R^2 + X^2} \qquad \text{Eq 4}$$

Alternatively, they can be found graphically by drawing a line from R, X through the point $1/R_o$, ie, $2/(S+1/S)$ on the zero reactance line. This line intercepts the circle at the point corresponding to a quarter wavelength along the line from R, X.

To find the impedance at any point, one of the two purely resistive points where the normalized impedance is S or $1/S$ is taken as a reference. The distance from the measuring point to that where the resistance is S, as shown in Fig 4, can be calculated from

$$\tan(a) = \cot(b) = ST$$

where

$$T = \tan(L) \text{ and } L + 360d/\lambda \qquad \text{Eq 5}$$

as in the sidebar.

We now have a graphical method whereby the impedance at any point on the line can be calculated if its characteristic impedance and the impedance at one point are known. For instance, the impedance at the input to a dipole can be calculated from a measurement at the bottom end of the feeder.

To illustrate this, consider the case of an 80-m dipole fed by 75-Ω twin balanced feeder 12 meters long, corresponding to 53°, or 0.146 wavelengths at 3.65 MHz. The impedance measured at the transmitter end is 100 Ω in series with −23-Ω capacitive reactance giving normalized values of 1.33 resistive and −0.33 reactive. Calculating the SWR gives 1.5, quite good for a dipole, but can it be improved by adjusting the dipole length?

To find out, the impedance at the dipole must be determined (Fig 4). Plotting the impedance on the circle diagram shows that the angle a_1 is 28°; the equivalent electrical angle L_1 on the feeder is found using Eq 5:

$\tan(a_1) = S\tan(L_1)$ so $\tan(28°) = 1.5\tan(L_1)$ giving L_1 as 19.5°. The total feeder length is 53°, and to get to the dipole impedance a further Eq 1 calculation is required for angle L_2 of 53−19.5=19.5° to get a_2. Here

$\tan(a_2) = 1.5\tan(33.5)$ whence $a_2 = 45°$

The line from $1/S$ at 45° cuts the circle at 1.08, +0.41, corresponding to a dipole impedance of 81 Ω in series with 31-Ω inductive reactance. The dipole is too long, and judicious pruning should reduce the standing wave ratio.

Notes

[1]Smith, Clive, G4FZH, "Circuit Concepts Explained," *Radio Communication*, Volume 69, No. 11, November 1993.

[2]Hills, R.C., G3HRH, "Some Reflections on Standing Waves," *RSGB Bulletin*, Vol 40, No. 1, January 1964.

RF

By Zack Lau, KH6CP/1

Designing VHF+ Yagis

Yagis are ubiquitous at VHF and above, so let's discuss some of the considerations to take into account when designing them. Of course, if you are just looking for duplicable designs, there are some excellent ones by Steve Powlishen, K1FO, in both *The ARRL Handbook* and *The ARRL Antenna Book* for 144, 222, and 432/446 MHz. He even has smaller 432-MHz antennas in *The ARRL Antenna Compendium, Volume III*. But his designs aren't always what *I* need for my application, so I need to come up with my own designs. You may want to do so, too. Of course, I strongly recommend reading Steve's material, as well as that by Günter Hoch, DL6WU, in *The ARRL UHF/Microwave Experimenter's Manual*.

Confirming Modeling Software

The first area of concern is suitable software to do the analysis of the design—I don't know anyone who can do the analysis by hand with pencil and paper. *MININEC* is known to have a frequency-offset problem at VHF: it will design a Yagi all right, but you'll often find it necessary to shorten all the elements, even when using a nonconductive boom. (I've modified one of those wire-stripper/screw-cutter tools for the purpose of clipping the ends off of Yagi elements. I just drilled out the hole that's threaded for 10-24 screws using a 3/16-inch cobalt steel bit.) Fortunately, if you buy a new *Antenna Book* it comes with *Yagi Analyzer* (*YA*), a program by Brian Beezley, K6STI. Not only is this software optimized for Yagis, but also Brian has calibrated it to NEC—the Numerical Electromagnetic Code—greatly reducing the frequency-offset problem, at least below 432 MHz. Brian also sells more advanced programs with optimization routines and advertises them regularly in *QST* and *NCJ* (National Contest Journal). At 903 MHz, 3/16-inch elements are too thick for *YA* to handle, according to Brian's documentation, although you could use 1/8-inch elements. Or, you could use NEC to do the analysis of your microwave antennas for any element size.

