ARRL's

HANDS-ON RADIO EXPERIMENTS

H.Ward Silver, NØAX

Volume 3

Basic electronics experiments from the pages of QST magazine

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Updates

There is often additional information regarding the Hands-On Radio experiments — you can find it on the Hands-On Radio Web page at **www.arrl.org/hands-on-radio**. The contents include links to reference articles and sources of supplies and equipment. There is also a Frequently Asked Questions section that provides explanations about many of the experiments. Readers have contributed their observations and sometimes tools and links that relate to the topic in the experiment. You may find the information helpful as you perform the experiments in this anthology.

Contents

Foreword

Welcome to the third volume of Hands-On Radio Experiments. The companion Volumes 1 and 2 span a 15-year set of columns in *QST*. When the column started long ago, we weren't sure if it would run more than a few months, but readers were certainly interested and here we are, some 179 columns later!

This volume is similar to the previous volumes in that it presents an eclectic mix of topics spanning the practical through theory. There are several columns that describe how components — even wires and PC traces — behave at RF, which is often quite different from their performance at lower frequencies and dc. Knowing how to account for the difference can make or break a project or repair.

Volume 3 has a bit more emphasis on transmission lines, antennas, and some associated circuit functions. Even if the column doesn't include an experiment per se, be sure to build one or two of the antennas or circuits to see how they work. Then be sure to compare the results to what you expected. That's a very important part of learning about radio and wireless in general.

The columns on grounding and bonding generated so much interest that I extended the conversation to the recent ARRL book of the same name. This is an area of many mis- and incomplete understandings, which lead to confusion and poor practices. It is the Hands-On Radio readership that deserves a great deal of credit for encouraging me to collect the necessary information and put it into one book where it can serve as a reference and a guide to station and antenna system building.

Finally, the last columns covered that most mysterious of topics, Maxwell's equations. With their somewhat advanced notation, they look very intimidating. I've tried to demystify both the symbols and the concepts, explaining how radio waves appear as a consequence of "inconveniencing electrons" by accelerating and decelerating them. Since we make use of radio waves in ham radio, it seemed like a good idea for us to understand where they come from!

This volume completes the Hands-On Radio series in its present form. There will be other columns from excellent writers, just as there were in the past. ARRL members will enjoy searching the ARRL archives for columns by legendary writers like George Grammar, W1DF, and Doug DeMaw, W1FB, who were certainly inspirations to me. Or why not write an article yourself? *QST* is our shared resource and depends on each of us for contributions, as well as readership. If you know a topic well, your first article might be close at hand.

Thanks for reading and 73!

Ward Silver, NØAX

About the ARRL

The seed for Amateur Radio was planted in the 1890s, when Guglielmo Marconi began his experiments in wireless telegraphy. Soon he was joined by dozens, then hundreds, of others who were enthusiastic about sending and receiving messages through the air — some with a commercial interest, but others solely out of a love for this new communications medium. The United States government began licensing Amateur Radio operators in 1912.

By 1914, there were thousands of Amateur Radio operators — hams — in the United States. Hiram Percy Maxim, a leading Hartford, Connecticut inventor and industrialist, saw the need for an organization to unify this fledgling group of radio experimenters. In May 1914 he founded the American Radio Relay League (ARRL) to meet that need.

ARRL is the national association for Amateur Radio in the US. Today, with approximately 167,000 members, ARRL numbers within its ranks the vast majority of active radio amateurs in the nation and has a proud history of achievement as the standard-bearer in amateur affairs. ARRL's underpinnings as Amateur Radio's witness, partner, and forum are defined by five pillars: Public Service, Advocacy, Education, Technology, and Membership. ARRL is also International Secretariat for the International Amateur Radio Union, which is made up of similar societies in 150 countries around the world.

ARRL's Mission Statement: To advance the art, science, and enjoyment of Amateur Radio.
ARRL's Vision Statement: As the national association for Amateur Radio in the United States, ARRL:

- Supports the awareness and growth of Amateur Radio worldwide;
- Advocates for meaningful access to radio spectrum;
- Strives for every member to get involved, get active, and get on the air;
- Encourages radio experimentation and, through its members, advances radio technology and education; and
- Organizes and trains volunteers to serve their communities by providing public service and emergency communications.

At ARRL headquarters in the Hartford, Connecticut suburb of Newington, the staff helps serve the needs of members. ARRL publishes the monthly journal *QST* and an interactive digital version of *QST*, as well as newsletters and many publications covering all aspects of Amateur Radio. Its headquarters station, W1AW, transmits bulletins of interest to radio amateurs and Morse code practice sessions. ARRL also coordinates an extensive field organization, which includes volunteers who provide technical information and other support services for radio amateurs as well as communications for public service activities. In addition, ARRL represents US radio amateurs to the Federal Communications Commission and other government agencies in the US and abroad.

Membership in ARRL means much more than receiving *QST* each month. In addition to the services already described, ARRL offers membership services on a personal level, such as the Technical Information Service, where members can get answers — by phone, e-mail, or the ARRL website — to all their technical and operating questions.

A bona fide interest in Amateur Radio is the only essential qualification of membership; an Amateur Radio license is not a prerequisite, although full voting membership is granted only to licensed radio amateurs in the US. Full ARRL membership gives you a voice in how the affairs of the organization are governed. ARRL policy is set by a Board of Directors (one from each of 15 Divisions). Each year, one-third of the ARRL Board of Directors stands for election by the full members they represent. The day-to-day operation of ARRL HQ is managed by a Chief Executive Officer and his/her staff.

Join ARRL Today! No matter what aspect of Amateur Radio attracts you, ARRL membership is relevant and important. There would be no Amateur Radio as we know it today were it not for ARRL. We would be happy to welcome you as a member! Join online at

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Electronic Circuits

Experiment #125 — The Schmitt Trigger

Once you start building radio gear, you learn that a great deal of "radio" has little to do with RF. This month we're going to work with one of the non-radio building blocks you'll encounter when an analog signal and a digital function come together — the *Schmitt trigger*. Long-time readers may recall the Schmitt trigger making an appearance in Experiment #11 about op-amp comparators.

[1]

,

[2]

Adding positive feedback to the basic comparator circuit creates *hysteresis* — a switching threshold that changes depending on whether the circuit's output is on or off. This turns out to be desirable in certain applications.

If you're not familiar with op-amp comparators, download and read Experiment #11 from the "Hands-On Radio" web page. Page two covers hysteresis and how to design the comparator-based Schmitt trigger with discrete components. If you have an LM311 (or one of the equivalents listed) build a Schmitt trigger circuit and perform the experiment.

The Logic IC Schmitt Trigger

The ability to switch reliably in the presence of noise is a valuable function for digital circuits that have analog signals as inputs. Op-amps and discrete components take up valuable printed circuit board space, so the Schmitt trigger function was packaged into an IC. The basic set of six hex inverters (7414-type or CD4069 ICs) and quad NAND gates (74132 or CD4093) are the most common Schmitt trigger ICs. They are inexpensive and widely available.

[3]

The difference in switching characteristics between standard logic gates and Schmitt triggers can be seen in the device data sheets. Download the Texas Instruments data sheets for the CD4011 (standard quad NAND) and CD4093 ICs from

www.datasheetcatalog.org

. Look in the dc or Static Electrical Characteristics tables and find the Input Low and Input High voltage specifications for a V_{DD} value of 5 V at 25° C. For the standard gate the typical values of V_{IH} and V_{IL} are 3.5 and 1.5 V, respectively, and the response to an input signal in that 2 V range is undefined. For the Schmitt trigger IC, V_N and V_P are 1.9 and 2.9 V, only 1.0 V apart — a much smaller switching window — and drawing (b) at the bottom of the CD4093 datasheet's first page shows the *transfer characteristic* of the gate. Note that the output voltage is defined for all values of input voltage. Let's put that to work.

Sensing Slowly Changing Signals

Any device controlled by a microprocessor needs to have a **POWER_OK** signal to prevent its digital circuits from attempting to operate before the power supply is fully up and running. Such premature operation can yield strange results.

Similarly, when power is lost, the same signal notifies the digital circuits to shut down in a hurry. Specialized power monitoring ICs are available for this task, but the simple circuit in Figure 1 can also do the job.

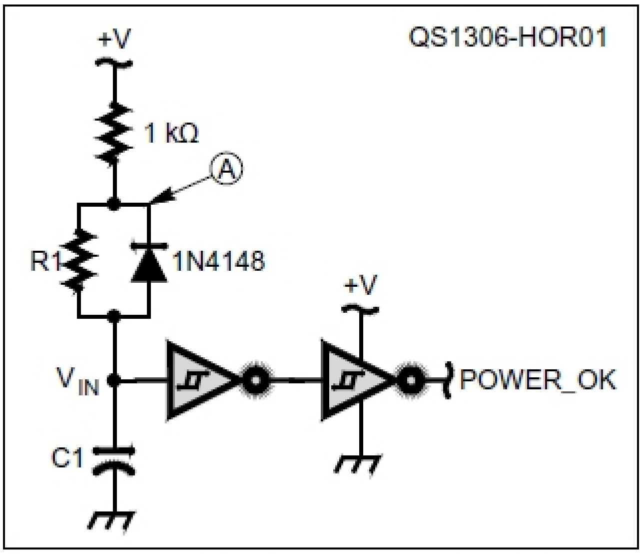

Figure 1 — A simple power-on circuit created from an R-C circuit and Schmitt trigger inverters.

When power turns **ON**, capacitor C1 charges slowly through R1, keeping V_{IN} below the buffer's V_P threshold for a time delay of about one time constant, R1 × C1. (The back-to-back inverters form a non-inverting *buffer*.) After that time, the buffer's **POWER_OK** signal goes high to indicate the power supply has had enough time to stabilize. If power is lost at any time during the charging process, or if power is turned **OFF** after stabilizing, C1 is rapidly discharged through the 1N4148 and the **POWER_OK** signal goes low.

You can build this circuit using two of the inverters in a 74HC14 IC powered by 5 V. To make it easy to observe the time delay, use a 1 MΩ resistor for R1 and a 1 µF capacitor for C1. The 1 kΩ resistor provides current limiting during the experiment and is small compared to R1, so it has an insignificant effect on the charging time of C1.

Build the power-down detector using a CMOS hex inverter such as the 74HC14 or CD40106. TTL versions (7414 or 74LS14) draw too much current for the 1 MW resistor to act as a pull-up. Also, in a real-world design, the inverter would be powered from a large capa- citor to allow it to hold **POWER_OK** low for several msec, insuring a controlled shut-down period.

Watching the **POWER_OK** signal with a voltmeter or oscilloscope, apply power to the circuit and verify that the signal stays low and does not go high until about 1 second has passed. If you connect one end of a clip lead or piece of wire to ground and simulate a power dropout by brushing the other end against the cathode of the diode (point A in Figure 1), the output signal should immediately go low, signifying power is *not* OK and there should be a 1 second delay before it returns to the OK state. **POWER_OK** might be used as a reset signal for a digital circuit.

Switch Debouncing

A switch may feel quite solid to you, leaving little doubt that when you close a switch, it instantly closes and stays closed. In truth, the contacts of almost all mechanical switches and relays literally bounce for a few milliseconds before settling down to stay closed. Because digital devices are so fast, software can react to those bounces as multiple switch closings and openings. While it's possible to "debounce" a switch signal in software, it can be done with hardware, too.