But I distrust computer programs. I like to verify that the numbers they generate make sense. Like most people, I don't have a VHF antenna range handy with which to accurately measure antenna patterns. The primary problem with making such a range is reflections that distort the patterns. But you may be able to use a simple range to determine the frequency accuracy of an analysis program. Measure the antenna pattern at a sweep of frequencies, looking for pattern symmetry. For instance, as you get farther away from the design operating frequency the pattern may degenerate into a perfect dipole. For such a pattern, any reflections on the range should have little effect on the front-to-back (F/B) ratio—I'd expect it to still be close to 0 dB. At the frequency where the pattern *measures* 0 dB F/B ratio, the computer should *predict* 0 dB. If there is a frequency offset in the program, you can change the analysis frequency in steps until you find the symmetrical pattern you measured. The difference between the measured frequency and the analysis frequency is the frequency offset of the program. To be more accurate, you could plot the F/B ratio versus frequency and compare it to what the computer predicts. This technique should also tell you whether you made any serious errors in designing the antenna.

Be careful of drawing conclusions from impedance measurements or SWR curves, too. Once, I swept the return loss of a 903-MHz Yagi that had a T-matching network and found that it was quite easy to get a match high above the frequency I wanted, but difficult to match it at 903 MHz. I incorrectly concluded that the array was cut

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too high in frequency. But further tests indicated that the array was unusable at 923 MHz—the array was cut too *low* for this frequency. What happened was that in my swept measurements, the matching network effectively masked the true operating frequency of the antenna. Thus, while impedance measurements *seem* like a good idea, I don't believe they're as trustworthy as pattern measurements.

For the ultimate validation of the design, you can take your Yagi to a test range and see how it compares against antennas others have built. The Central States VHF Conference seems to be the premier event for antenna testing in the USA. This year it will be held in Colorado Springs, CO, on July 27-30. (See "Upcoming Technical Conferences," elsewhere in this issue of *QEX*.) Typically, measurements at Central States cover amateur bands from 144 MHz to 24 GHz.

Electrical Design Considerations: Loop Elements versus Rod Elements

One popular design approach is the "loop Yagi." A loop Yagi is actually a microwave version of a Quad antenna. Both use loop elements, but loop Yagis have the loops attached to the boom with screws. At microwave frequencies, the loops are small enough to be easily supported at a single point. They can be squashed by rough handling but are relatively easy to restore to their original shape. It is also easy to make spare elements since only a few loop sizes are used. Ordinary Yagis normally have rod elements supported at their centers by the boom or by some sort of fitting attached to the boom. And rod-Yagi designers often find they have to make each director a different length to maximize electrical performance.

I'm inclined to believe that for good designs, the difference in gain between the two approaches is small—approximately 1 dB for antennas of the same boom length. So it makes a whole lot of sense to copy a popular existing design, at least as a reference, then compare it to your new design. (This assumes that the design has been tested and optimized correctly.) However, rod elements do appear to be a lot easier to model using a computer. Of course, having a computer isn't a necessity; people have come up with excellent designs with a good test range. In fact, many designs have been optimized on a range, as opposed to a computer.

Modeling software works quite well for elements using rod elements because there is excellent measured data to calibrate the programs against. This is what Lawson did in his book, *Yagi Antenna Design*. He used experimental data published by Peter Viezicke of the National Bureau of Standards (NBS), now the National Institute of Science and Technology (NIST). (Buying this book from the ARRL is a lot easier than tracking down NBS report 688, which is out of print.) Unfortunately, there doesn't appear to be an equivalent set of data with which to calibrate parasitic loop element designs. On the other hand, loop Yagis seem to be more tolerant of typical construction tolerances—even at fairly high frequencies. Down East Microwave sells 3456-MHz loop Yagis—and they work, although electrical performance is noticeably inferior to a 2-foot dish, as one might expect from theoretical calculations. Then again, the significantly lower wind load of the loop Yagi is a decided advantage.

Frequency Optimization

The proper frequency to design a Yagi for is a tradeoff between gain and the effect of detuning by rain and ice. The gain gradually increases with frequency, then drops off quickly. Water on the array usually lowers the resonant frequency of the elements. Thus, if you set the frequency exactly at the point of highest gain, you get an array that quickly becomes nearly useless if loaded down with enough rain or ice. K1FO discusses the effect of rain in the *UHF/Microwave Experimenter's Manual* but doesn't try to tackle the tougher problem of ice loading. Another issue to consider is the effect of other Yagis in an array; the Yagis can detune each other.