Reconfigure your circuit as shown in Figure 2; only one inverter section is needed here. That wire you just used to simulate a power dropout can also simulate the noisy signal from a switch, or you can use a real momentary switch. The two time delays, t_{CL} and t_{OP} depend on the values of R1 and R2, respectively:

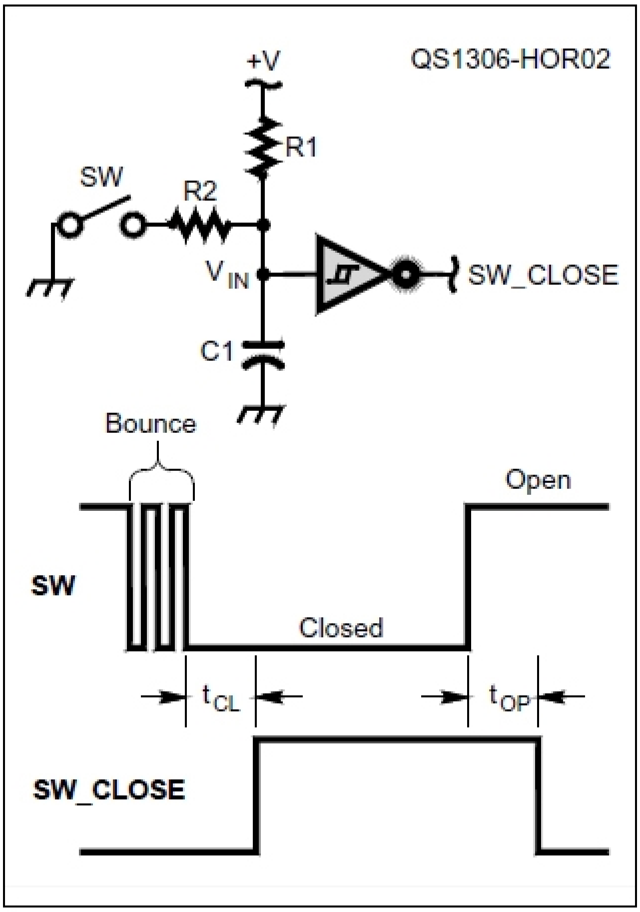

Figure 2 — Switch debouncing by using an R-C circuit and a Schmitt trigger inverter. R1 must be much greater than R2.

$t_{CL} \approx R2 \times C1$ and $t_{OP} \approx R1 \times C1$

Start with R1 = 1 MΩ, R2 = 10 kΩ, and C1 = 1 µF. You'll need an oscilloscope to see the bounces of the switch contacts and the short delay, t_{CL}.

If you swap R1 and R2, you'll find that R1 has to be much larger than R2 for **SW_CLOSE** to go high. If R1 is too small, closing the switch does not discharge C1 below the V_N threshold for the inverter because R1 "overpowers" R2 and keeps the capacitor charged to higher than V_N. Experiment with different values; you'll find that R1 has to be about three times larger than R2 to get reliable results.

The squelch function in your radio also requires a continuously changing signal to cause switching at a threshold. If the input signal to R2 is the rectified and filtered output of a receiver's audio amplifier, the output of the inverter indicates whether a signal is present.

Edge Detectors

In many cases you want to be able to tell when the input signal changes state without having to monitor it continuously. This requires an elementary form of memory that will allow a circuit that can compare "then" to "now." The circuit in Figure 3 uses a two-input XOR gate to make the comparison and an R-C circuit to create the time delay that acts as memory. The Schmitt trigger input is required because of the slowly changing R-C circuit output.

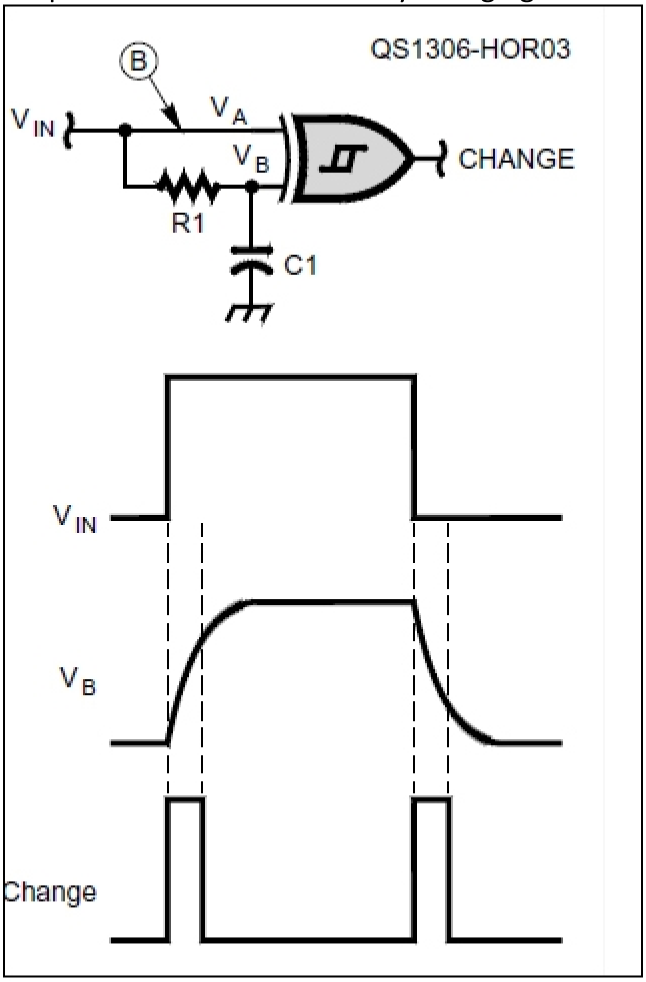

Figure 3 — The edge detector circuit uses the memory created by the R-C delay circuit. Inserting a Schmitt trigger inverter at B inverts the output pulses.

V_A is the "now" state of V_{IN} while V_B is the "then" state. An XOR gate's output is high only when just one of its inputs is high. If both inputs are in the same state — high or low — the output, **CHANGE**, is also low. During the delay period while C1 is charging or discharging (approximately equal to the time constant R1 × C1), one input lags behind the other so that the XOR function is true and **CHANGE** is high — but *only* during the charge/discharge period. The Schmitt trigger action insures that the slowly-changing voltage V_B causes only one pulse with every transition of V_{IN}.

You can build the circuit in Figure 3 using one section of a 74HC86 quad exclusive-OR gate with Schmitt trigger inputs. R1 and C1 values of 100 kΩ and 0.01 µF will provide output pulses about 1 msec long. Use a 50% duty- cycle, 5 V pulse output from a function generator or a 555 timer circuit as V_{IN} (don't use a square wave with a negative voltage) with a repetition rate of about 100 Hz. Vary the time constant of R1-C1 to see the effect on the output pulse width, viewed on an oscilloscope.

Another name for this circuit is a *frequency doubler*. Two output pulses occur for every input pulse — one at each edge of the input pulse. If you insert a Schmitt trigger inverter at point B, the **CHANGE** pulses change from positive pulses at each edge to negative pulses.

Shopping List

74HC14 hex inverter

74HC86 quad XOR gate

1N4148 diode

1⁄4-watt resistors: 1 kΩ, 10 kΩ 100 kΩ, 1 MΩ

Capacitors: 0.01 µF, 1 µF (ceramic or electrolytic)

Momentary switch (optional)

[1]

All previous “Hands-On Radio” experiments are available to ARRL members at www.arrl.org/hands-on-radio.

[2]

The Schmitt trigger is named for Otto Schmitt who identified the function when studying properties of squid nerves in the 1930s!

[3]

7400-family model ICs are available in many different logic families such as HC, LS, and AC.

Experiment #129 — Wye-Delta and Pi-T Circuits

I didn't realize Hands-On Radio had so many electricians and power engineers as readers! Experiment #127's hypothetical example of three-phase power to explain how phasors operate (combined with a sloppy math error) was not intended to be a tutorial on ac wiring practices, but the references cited in the article and on the Hands-On Radio web page should clear up any confusion I unintentionally generated on this important topic.
[1]

During the post-column post-mortem, I realized that I'd been given a golden opportunity to dig deeper into some important supporting circuit concepts and get us back to radio at the same time.

Why a Delta?

When studying three-phase power systems, you'll soon encounter the terms *wye* (pronounced "why") and *delta*. These refer to how the three individual phases are connected with respect to a neutral reference. Figure 1 illustrates the basic idea. The coils shown here — typical of motor or transformer windings — could also be voltage or current sources, resistors, capacitors, or generic impedances. The origin of the names for each type of system is clear — the schematics for each system take the shape of a **Y** or a Δ.

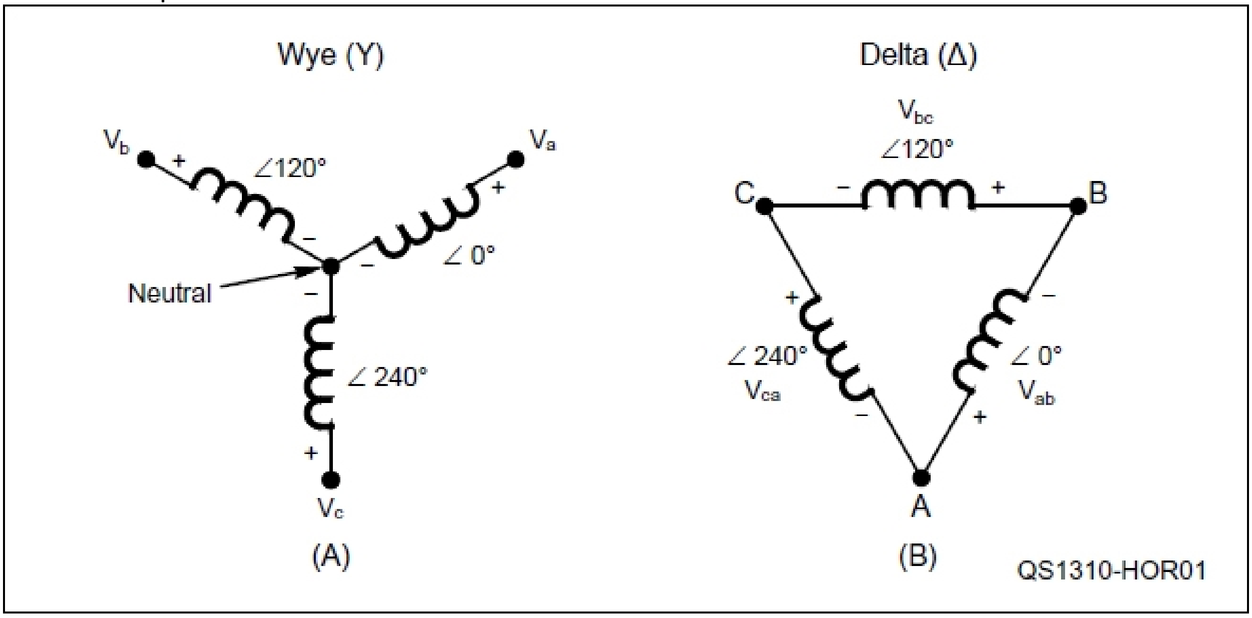

Figure 1 — Schematics for circuits connected in wye (A) and delta (B) configurations.

In a wye system, the three phases (a, b and c) all share a common neutral point, so there are four connections: Phase A, B and C (also commonly labeled Line 1, 2 and 3) and neutral. In a delta system, there is no common neutral connection because the sources or loads are connected together in a loop. That means there are only three connections and the voltages between them. (A neutral reference point can be created in a delta network through various techniques discussed in the reference articles.)

The angle between the phase voltages is always 120°, but whether the angle is positive or negative depends on the *phase sequence* which can be a-b-c or a-c-b in order of increasing angle. In Figure 1A, the phase sequence is a-b-c, which is positive rotation.

Say — how can three voltages be connected together in a loop that doesn't contain any resistance and not have the current go to infinity? If the voltages were dc, we would indeed have a problem! Instead, these are ac sine waves with the same voltage magnitude (V) but different phase angles. Adding up the voltages around the circuit gives us $V\angle 0^\circ + V\angle 120^\circ + V\angle 240^\circ$. Changing the phasors to rectangular coordinates allows us to calculate the sum: $(V + j0) + (-0.5V + j0.866V) + (-0.5V - j0.866V) = 0$. So the net voltage around the loop is zero and no circulating current flows at all!

In both wye and delta systems the loads can be connected between phase or line voltages. In a wye system, a load can also be connected between a phase voltage and neutral. Imbalanced loads in either type of system can cause substantial *error currents* to flow. Non-linear loads such as switchmode supplies and loads controlled by SCRs and TRIACs create harmonic currents. Both of these cause problems, too. Obviously, generating, transferring, and using multi-phase power is a complex subject. You can learn more about it in the references mentioned earlier and at
www.allaboutcircuits.com/vol_2/chpt_10/1.html
.

But what do wye and delta power systems have to do with radio? From the standpoint of ac power, not much — unless you happen to need a *really* big power supply. How- ever, we use wye and delta all the time in our circuits — we just refer to them as *Pi* and *T*!

Having T with Pi

Figure 2 shows two circuits made of generic impedances — one is a Pi network like you'll find in nearly every tube-type amplifier and the other takes the shape of a T network that you'll find in most antenna tuners. Figure 2 also shows how a

Pi network is the same as a delta network and a T network is a wye network. Who knew?

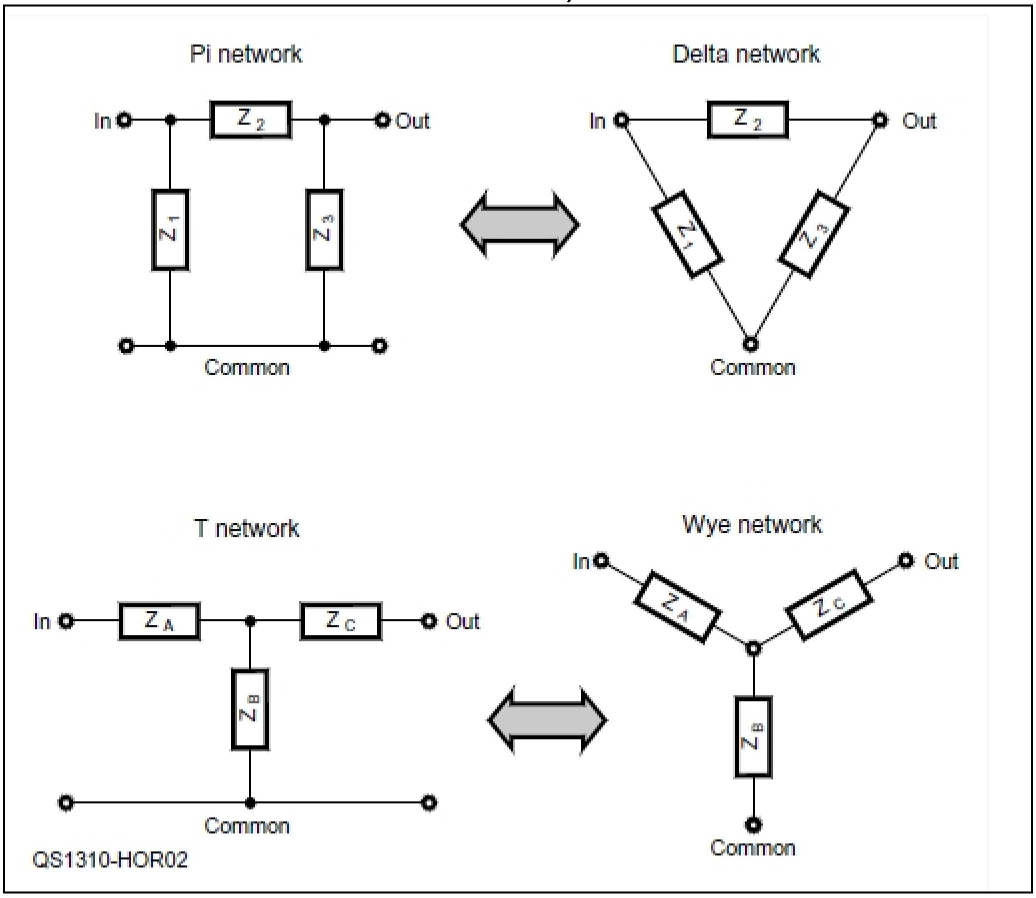

Figure 2 — A Pi network is equivalent to a delta network and a T network is equivalent to a wye network.

That's handy to know, but there is another neat trick to apply. You can turn a circuit of one type (Pi or T) into its exact equivalent circuit of the other type (T or Pi) by using some math called the *wye-delta transformation*. From the perspective of the input and output connections, the equivalent circuit will behave *exactly* the same (with a caveat explained later). The following equations show the math, although you can use an online calculator such as this one at

www.elektro-energetika.cz/calculations/transfigurace.php?language=english

2

Pi (Delta) to T (Wye)

$Z_A = Z_1Z_2 \ / \ \Sigma Z$; $Z_B = Z_1Z_3 \ / \ \Sigma Z$; $Z_C = Z_2Z_3 \ / \ \Sigma Z$

Where $\Sigma Z = Z_1 + Z_2 + Z_3$

T (Wye) to Pi (Delta)

$Z_1 = Z_P \ / Z_C$, $Z_2 = Z_P \ / Z_B$, $Z_3 = Z_P \ / Z_A$

Where $Z_P = Z_AZ_B + Z_AZ_C + Z_BZ_C$

Using the Transformation

More than just math sleight-of-hand, transforming the circuits from one form to the other can be quite useful. Let's say the circuit you start with has component values that are hard to make work well — maybe they are very large or very small values. By changing the circuit from one form to another, the component values also change and may become more reasonable. Let's try an example:

The circuit in Figure 3A is a 40 dB T network attenuator with symmetrical input and output impedances of 50 Ω. The 49 Ω series resistors aren't an issue, but the 1 Ω parallel resistance could be significantly affected by extra wiring resistance in the common con- nection. Transforming the circuit into its Pi equivalent in Figure 3B changes the resistors to 51 and 2500 Ω. The 2500 Ω resistance is much less affected by wiring resistance and other small variations. You can use an online calculator such as

www.microwaves101.com/encyclopedia/calcattenuator.cfm

or tables of resistor values for Pi and T attenuators allow you to pick the form that makes the most sense.

3

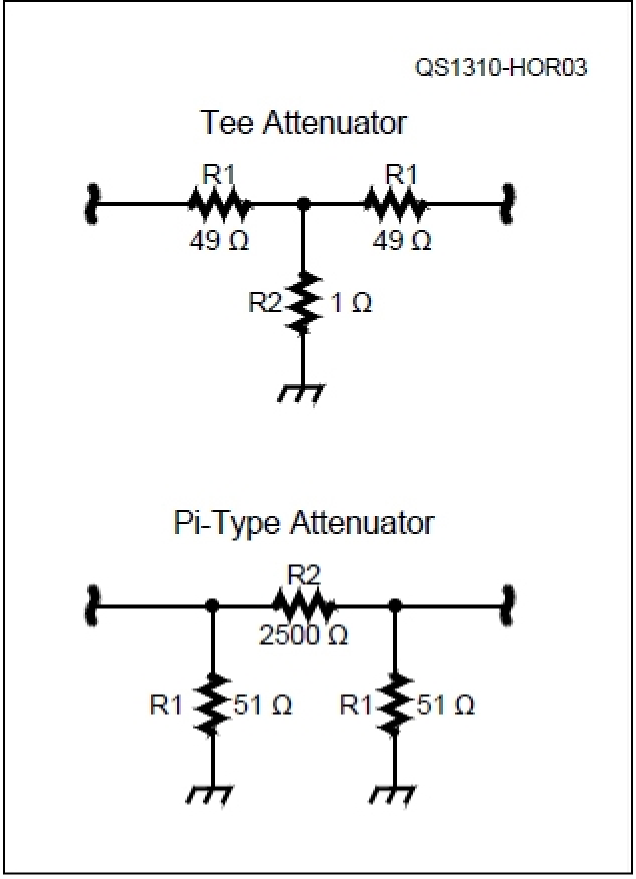

Figure 3 — A resistive 40 dB T attenuator and its equivalent Pi attenuator. The Pi attenuator's 2500-Ω resistor is less affected by wiring resistance than the T attenuator's 1 Ω resistor.

This can also work with circuits made out of reactances (Ls and Cs). Let's try an example in which we are transforming an output im- pedance of 800 Ω to 50 Ω at a frequency of 7 MHz and with a circuit Q of 6. If we use a typical antenna tuner's T network with series capacitors and a parallel inductor, the component values are C_{IN} = 76 pF, L = 5.9 µH, and C_{OUT} = 25 pF (calculated with the T match calculator at

www.eeweb.com/toolbox/t-match

). The value of L is fine but the values for C_{IN} and C_{OUT} are small enough that it would not take much stray capacitance to upset the tuning of the network which makes the settings "touchy." In addition, the higher the reactances of the series capacitors (–*j*300 and –*j*910 Ω, respectively) the higher the voltages across them for a given current.

What happens if the T (wye) network is transformed to a Pi (delta) network? The resulting reactances are arranged with a shunt (parallel) inductor at the input (*j* 44 Ω or 1 µH) and output (*j* 132 Ω or 3 µH) and a larger series capacitor (–*j*155 Ω or 147 pF) which would reduce the effect of stray capa- citance and lowers series reactance, too. The online calculator can also change the Pi net- work to a series-L form if that's more convenient.

[4]

Here's the caveat mentioned earlier: When using LC networks it's important to remember that when using LC networks, the transformation works *only* at the frequency for which the components have the reactances you specify. As frequency changes, so do the reactances, and you'll have to re-calculate the component values to get an exactly equivalent circuit. Using variable components allows you to use the circuit at different frequencies.

Putting the transformation calculators in your software toolbox is very helpful when you are trying to select and design an impedance matching circuit. This is particularly true at QRO power levels where heavy-duty (i.e. expensive) components are required to stand up to the high voltages and currents. A few iterations of your design over the range of frequencies and impedances you want helps avoid extreme component values and the high voltages and currents that often go with them.

References

[1]

All previous Hands-On Radio experiments are available to ARRL members at

www.arrl.org/hands-on-radio
.
2
Step 1 of the calculator asks for the “Shape of the complex numbers,” meaning that you should select either rectangular (R + *j*X) or phasor (Z∠θ) form for the impedances. Use a comma for the decimal in the European convention. i.e. 1.0 becomes 1,0.
3
See the Component Data and References chapter of the *ARRL Handbook*, available from your ARRL dealer, or from the ARRL Store, ARRL order no. 6948. Telephone toll-free in the US 888-277-5289, or 860-594-0355; fax 860-594-0303;
www.arrl.org/shop/
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.
4
The calculator can switch between series-C (block dc current) and series-L (pass dc current) forms of the network.

Experiment #132 — Resistor Networks

Resistors may seem to be the lowliest of components, but they are by far the most common. This month's column discusses several different types of useful resistor circuits that are handy designs to have in your toolbox. Dig in and try a few on your own.