Stacking Yagis

As far as I can determine, none of the experts have been able to devise simple algebraic stacking rules—there is no simple formula for stacking Yagis for more gain. I think the experts vary the stacking distance and look at the computer results till they get the best compromise between gain and unwanted sidelobes. Of course, if you don't care what the pattern is like you can stack them far apart and get nearly 3 dB of gain—along with nasty sidelobes that may trick you into pointing the antenna incorrectly. (In textbooks, unwanted sidelobes are also called grating lobes.) A terrible pattern, coupled with reflective objects to provide bounce paths, can result in a terribly confused VHF contester. Where *is* that weak signal coming from?

Mechanical Considerations: Boom-to-Mast Clamps

In order to get anywhere near the theoretical gain, the Yagis in a multiple-Yagi array all have to be pointing in the same direction. This isn't always trivial, due to the poor mechanical mounts used by many amateurs. Your typical "one size fits all" U-bolt and saddle assembly offers very little surface contact; it requires very little torque to rotate the antennas out of alignment. Ideally, blocks of aluminum would be machined into perfectly matching surfaces, but this is out of the realm of practicality for most of us.

I've had good results making saddles out of square extruded aluminum. You can get an excellent matching surface by using an appropriate size of Greenlee hole punch. These devices were once popular for making holes for tube sockets. They are still available but are no longer a standard in the amateur's tool box. If necessary, a reamer can be used to accommodate odd-size tubing, should that be necessary. After the holes are enlarged to the proper size, the extrusion is cut in half with a saw, then smoothed with a file. These saddles work well with the little antennas I've built.

Wooden Booms

Wooden booms are used in the popular "Quagi" antennas designed by Wayne Overbeck, N6NB. Nearly 20 years old, these designs still appear in the *ARRL Antenna Book*. While not the ultimate in gain or pattern, Quagis are easy to make out of readily available materials. Hey, another dB of gain is meaningless unless you can get the antenna built!

Wooden booms are also used in a set of Yagis designed by Kent Britain, WA5VJB, for 144 through 1296 MHz. Kent reinvented a simple feed that appeared in *Understanding Amateur Radio* (ARRL, 1963). Wood is easy to work with, both mechanically and electrically. Unless the wood boom is very thick, its effect on the Yagi elements can often be ignored. This eliminates a major source of uncertainty when designing Yagi antennas. This only applies if you support the elements at the center points, where you have a voltage null. The primary disadvantage of wood is its poor strength-to-weight ratio, at least when compared to high-quality aircraft aluminum.

Conductive Metal Booms

Metal booms are used in loop Yagis, originally invented by M. H. Walters, G3JVL. A 23-cm version using materials with American dimensions was developed by Chip Angle, N6CA, and appears in *The ARRL UHF/Microwave Projects Manual*. You can read up on Walters' work in Volume 1 of the RSGB's *Microwave Handbook*, and Walters' program for designing loop Yagis is on the companion disk to the *UHF/Microwave Experimenter's Manual*. These are elegantly designed quad antennas for the lower microwave bands.

The problem with conductive metal booms is the necessity of maintaining good electrical contact. A quad loop does this automatically, since it has a natural tendency to spring apart and press against the screw that holds it down onto the boom. Occasionally, I find it necessary to retighten the screws on my loop Yagis, which isn't a problem for a portable station stored in my apartment. I think this design is elegant because many nasty problems are solved simply. The simple feed that doubles as a balun, for instance. Also, you can get away with only the top holes in the boom being lined up properly. The bottom holes can be a little off without affecting electrical performance. With designs that have the elements sticking through the boom, you have the much tougher task of drilling holes for the elements that are precisely aligned on both sides of the boom.

The main disadvantage of loop Yagis is getting thin strips of metal for the elements. You want thin strips, but not *too* thin. A lot of rather frail loop Yagis have been built! My 903-MHz loop Yagi uses 0.050-inch 6061-T6 aluminum sheet. Two years ago, at an antenna measuring session, a number of people commented on how ruggedly it was built. The loop elements really have to be cut with some sort of shear to get the needed accuracy, and not everyone has access to a shear. In fact, I've even heard that some high-tech machine shops have difficulty cutting the thin strips. This isn't a problem with old-fashioned, foot-operated shears.

The driven element of a loop Yagi is usually mounted with a large-diameter brass bolt, and the bolt is drilled through lengthwise to pass the feed line into the loop. Drilling this bolt can be a problem—there isn't much room for error. Note that, while hex-head bolts are specified in Chip Angle's article in the *Antenna Book*, the exact head type isn't important. I've also used bolts designed to hold toilets in place—check out the plumbing section of the hardware store to see what is available.