Finding a Needed Parallel Value

We should all know the basic formulas for series and parallel resistors. In series, just add the values together: $R_{SER} = R1 + R2 + R3$…etc. For parallel resistors, the equivalent value is the "reciprocal of the sum of reciprocals": $R_{PAR} = 1 / (1/R1 + 1/R2 + 1/R3 + …etc)$. Happily, for two resistors this simplifies to $R_{PAR} = R1\,R2 / (R1 + R2)$. But what if you need to create a certain resulting value of R_{PAR} — what two resistances in parallel can you use? Start by picking a value for $R1$ that is higher than the desired value. Once a little algebra is applied to rearrange the equation, the needed resistor's value is

$R2 = R_{PAR}\,R1 / (R1 - R_{PAR})$ [1]

This is a handy formula to have in a calculator or spreadsheet so I've created an *Excel* spreadsheet containing all the formulas in this column. It's available on the "Hands-On Radio" web page.
[1]

Power Dissipation

Power dissipation for a single resistor is equal to V^2/R or I^2R, but what happens when you have more than one resistor in series or parallel? It sounds complicated, until you realize that in a series circuit the same current flows through all of the resistors — use I^2R. Similarly, the same voltage appears across all of the resistors in a parallel circuit — use V^2/R.

Voltage Dividers

If you are given both resistor values in the voltage divider shown in Figure 1 and for a moment ignore the load resistor, RL, it is straightforward to figure the output voltage: $V_{OUT} = V_{IN}\,R2 / (R1 + R2)$. But what if you need a specific division ratio (V_{OUT} / V_{IN}) and want to know what resistor values to use? Start by choosing the total resistance of the divider, $R_{TOT} = R1 + R2$, then

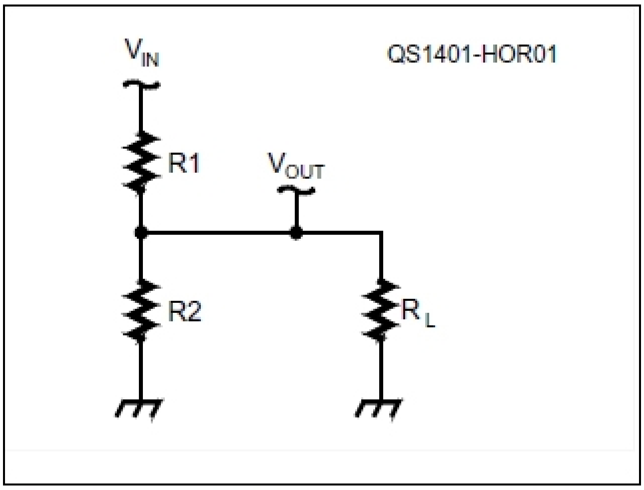

Figure 1 — A simple voltage divider. The effect of the load resistor, RL, on the voltage division ratio becomes significant as RL is reduced to less than 10 times the value of R2.

$R2 = R_{TOT}\,(V_{OUT} / V_{IN})$ and $R1 = R_{TOT}\,(1 - V_{OUT} / V_{IN})$ [2]

Or maybe you know the division ratio and already have a value for $R2$ — note that the division ratio is inverted in this equation:

$R1 = R2\,(V_{IN} / V_{OUT} - 1)$ [3]

The voltage division ratio is also affected by the load, RL, attached across $R2$ at the output of the divider. If you need a precise ratio, remember to include the effect of RL by substituting $RL\ //\ R2$ for $R2$ in equations 2 and 3. (The symbol // means "in parallel with.")

A good rule is that to avoid large effects on the voltage division ratio, RL should be at least 10 times greater than $R2$. If this is not practical, an alternative is to connect the voltage divider output to a high-impedance buffer circuit such as an emitter-follower ("Hands-On Radio" Experiment #2) or op-amp voltage follower (Experiment #3).

Current Dividers

Sometimes, instead of dividing a voltage, you need to divide a current, such as when a load current is too large to measure directly. For resistors in parallel as in Figure 2A, the current through any one of them, R_N, is

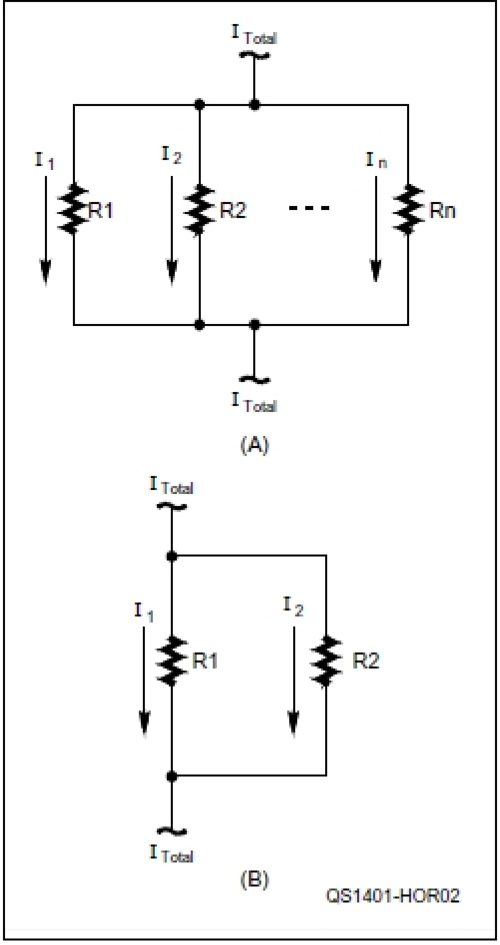

Figure 2 —A general current divider at A and the more common two-resistor current divider at B. The schematic in B also describes a meter shunt with R2 representing the meter.

$I_N = I_{TOT} (R_{PAR} / R_N)$ [4]

where I_{TOT} = total current through all resistors and R_{PAR} is the combined parallel resistance of all the resistors.

Let's do a sample calculation for two resistors. If you have 1.5 mA flowing through the parallel combination of a 1 kΩ and a 470 Ω resistor, how much current is flowing through the 470 Ω resistor? And through the 1 kΩ resistor?

I_{470} = 1.5 mA (1 kΩ // 470 Ω / 470 Ω) = 1.5 mA (320 Ω / 470 Ω) = 1.0 mA

I_{1k} = 1.5 mA (1 kΩ // 470 Ω / 1 kΩ) = 1.5 mA (320 Ω / 1 kΩ) = 0.5 mA

This makes sense — the lower-value resistor carries more current. It's always good to check on your calculations!

Here's a more common situation: diverting a specific fraction of the circuit's total current, I_{TOT}, through a measurement circuit as in Figure 2B. With two resistors in parallel and $R2$ the resistor in the measurement path, the current division ratio is $I_2 / I_{TOT} = I_2 / (I_1 + I_2)$. To limit the effects of adding resistance to the circuit, you must also pick a suitably small value for the total amount of resistance you are adding, $R_{PAR} = R1 // R2$. By rearranging equation 4, we can find $R2 = R_{PAR} / (I2 / I_{TOT})$ and rearranging equation 1 gives us the value of $R1 = R_{PAR} R2 / (R2 - R_{PAR})$.

An example would be nice! Let's say I want to divert 1% of a 750 mA current through the measurement branch of my circuit. (The division ratio for 1% is 0.01.) If I can add 1 Ω of total resistance to the circuit, I have $R_{PAR} = 1$. Start by finding the measurement branch resistance, $R2$ = 1 / (0.01) = 100 Ω. The main current-carrying resistance must be $R1$ = 1 × 100 / (100 – 1) = 100 / 99 = 1.01 Ω. One more thing — make sure R_1 can dissipate the total power of $0.75^2 \times 1.01 = 0.57$ W. And also remember that resistors change value when they heat up, so it might be smart to use a 5 W or larger resistor for $R1$ to keep its temperature rise to a minimum.

Meter Shunts

Another very common application of current dividers is the meter shunt. When you see a meter calibrated in amps of current, the delicate meter movement itself is not carrying all that current. Usually the meter is a milli- or microammeter that measures a small sample of current diverted through it as we just described. Analog meters with a full-scale current (I_{FS}) of 100 µA to 10 mA are fairly common. Here's how to use them for measuring larger currents.

If you don't know the meter's I_{FS}, connect a 1.5 V battery or other low voltage source, a 10 kΩ potentiometer, and DMM (in current-measuring mode) in series with the meter. Adjust the resistance for full-scale current (start at maximum resistance) and read I_{FS} from the DMM.

Next determine the meter's internal resistance, R_M, but don't hook it up to an ohmmeter! The current from the ohmmeter could damage a sensitive meter movement, so use the circuit in Figure 3. Meters with I_{FS} of less than 1 mA typically have an R_M of 1000-5000 Ω and for I_{FS} of 1 to 10 mA, R_M ranges from a few to several hundred ohms. Open S1 and adjust *R1* for full-scale deflection. Now close S1 and adjust *R2* for half-scale deflection. Remove the voltage source, disconnect *R2* and measure it — it has the same value as R_M.

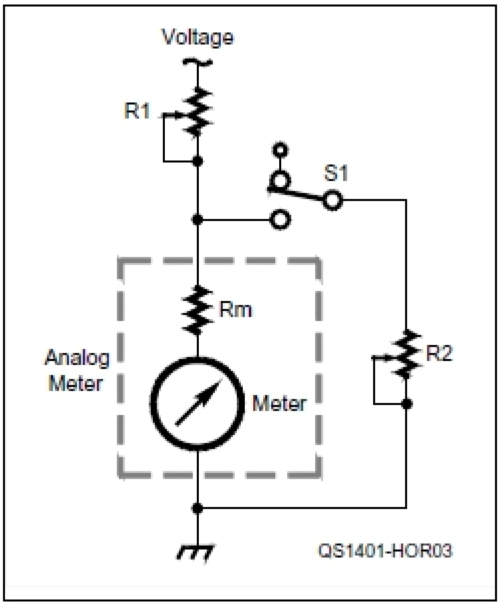

Figure 3 — A circuit for measuring an analog meter's internal resistance. See the text for instructions on its use.

Now that you know the meter's full-scale current and internal resistance, you can calculate the value of shunt resistance necessary for a current, *I*, through the shunt to cause a full-scale reading on the meter:

$R_{SHUNT} = R_M I_{FS} / (I - I_{FS})$ [5]

Let's give this a try: find the shunt resistance that causes a 1 mA full-scale meter with a resistance of 150 Ω to indicate at full-scale with a current of 1 A through the shunt.

$R_{SHUNT} = 150\ \Omega \times 0.001\ A / (1 – 0.001) = 0.150\ \Omega$

Remember that the shunt may have to dissipate some power and that it shouldn't get very warm to avoid temperature effects. A common choice for shunt resistors is a coil of small enameled wire. Tables of resistance for copper wire of different sizes are available in *The ARRL Handbook* and other sources.
[2]

For example, AWG 20 wire can carry 1 A of current and has a resistance of 10.12 Ω per 1000 feet. We'll need 1000 × 0.150 / 10.12 = 14.82 feet of wire to make our shunt.

These circuits are simple but they are everywhere, once you start looking for them. Troubleshooting and circuit design become easier if you understand how they work.

1

All previous Hands-On Radio experiments are available to ARRL members at
www.arrl.org/hands-on-radio
.

2

The ARRL Handbook is available from your ARRL dealer, or from the ARRL Store, ARRL order no. 0007. Telephone toll-free in the US 888-277-5289, or 860-594-0355; fax 860-594-0303;
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Experiment #141 — Window Comparators and Null Detectors

In the early days of Hands-On Radio, one of the first experiments (#11) described comparator circuits. [1]

These are pretty handy for detecting when the voltage at one of their inputs is greater or less than at the other input. I'm finally getting around to describing two variations of the comparator — the *window comparator* and the *null detector*. There are plenty of applications for these circuits in radio and for test circuits.

The Window Comparator

Figure 1 shows a comparator circuit that incorporates hysteresis to shift the setpoint "a little bit" when the output changes state. That helps prevent "chatter" — rapid on-off switching of the output due to noise on the input signal or voltage reference. This comparator-with-hysteresis is known as a *Schmitt trigger*. Schmitt triggers are so useful that logic ICs intended for interfacing to switch and analog signals use them at gate inputs. Typical parts are the CD4093 and CD40106, along with variations of the 7414.

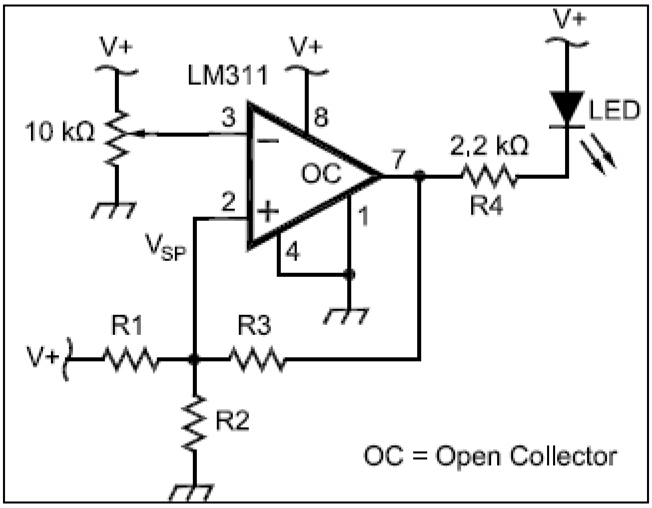


Figure 1 — The Schmitt trigger circuit is a comparator with hysteresis added to its switching setpoint through the addition of R3. See Hands-On Radio Experiment #11 for information on calculating the value of R3.

The comparator provides information about a single level — is the input voltage higher or lower than the setpoint? It is often more useful to know if an input voltage is between an upper and lower limit. This type of comparator is a *window comparator*, sometimes called a *range detector.*

The window comparator in Figure 2 is simply a pair of comparator circuits — one detects whether the input is below the lower limit and the other detects when it is higher than the upper limit. Hysteresis is not included in order to keep the circuit simple. (Power and ground for the LM393 are on Pins 8 and 4 in the DIP and SOIC packages.)

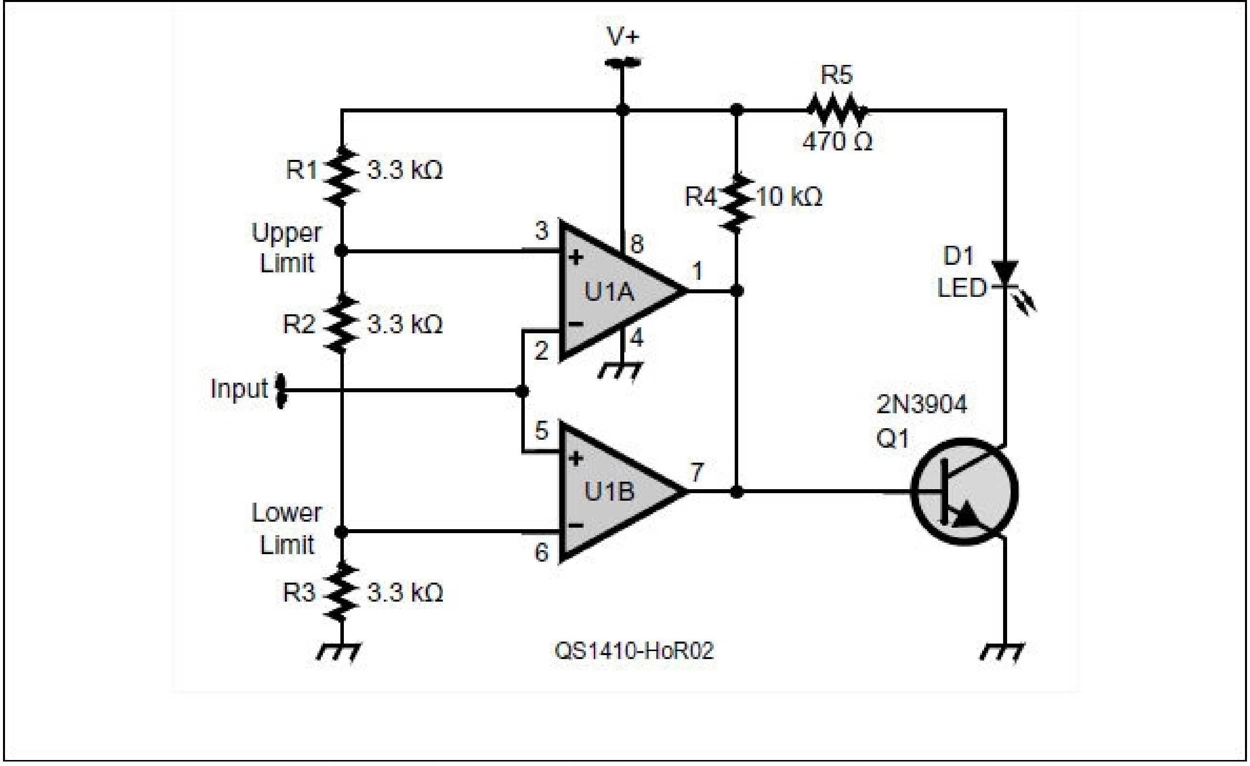


Figure 2 — A simple window comparator circuit using the dual-comparator LM393. The output of both comparators must be HIGH (transistor off) to allow the voltage to rise at R4, turning on Q1 and the LED. This occurs when the input voltage is between Upper Limit and Lower Limit.

The comparator setpoint voltages (labeled Upper Limit and Lower Limit) are easy to calculate: Upper Limit = V+ × (R2+R3)/(R1+R2+R3) = 2/3 V+ and Lower Limit = 1⁄3 V+. If V+ is 12 V, then the two limits are 8 V and 4 V, creating a 4 V window from 4 to 8 V.

Remember that the output of the comparator is an open-collector transistor acting as a switch that can be open or closed. By connecting the output to a pull-up resistor (R4 in Figure 2) the output voltage is **HIGH** when the transistor is off (approximately V+) and **LOW** (at the transistor's collector-emitter saturation voltage) when the transistor is on. When the comparator's non-inverting (+) input is at a higher voltage than the non-inverting (–) input, the output is **HIGH**. When the opposite is true (non-inverting input lower than the inverting input) then the open-collector transistor output is **LOW**.

Connecting both comparator outputs together creates a *wired-AND* gate in which *both* output transistors must be **HIGH** to allow the voltage at R4 to rise and turn on Q1, lighting the LED. If either comparator output is **LOW**, Q1 remains off. For the window comparator, if the input voltage is higher than Upper Limit, the output of U1A is **LOW** and Q1 is off. If the input voltage is lower than Lower Limit, the output of U1B is **LOW** and Q1 is off. The only condition in which both comparator outputs are **HIGH** occurs when the input voltage is between Upper Limit and Lower Limit.

Look In the Window

Build the circuit, using either a variable power supply or adjustable resistor to create the input voltage. Vary the input voltage from below 4 V to above 8 V, while watching the LED. It will turn on when the input voltage is in the target window.

Now experiment with the value of R2 by substituting fixed-value resistors or using an adjustable resistor. What happens to the window range as R2 is increased? (The window range increases.) Why? (A higher value for R2 lowers Lower Limit and increases Upper Limit.) Decrease R2 and verify that the opposite effect occurs. Return R2 to 3.3 kΩ and increase the value of R3 — what happens to the window? (The window shifts higher.) Return R3 to 3.3 kΩ and increase the value of R1 to observe the effect on the window. (It shifts lower.)

If you used a different voltage for V+ the values of Upper Limit and Lower Limit would change right along with it. What if there was noise on the power supply output? Or if you were using a battery supply — what would happen as the battery discharged? Obviously, it's not a good idea to use an unregulated voltage source for your setpoints. Some kind of regulated voltage is necessary. (Note that if you use a bipolar supply with Pin 4 of U1 and the negative end of R3 connected to the negative supply voltage, the window limits can be made positive or negative.)

This is such a common circuit that many manufacturers offer single and dual comparator ICs with a voltage reference built in. One such IC is the Linear Technology LTC1442.[2] You can also use a Zener diode or voltage reference IC. Regardless, for consistent and reliable operation, the voltage source to which the voltage divider string of R1, R2, and R3 must be clean and stable. It is good practice, particularly around RF sources like transmitters, to include small-value capacitors (such as 0.01 µF) from each setpoint to ground.

Window comparators are useful for lots of radio-related chores: making sure your power supply or battery voltage is in the right range, for example. If you sample some RF power and use a peak detector as described in the previous column, a window comparator can be used to make sure the power is within a desired range. Another use is for decoding the Icom transceiver BAND data output that changes from 0 to 7 V with the band selected. K6XX has designed a circuit that uses an LED meter driver IC with many window comparators built in to sort the voltage levels into digital-compatible outputs.[3]

Null Detector

There are many instances in which it's useful to adjust a circuit or system to produce a voltage exactly equal to some reference level, such as when balancing a bridge circuit or trying to adjust dc offset in an amplifier or demodulator circuit. A special type of window detector called a *null detector* responds to a narrow range right around a 0 V difference between its two inputs. "Null" in this case means "no difference between" the voltage being measured and a reference voltage.

Figure 3A shows a null detector circuit that uses an analog center-zero meter to allow fine adjustments to obtain the null. The input section consisting of U1A, R1, and R2 sets the gain of the detector equal to R2 / R1. The higher the gain, the more sensitive the detector. Note that U1 is a dual op-amp and not a dual comparator! A bipolar power supply of at least ±6V should be used.