For rod elements, an approach used by Ranier Bertelsmeier, DJ9BV, and Günter Hoch, DL6WU, is to force aluminum rods through slightly small holes in an aluminum boom. They highly recommend painting the antenna to avoid corrosion. Square aluminum stock, if you can get it, will make hole drilling much easier. This is discussed in the 2/94 issue of *DUBUS*.

With rod elements passing through the boom, it is important for both sets of holes to be precisely drilled. The ideal tool for this is a vertical mill with an extremely long bed. Second best is perhaps a drill press. If you use a drill press, it makes little sense to drill a hole all the way through the boom if you are using a thin drill bit. The second hole is rarely centered properly because the bit invariably bends. It is a lot easier to drill the holes properly if you use square tubing. Unfortunately, I don't have a source of supply to recommend.

Insulated Metal Booms

K1FO insulates the elements from the metal boom in his high-performance 144, 222 and 432-MHz Yagi designs that appear in the *ARRL Handbook*. The advantage of using insulated elements is the elimination of possible noisy contacts while still using a rugged aluminum boom. I've used this technique on many of my VHF Yagis, but it requires more hardware parts than uninsulated element mounting. Unfortunately, Rutland Arrays, which was an excellent source for the stainless-steel retaining rings and machined Delrin insulators, is out of business due to the untimely death of Tom Rutland.

Design Guidance

The general consensus is that loop Yagis are good choices from 903 to 2304 MHz. For 2.4-GHz satellite work you generally can't afford the polarization loss of 3 dB, though this may change when Phase 3D goes aloft. Helical antennas make a lot of sense if you need circular polarization. At 3456 MHz and higher, parabolic reflectors or horns generally make more sense. At 432 and lower, Yagis with rod elements are usually preferred, though the Quagi antenna is easy to construct. Interestingly, Quagis and loop Yagis are both within about 1 dB of what an optimized Yagi can do. Thus, if you really need that dB, it may make sense to design an optimized Yagi. On the other hand, many people might be better off simply copying an existing design, rather than expending a lot of work for relatively little benefit.

A heavily optimized Yagi can make a lot of sense if you are building a large array of them. Fractions of a dB can add up to a serious savings in cost for a given gain goal. Not only can you get improvements in forward gain, but cleaner patterns may allow you to get closer to that 3-dB theoretical stacking gain.

While computer modeling has come a long way, I think it often comes up short when it comes to matching Yagis to the feed line. I wouldn't worry if you find you need a lot of experimentation to get the antenna matched. In fact, the Gamma match came about because the inventor couldn't get his T match to work! (I have noticed that if you do use a T match, the proper length of the driven element is often nearly the length of the reflector.)

If you want a no-tune Yagi, perhaps the best technique is to split the element and use a matching network that doesn't change the accuracy of the driven element model. An example would be a quarter wavelength of RG-83/U 35-Ω coax, which would match 24.5 Ω to 50 Ω. This exotic coax costs about $3 a foot, so it might make more sense to use two pieces of 75-Ω coax in parallel. In this case, the optimum impedance for the antenna would be 28 Ω.

I find that using brass tubing for rod elements—instead of solid rod—makes it easier to vary the element lengths. You can slide in another piece of tubing or rod to vary the length. Once you get real close, you can use the inserted tubing as a joining section so the element is of constant thickness. The advantage of using brass is the ability to solder the electrical connection, preventing a noisy or intermittent contact.

I've noticed a problem with Teflon type-N connector jacks—the newer ones have their center pins held in via friction. It is quite possible to move the pin when attaching it to the antenna. This can destroy the pin by causing the metal fingers to splay apart when you attach a cable. Or, there may be little or no contact, resulting in a poor electrical connection. You might use a trick I picked up from Mark Wilson, AA2Z—when you're not sure if all the parts of your connector are positioned correctly, compare it to an expensive adapter, which usually has the right mechanical dimensions.

Power Dividers

I've noticed in building power dividers for stacking purposes that there is a bit of uncertainty concerning the formula for coax with a square outer conductor. The power divider articles all seem to use $Z_0=138\log(1.08D/d)$, while you see $Z_0=138\log(1.178D/d)$ in tables for calculating transmission-line impedance. It appears that the former is an experimentally derived number that works well for building power dividers with quarter wavelengths of transmission line, while the latter is theoretically derived. I get much better results with the 1.08 constant, although I lack the precision equipment necessary to be sure enough to give anything besides rather fuzzy numbers.