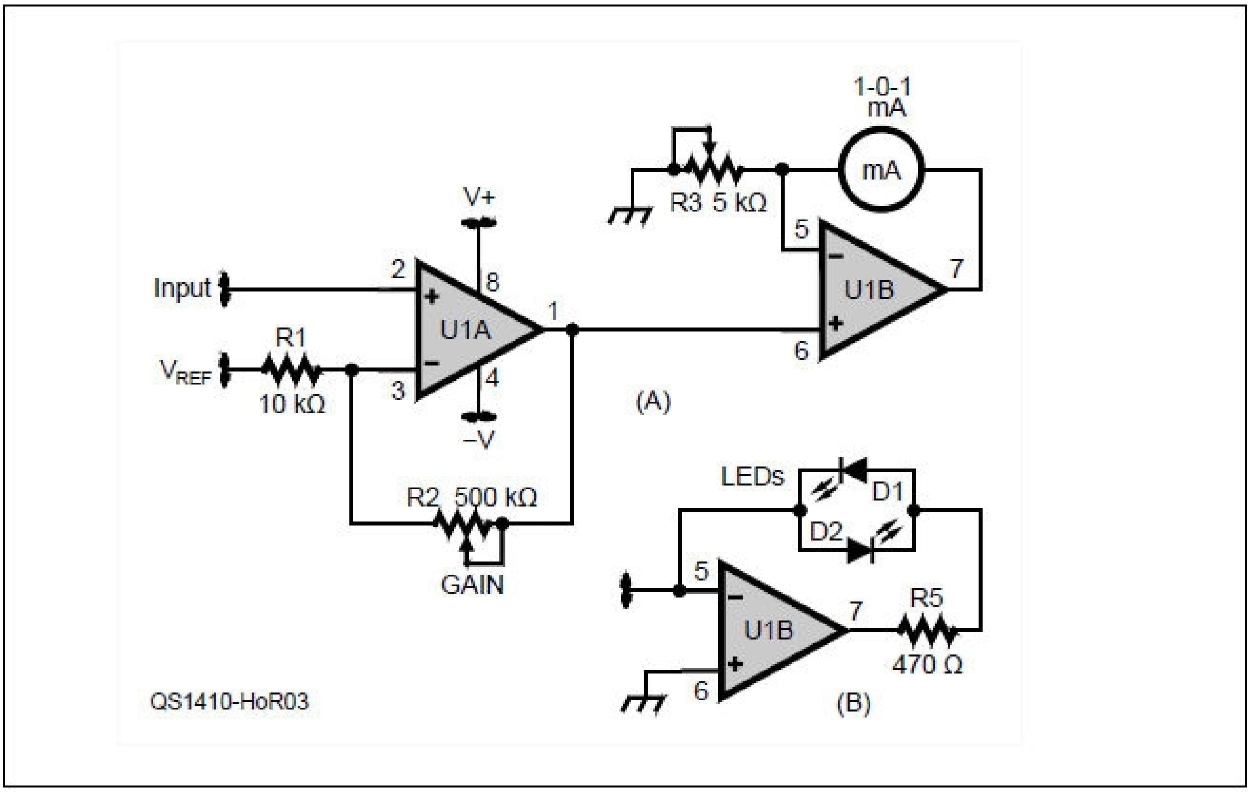


Figure 3 — The null detector circuit converts a positive or negative voltage difference between Vref and Input into a current (3A) or lights an LED (3B) to indicate the polarity of the difference. Both V_{REF} and Input voltages must be ground-referenced.

U1B converts the voltage at the non-inverting input into current through the meter. R3 calibrates the voltage-to-current ratio: I = V / R3. If R3 is 1kΩ, then 1 V at the non-inverting input to U1B is converted to 1 mA of current through the meter. (1-0-1 mA center-zero meters are often used with a shunt for battery charging systems.)

To calibrate the meter; set R2 to 10 kΩ so that gain equals 1, ground V_{ref}, then connect a voltage of 0 to 1 V to the input. Measure the voltage at the output of U1A and adjust R2 so that the meter shows the same number of mA. i.e. – for a voltage of +0.5 V, the meter should indicate 0.5 mA in the positive direction.

To use the null detector, connect V_{ref} to the reference voltage desired, whether 0 V or some other value, and adjust input gain (R2) to the desired sensitivity. For initial adjustments of the external circuit being measured, keep gain low. As the null is reached, increase input gain for more and more sensitive adjustments.

An alternative to the analog meter is shown in Figure 3B, a pair of LEDs connected back-to-back. The calibration pot, R3, is adjusted with V_{ref} and Input shorted so that both LEDs are dark. (Noise, hum, or ripple on either signal may cause the LEDs to light dimly. A 0.1 µF capacitor across R2 creates a low-pass filter to avoid responding to ac components.) If the input voltage is higher than V_{ref}, then the bottom LED will be dark and the top LED will be lit. This is often sufficient to adjust a circuit or balance a bridge without requiring a high-precision display.

Parts List

LM393 dual comparator (or equivalent)

TL082 dual op-amp (or equivalent)

2N3904 NPN transistor

Standard LED (2)

470 Ω, 3.3 kΩ (3 ea.), 10 kΩ (2 ea.) ¼-watt, 5% resistors

5 kΩ, 10 kΩ trimpot

1-0-1 mA center-zero dc milliammeter (optional)

[1] All previous Hands-On Radio experiments are available to ARRL members at www.arrl.org/hands-on-radio.

Experiment #143 — Delay Circuits

The Warner Brothers cartoon character Marvin the Martian seethes, "delays, delays!" But there is no need for a ham radio electronics designer to become "verry annngry." Not at all! Delay circuits are found in many types of ham radio gear, and might even prevent an Earth-shattering ka-boom! This month's column serves up two sample circuits to satisfy your search for spare seconds.

Pulse Stretcher

There are many applications to "stretch" a short pulse into a longer one. A circuit that detects RF might only generate a very short pulse if the incoming signal is brief or weak. That might generate a pulse too short to reliably trigger a logic circuit, be detected by a microprocessor, or light an indicator long enough to be easily seen by the naked eye. Switching or power transients are another notoriously unreliable input signal. One simple and inexpensive solution is to use a spare logic gate or two and an RC timing network, as shown in Figure 1.

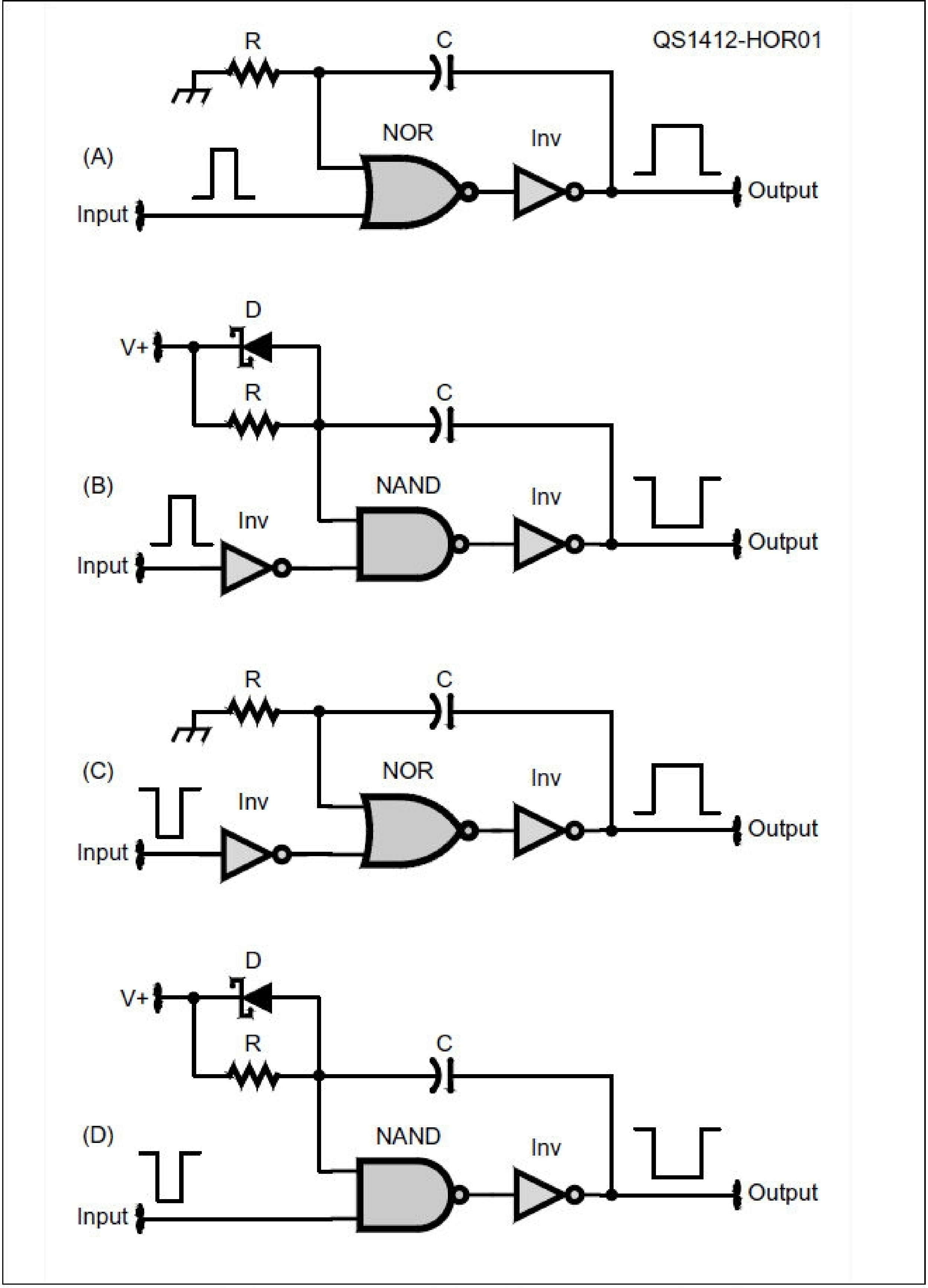


Figure 1 — Four basic pulse stretching circuits. A and D stretch the pulse without inverting it while B and C invert the input pulse. Diodes across the timing resistors prevent switching transients from exceeding the power supply voltage and possibly damaging the logic gates.

Let's take the circuit in Figure 1A as an example. In its resting or *quiescent* state, the input signal is **LOW**, the input connected to R and C is **LOW**, and the output of the **OR** gate formed by the combined **NOR** gate and inverter is also **LOW.** As soon as the input pulse changes to **HIGH**, the output of the **OR** gate also goes high. This causes current to flow through C, creating a voltage across R. Since C is assumed to be discharged, the initial voltage across R is approximately the same as the output signal, close to V+. Then it begins to drop according to the RC circuit's time constant $\tau = RC$. In a bit more than one time constant τ the voltage will have dropped enough at the **OR** gate's input to be a logic-level **LOW**. If the input pulse has ended by then, both inputs to the **OR** gate will be **LOW** and so the output of the **OR** gate will return to **LOW**. If τ is longer than the input pulse, the output pulse has been "stretched" to approximately *RC* seconds. The output pulse will never be shorter than the input pulse. Why? (The output of the **OR** gate will be high as long as either input is high.)

You can follow similar steps to figure out how the circuits work in Figures 1B – 1D. In all, there are four circuits for stretching and inverting either positive- or negative-going pulses. You'll find that using an oscilloscope is the best way of watching both the input and output pulses. Use a 555 timer circuit as described in Experiment #5 as your pulse generator. [1]

The exact amount of stretching depends on the logic switching thresholds of the logic family you are using. The closer the gate switching thresholds are to V+ and ground, the more the pulse will be stretched. For example, switching thresholds for the 4000-series of CMOS logic are about 10 and 90 percent of V+. Pulses will be stretched longer for this family of logic than for a logic family with thresholds closer to ½ V+. Why? (Because the voltage across R will have to decay longer to reach the lower switching threshold and that means the output pulse will stay **HIGH** longer.)

Soft-start Circuit

Linear amplifiers and other equipment with high-voltage (HV) power supplies need a bit of delay between the time the power switch is turned **ON** and the time full voltage is applied to the HV rectifiers and filter. The reason is *inrush current*. If a linear power supply's filter capacitors are discharged when power is applied, they act like a short-circuit during those first few cycles of rectified ac. This causes very high current pulses in the transformer windings and through the rectifiers as the capacitors charge up.

After a few cycles of ac and depending on the resistance of the rectifiers and transformer secondary winding, the filter capacitors are charged to near their peak value and the amount of current drops dramatically. During the charging period however, peak currents can be 10 to 20 times normal current or even higher, placing significant stress on all of the power supply components.

Circuits have been employed that slowly increase transformer input voltage with a **TRIAC** or other variable ac source. The *soft-start* circuit presented in Figure 2 is simpler and satisfies quite nicely the requirement to limit that surge current. It limits inrush current with a 10 Ω power resistor. Until the relay activates, the 10 Ω resistor is in series with the primary winding of the main power transformer. After a suitable delay, the relay contacts short out the 10 Ω resistor and full power is applied to the main transformer.

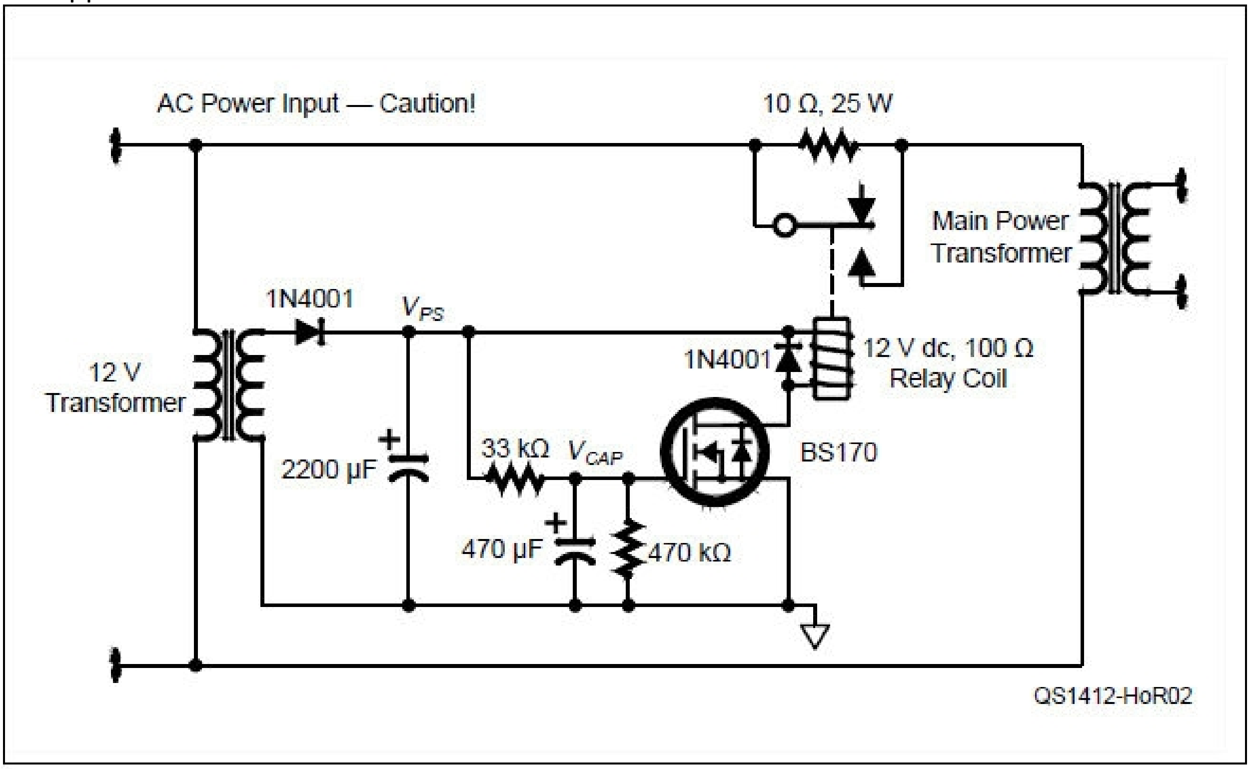


Figure 2 — A soft-start circuit reduces inrush current to the large filter capacitors in a high-voltage power supply. The 10 Ω resistor limits primary current while the capacitors charge. After a couple of seconds, the relay shorts out the resistor and applies full power to the secondary.

To power the relay, an auxiliary power supply circuit is required. A small 12 V transformer supplies a 1N4001 diode and 2200 µF in a half-wave rectifier circuit. At light loads, the filtered output voltage, V_{PS}, will be about $12 \times 1.4 - 0.7 \approx 16$ V. (V_{PS} will drop closer to 12 V when the relay coil draws current from the supply.)

The timing of when the relay switches is determined by the 33 k Ω resistor and 470 µF capacitor. When power is applied with the 470 µF capacitor completely discharged, it begins charging towards 16 V with a time constant of $\tau = RC = 33 \times 10^{3} \times 470 \times 10^{-6} = 15.5$ s.

As the voltage across the 470 µF capacitor (V_{CAP}) increases, it approaches the *gate threshold voltage* ($V_{GS(Th)}$) of the BS170 FET. This is the point at which the FET will rapidly turn on and conduct drain current, acting more or less like a switch. How quickly the gate voltage will reach 2 V is determined by the equation:

$$t = -\tau \ln[1 - \frac{V(t)}{V_{PS}}] = -15.5 \ln[1 - \frac{2}{16}] \approx 2\,\mathrm{s}$$

Two seconds is plenty of time for a power supply's filter capacitors to charge up. (This and similar formulas for RC timing circuits are posted on the Hands-On Radio web page for this experiment.)

Most 12 V relay coils have a resistance of around 100 Ω and so draw about 120 mA from a 12 V supply. If the *on-resistance* ($R_{DS(On)}$) of the FET is a few ohms, it will dissipate $P_D = I_D^2\ R_{DS(On)} = 0.014\ R_{DS(On)}$ watts, which is a minimal amount of heat. The 470 kΩ resistor discharges the timing capacitor when power is removed so that the circuit will operate properly when power is again switched on. The second 1N4001 diode clamps the coil voltage so that a nasty *inductive kickback* transient doesn't destroy the FET when the relay is turned off.

All of this is quite loosely estimated, which means there is plenty of room for experimentation by you! There is a lot of variation between relays — not only in the coil's resistance but the relay's *pull-in voltage* at which the coil will actually switch the contacts. Try changing the timing components, use different types of FETs or redesign the circuit to use an NPN transistor like the 2N4401, or maybe scrounge up some different relays and try them out. This will help you get a feel for how much variation you can expect out there in the real world.

You needn't actually apply the soft-start circuit to a high-voltage supply — it will operate just fine on its own. You can get a sense for the timing just by listening to the relay or wiring an LED circuit through the normally-open contacts. If you use this circuit on an actual ac power supply, though, please remember that the 10 Ω resistor and the relay contacts carry the full ac mains voltage. That hazard is easy to forget when working with low-voltage circuits! Make sure you keep the ac wiring insulated and well away from the low-voltage dc circuits. No surprises, please!

1

All previous Hands-On Radio experiments are available to ARRL members at www.arrl.org/hands-on-radio.

Experiment #155 — Negative-Voltage Circuits

It may come as a surprise in this 12-volt world, but not every circuit is powered by 13.8 ± 0.5 V dc. Negative voltages were quite common in the days of tubes, but today's digital logic and analog circuitry mostly requires only one voltage polarity — positive. Is there nothing left "below ground?"

ALC Control

You don't have very far to look for a circuit that depends on a negative voltage, and that is your radio's Automatic Level Control (ALC) input. Just like the radio's internal ALC system, an external amplifier generates an ALC signal to reduce transmitter power and prevent overdrive. The ALC signal is a dc voltage varying from 0 V (full power, no ALC action) to a few volts negative. For example, the ALC range of my TS-590S is 0 to –7 V. Armed with this knowledge, you can turn your 100 W radio into a milli-watter with the simple battery-powered circuit shown in Figure 1.

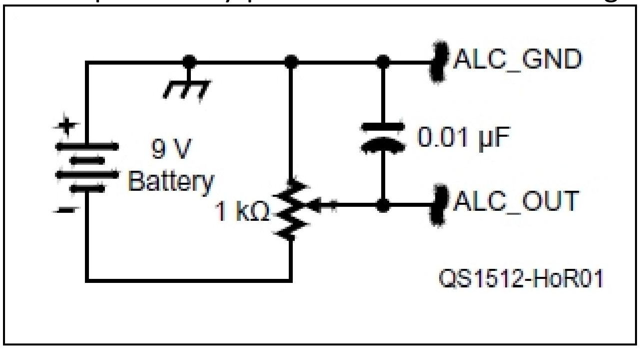


Figure 1 — This simple battery-powered resistive divider can be used to control the output power of radios with a standard, negative-voltage ALC input. Using the ALC input allows output power to be reduced more than with the front-panel RF PWR control.

Set the 1 kΩ pot to about 500 Ω between the wiper and either end of the element. Connect a 9 V battery (paying careful attention to which terminal is positive or negative) across the element of the pot. (You can't use a ground-referenced power supply for the 9 V battery!) The output ALC signal is created between the wiper of the pot and positive terminal of the battery, which is connected to the radio's chassis ground. Use a voltmeter to verify that the **ALC_Out** signal is negative and about half of the battery voltage, then adjust the pot for an output of about –0.1 V. The 0.01 µF capacitor filters out any RF that might be picked up by the connecting cable.

Disconnect the battery and fashion a cable to connect the **ALC_Out** signal to your radio's ALC input. (This might be a dedicated phono jack, but it's probably two pins on an accessory connector.) Be sure the positive battery terminal will be connected to the radio's common or ground pin. Reconnect the battery and turn on the radio. Set the radio for full power output, then generate a steady carrier level into a dummy load. Adjust the pot so that **ALC_Out** becomes more negative and output power from the radio should eventually drop. Generate some CW and adjust the pot until you have 5 W output. *Voilà* — you have a QRP rig. Or, you can keep going and see what you can work with 100 mW!

Un-common Emitter

The common emitter (CE) amplifier circuit from "Hands-On Radio" Experiment #1 makes another appearance in Figure 2 — almost!

1

Look closely and you'll see that Q1 is a 2N3906, the PNP twin of the common 2N3904. They have very similar specifications for gain, voltage and current rating, switching speed, etc. You'll also notice that the battery polarity is inverted so the "top rail" is the circuit's common.

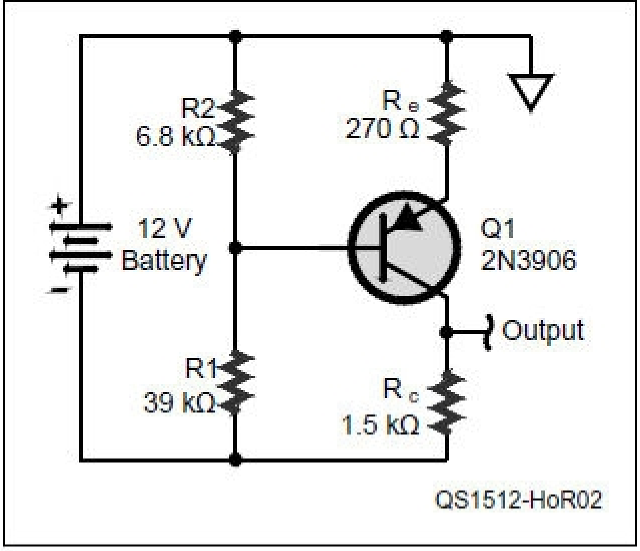


Figure 2 — By connecting a battery as a –12 V supply, the circuit should produce I_C = –4 mA (the minus sign indicating current flowing out of the collector electrode) and V_{CE} = –5 V. Most silicon PNP transistors with dc current gain (beta) of 50 or more will produce similar results.

Everything about this PNP version of the CE is inverted. Instead of pushing current *into* the base to turn on the transistor, in this circuit you pull current *out* of the base. The more current you allow to flow out of the base (by lowering the value of R1), the more I_C will flow. Even the equations for gain and dc operating point are the same.

Wire up this circuit (use a battery or isolated power supply) and verify that I_C is about –4 mA and the voltage from collector to emitter, $V_{CE,}$ is about –5 V. (The minus sign for I_C indicates that current is coming *out of* instead of *into* the collector.) You can experiment with resistor values to see the effect changing R1 and R2 has on I_C. In fact, building both the NPN CE circuit and the PNP CE circuit side by side is a good way to get more comfortable with PNP transistors.