Improving the Antenna Book *1296-MHz Loop Yagi*

I've a few comments about the design in Chapter 18 of the *Antenna Book*. To begin with, the sheet aluminum alloy isn't specified. I've had good results making everything out of 6061T-6 aluminum. Not only is this alloy inexpensive and rugged, but also it is easy to drill nice, clean holes in those mounting plates. A possible disadvantage, the springiness that makes forming loops difficult, is actually an advantage. I'd rather spend the extra time carefully putting it together than spend more time out in the field reforming elements that got bent during transportation and installation. Wearing safety goggles when assembling the elements might be a good idea—those little loops can get away from you! I've also heard of people making elements of stainless steel, which is even more rugged than aluminum. The losses are higher, though, due to the increased resistivity of the material.

Instead of the specified copper driven element, I used a brass strip that is more rugged. I also eliminated the butt joint shown, so the brass completely surrounds the semi-rigid coax shield. This is much less likely to break—my copy of the original design broke while getting up to Mt Washington. (Fortunately, we had a gas generator and a soldering iron.) Similarly, I recommend folding the center conductor of the coax against the brass to make the joint more rugged. Instead of using a relatively exotic type-N connector designed to attach to UT-141, I've successfully used a UG-58 panel receptacle with a UG-177 panel hood. The shield of the semi-rigid coax is soldered to the hood. It seems to work just fine at 1296 MHz, even if you don't do anything special to avoid an impedance bump. An aluminum bracket fastens the coax jack to the boom.

Running QRP in the ARRL and CQ VHF Contests

An inspection of the rules reveals a significant difference in allowed power levels between the ARRL and CQ VHF contests—the ARRL only allows 10 W, while CQ allows 25 W. Maybe 4 dB isn't significant for most contacts, but it does make a difference when probing the outer limits of your troposcatter range. One way of designing a station to accommodate both limits is to run stacked antennas fed with separate 10-W power amplifiers. For the ARRL contest you can run a single amplifier before the power divider.

For example, on 13 cm you could run two 10-W bricks, each to 20-dBi loop Yagis. This would give you an EIRP of around 36 dBW, a rather potent QRP signal—some multioperator stations don't have a signal this big on this band! Since microwave antennas are often degraded if a mast runs through them, I recommend putting the loop Yagis on an H-frame. For example, W1XX/3 had a pair of 23-cm Yagis on one side and a pair of 13-cm Yagis on the other side of the frame. To make the antennas easier to point, the antennas were stacked vertically. It would make more sense from a mechanical point of view to stack them horizontally, so the array would be balanced, but the broader azimuthal pattern of the vertical stack makes the antenna easier to point. Interestingly, for FM work you can often use a sharp pattern to better sort out stations on the same channel. Coupled with the fact that vertically polarized Yagis lend themselves to horizontal, as opposed to vertical, stacking on a cross boom, horizontal stacking is much more popular for FM work.

The ARRL June VHF contest will be June 10-12, while the CQ VHF contest will be on July 8-9. See you in the contests!

Upcoming Technical Conferences

The Central States VHF Society Conference

July 27-30, 1995, Sheraton Colorado Springs Hotel, 2886 South Circle Drive, Colorado Springs, Colorado 80906, 1-800-325-3535 or (719) 576-5900.

Contact: Lauren Libby, KXØO, President at (719) 593-9861, email: 75151.2442@compuserve.com or Hal Bergeson, WØMXY, Vice President and Program Chairman, 809 East Vermijo Avenue, Colorado Springs, CO 80903, (719) 471-0238, email: bergeson@ppcc.colorado. edu.

Events: Thursday night, side trip to "Flying W Ranch" for western music and beef barbeque. Friday and Saturday, technical and operating talks for all, beginners to the experienced ham. There will also be a special "Young People's" VHF program for those under 21. Saturday night's banquet features a presentation by Arnie Coro, CO2KK, on VHF opration in Cuba and a taste of Cuban culture. Sunday morning, SMIRK will sponsor a breakfast get-together.

Special Programs: Several "wives and kids" activities are being planned, including a "British High Tea" and a tour of the Glen Eyrie castle, a historical landmark in Colorado Springs. Shopping, galleries, local landmarks, unique shops and eating places are also part of this program.

Special Note: Colorado Springs is a summer tourist town. The town and surrounding area is a great place for a summer vacation. However, that presents special problems when it comes to booking room reservations. The Central States VHF Conference has a large block of rooms reserved for the Conference. Please take advantage of that as you will find it difficult to find rooms otherwise. Get your reservations in early; when our block of rooms is gone, you may have trouble finding another room.

1995 ARRL Digital Communications Conference

September 8-10, 1995, La Quinta Conference Center, Arlington, Texas—just minutes from Dallas/Fort Worth Airport. Co-hosted by Tucson Amateur Packet Radio(TAPR) and the Texas Packet Radio Society.

For more information contact the TAPR office at 8987-309 E. Tanque Verde Road #337, Tucson, AZ 85749-9399, tel: (817) 383-0000; fax: (817) 566-2544; Internet: tapr@ tapr.org

Call for papers: Deadline for receipt of camera-ready papers is **July 21, 1995**. Contact Maty Weinberg at ARRL HQ (tel: (203) 666-1541; fax: (203) 665-7531; Internet: lweinberg @arrl.org) for infomation on submitting papers.

The 1995 AMSAT Annual Meeting and Space Symposium

October 6-8, 1995, in Orlando Florida. For more information contact: Bob Walker, 6601 SW 16th Street, Plantation, FL 33317, (305) 792-7015, email: n4cu@amsat.org.

Call for papers: Those interested in submitting papers are asked to send in a summary by July 1, 1995. The deadline for camera-ready copy is August 12. Inquiries should be sent to Bob Walker at the above address.

Microwave Update 95

October 26-28, La Quinta Inn, Arlington, Texas.

For more information contact: Al Ward, WB5LUA, 2306 Forest Grove Estates Road, Allen, TX 75002 or Kent Britain, WA5VJB, 1626 Vineyard, Grand Prairie, TX 75052-1405.

Bits

Free SPICE Newsletter Previews New Simulation Techniques

The *Intusoft Newsletter* is a unique free publication dedicated to discussing the SPICE circuit simulation program and its use. The April 1995 DAC Preview issue contains information on simulating digital filters and current-mode designs, as well as announcements on several new analog and mixed-signal products.

The first article discusses new techniques for simulating digital filters with *IsSpice4*. The *IsSpice4* program includes a unique event-driven simulator that can simulate sampled data systems in addition to digital elements. This capability is not available in other SPICE programs. The event-driven simulator is integrated with the analog transient analysis, enabling mixed systems to be described by a single schematic design. As examples, a 2nd-order IIR low-pass filter and an FIR filter are simulated. The FIR filter is used in a signal averaging circuit with results presented.

Current-mode design techniques are not new, but the development of high-performance complementary bipolar processes with excellent matching properties has allowed current-mode principles to be fully exploited. The emergence of monolithic devices has led to a renewed awakening of the benefits of current-mode design and produced a wealth of circuits with broadband properties. This article discusses how one type of current-mode element, current conveyors, can be simulated in *IsSpice4*. An example of a 30-MHz precision full-wave rectifier is presented.

Another article introduces a new simulation concept: "Mixed Environment" simulations. Intusoft's new Array Processing Code Model Library (APCML), to be introduced at DAC, provides a variety of array processing functions for use with the *IsSpice4* simulator. The APCML also provides a generalized OLE 2.0 interface for users to add their own OLE 2.0 in-process servers. The APCML allows SPICE users to incorporate real-time sound and images in their electrical simulations, a capability never before incorporated with SPICE.

Other articles discuss Intusoft's new multithreaded, multiprocessor version of IsSpice4, as well as Intusoft's new Power Macintosh SPICE offerings.

The newsletter also lists where to get a free working copy of the ICAP/4 Windows SPICE software on the Internet. This demo is fully functional and can simulate circuits up to 15 components. It includes schematic entry, a SPICE 3 simulator, a small-model library and a waveform processor.The Internet site can be reached at: ftp.iee.ufrgs.br. After logging in as anonymous, type "cd intusoft."

An optional floppy disk, containing all of the schematics and SPICE models in the newsletter along with new models for SGS-Thomson, Analog Devices and élantec op-amps is available for a nominal fee.

An electronic version (Replica document) of the *Intusoft Newsletter* is posted on CompuServe (CADD/CAM/CAE Vendor Forum). The newsletter circulation is currently at 34,000 readers in over 110 countries.

For more information contact: Charles Hymowitz, Vice President Intusoft, PO Box 710, San Pedro, CA 90733-0710; tel: 310-833-0710; fax 310-833-9658; e-mail: 74774.202@compuserve.com; Internet: ftp.iee.ufrgs.br.