Grid-block Keying

Many of us have electronic keyers that ground our rig's positive-voltage keying input to turn on the transmitter. What about keying a vacuum tube rig? Many later model tube rigs used *grid-block keying*, in which a high negative bias voltage (–50 to –150 V) on a tube's grid cut off plate current during non-transmitting periods. Grounding the negative voltage allowed plate current to flow, turning on the transmitter. This was no problem with a hand key, "bug," or relay, but today's solid-state keyers can't handle negative voltage. What to do?

Chuck Olson, WB9KZY, of Jackson Harbor Press (
wb9kzy.com/ham.htm
), designed the circuit shown in Figure 3 to adapt the common positive-voltage keyer to grid-block vacuum tube circuits.
[2]

In this circuit, the keyer's output (usually an NPN transistor, as shown) is connected between J1 and J3. To key the tube transmitter, the keyer's output transistor grounds the base of Q1 through R1, turning on the PNP MJE350, which is rated for the higher grid-block voltage. This connects J2, the radio's grid-block input, to +5 V, turning on the transmitter. (Voltage at J2 is assumed to be negative.) R2 is a pull-up resistor that keeps the MJE350 turned off when the keyer transistor is off. C1 and D1 protect the transistor from transients and keep RF out of the circuit.

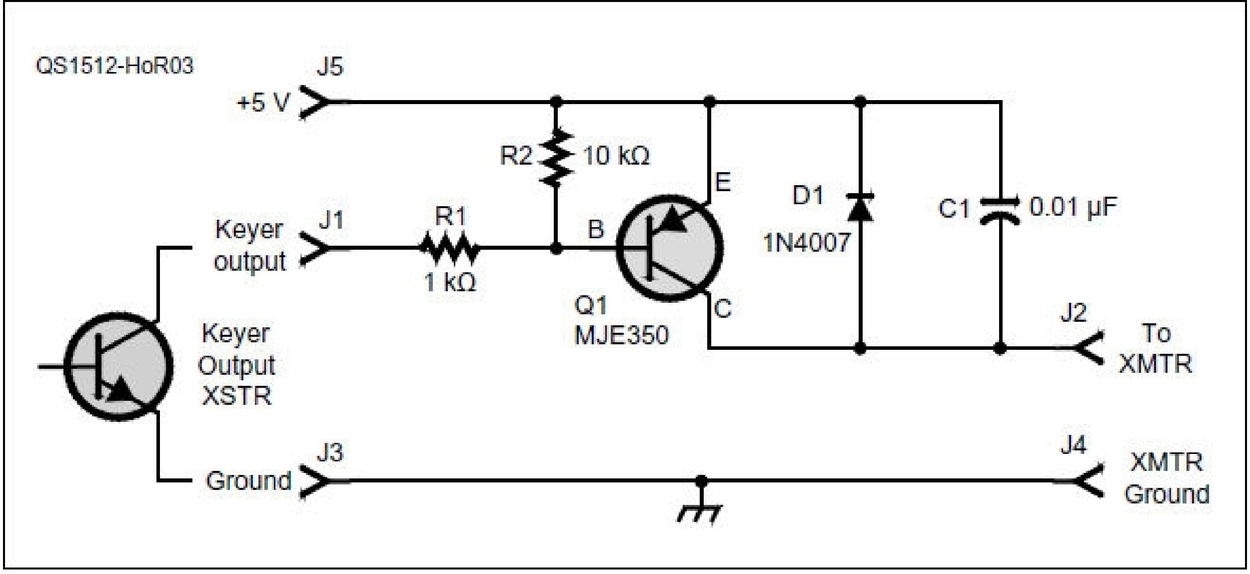


Figure 3 — This adapter circuit allows output of a regular electronic keyer designed for +12 V radios to key tube transmitters with negative-voltage grid-block keying. The grid-block keying line is connected to +5 V through the MJE350 when the output transistor of the electronic keyer turns on. [Circuit provided courtesy of Jackson Harbor Press]

Bipolar Op-Amp Circuits

Another place you'll find a requirement for a negative voltage supply is in op-amp circuits, such as the one shown in Figure 4. This circuit is from "Hands-On Radio" Experiment #3 — a *difference amplifier* with an output equal to the difference between V_{in1} and V_{in2}. (All the resistor values must be equal for this to be true.) While the venerable 741 is shown on the schematic, any common op-amp will do the job at audio frequencies or below. The inputs can be dc, ac, or a combination of the two. It doesn't matter if both input voltages are positive or negative or which is greater. (Just keep them between the power supply voltages to avoid damaging the op-amp.)

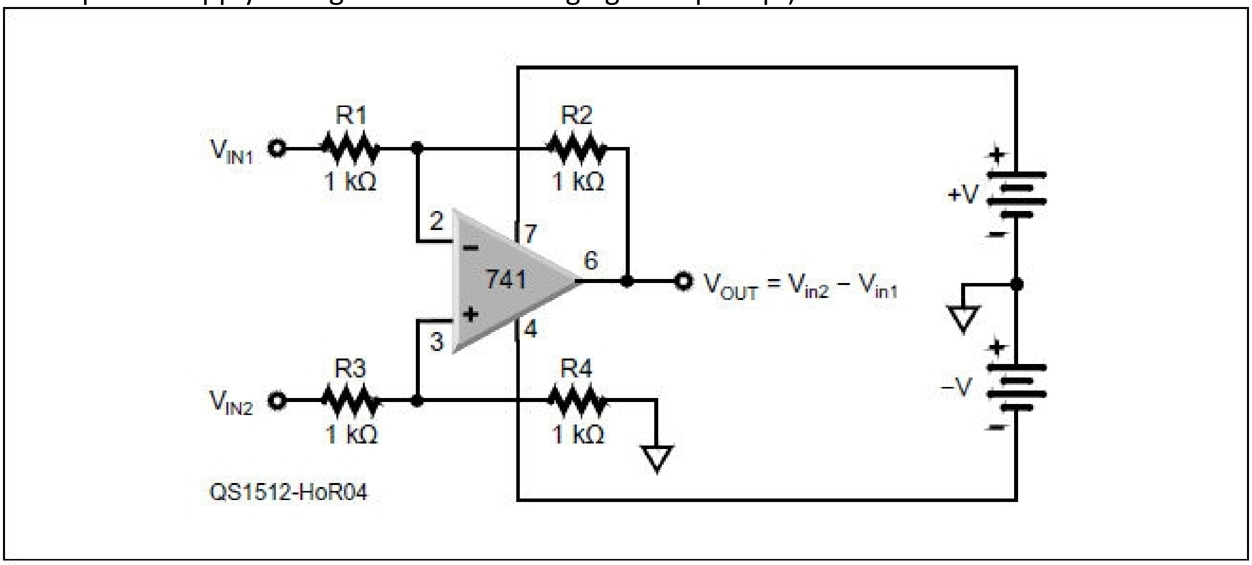


Figure 4 — The difference amplifier is an example of op-amp circuits that require a negative voltage. You can generate the negative voltage with a bipolar power supply, a string of batteries, or a voltage converter IC as described in the text.

The output can be of either polarity so the power supply must supply both polarities. Figure 4 also shows a handy (and portable) means of generating a negative supply — splitting a set of batteries. A string of D-cell batteries will supply clean dc power for a long time at the low loads of most circuit experimenting. Use at least four cells for each power supply voltage to generate ±6 V. If you have a large multi-cell holder, you can create the circuit's common voltage by soldering a wire to a bit of brass or copper shim and slipping it between adjacent cells. A pair of surplus rechargeable gel-cell batteries is a good choice, too.

Generating Negative Voltages

Batteries are terrific, but over the long-term, even a small current drain will deplete a good-sized battery. If you have a reliable source of positive voltage you can use a voltage converter IC and a couple of capacitors to create your negative supply instead.

The ICL7660 has been a staple of the analog designer for a long time, along with its many variations. Start by downloading and reading the ICL7660 data sheet from Maxim Integrated.
[3]

Not only will you get a detailed description of how the IC works, but also plenty of interesting and innovative ways to use it.

The ICL7660 uses *switched-capacitor* technology to convert positive voltages to negative voltages, double a positive voltage, divide voltages, can be connected in series to create higher negative voltages, or connected in parallel to increase the available current.

If you build a simple negative voltage converter, it can be used as the power supply for the ALC power control circuit in Figure 1. Input power can be obtained from the radio's +12 V supply. Place a Zener diode across the circuit's output, in parallel with the 0.01 µF capacitor to limit the output voltage to the maximum for your radio. By packaging the circuit in a simple plastic container or enclosure, you create a handy power-control accessory for your station. Just right for the QRP ARCI's "1000 Miles Per Watt" award — how low can you go?[4]

Notes

1 All previous "Hands-On Radio" columns are available to ARRL members at www.arrl.org/hands-on-radio.

2 Jackson Harbor Press sells the Keyall and KeyallHV kits if you prefer to purchase the components and pre-made PCB to make your own adapter.

3 datasheets.maximintegrated.com/en/ds/ICL7660-MAX1044.pdf

4 qrparci.org/awards

Experiment #156 — Designing a Broadcast Reject Filter

If you live within a few miles of an AM broadcast station, you will be interested in broadcast-reject filters to prevent overload, this month's subject. This will actually be a "two-fer" column because we're going to use the latest version of Jim Tonne's, W4ENE, *ELSIE* filter design software to do the design work. This handy software makes it easy to design your own passive LC filters in a wide variety of configurations. You can juggle design inputs to your heart's content, watching the filter respond until it's just right.

Downloading *ELSIE*

Begin by downloading the set of utility programs from *The ARRL Handbook*'s web page.[1]

Open the ZIP file, double-click on *LCinstall275.exe* and after installing, run *ELSIE*. Be sure you are running version 2.75 (look at the lower left of the opening display) so that the directions in this article will agree with what you see on your screen. Begin by clicking **NEW DESIGN**.

Specifying the Filter

Now it's time to tell *ELSIE* what kind of circuit you want. This is the filter's *topology* describing the general arrangement of the filter components. (Filter basics were covered in "Hands-On Radio" experiments #50 and #51.[2]) Because we are designing a broadcast-reject filter for 160 meter reception, we will want a high-pass response. *ELSIE* gives us two choices for high-pass filters: capacitive input and inductive input. A capacitor in series with the filter at the input (capacitive input) blocks any dc and low-frequency signals, so select that topology.

Next, we must select from the many types of LC filter circuits, called *families*, and each has a slightly different set of characteristics. For example, the Butterworth family has a smooth rolloff between the passband and the stop band. The Chebyshev family allows some ripple in either passband or stop band in trade for a steeper rolloff. Bessel filters have a constant time delay through the filter in the passband. If you click the **?** button next to **BUTTERWORTH** in the **FAMILY** section, a pop-up window will show the general behavior for each family.

In our case, we need a very sharp rolloff, passing signals with little attenuation at 1.8 MHz, the bottom end of 160 meters, but lots of attenuation at 1.6 MHz, the highest frequency of the AM broadcast band at which full-power stations are permitted.[3]

Chebyshev would be a good choice, but the Cauer family is even better at creating the necessary steep rolloff. The tradeoff apparent from the filter characteristics is that attenuation of the Cauer filters varies quite a bit in the stop band. That's okay, as long as a certain minimum attenuation is maintained, so select **Cauer** as the filter family.

Now the program needs some performance specifications entered at the right-hand side of the screen. How much attenuation (**STOP BAND DEPTH, A_S**, in dB) is enough for our filter? In my experience, 40 dB is enough to keep even nearby AM stations from clobbering the front end of a late-model receiver. For **RIPPLE BANDWIDTH (F_C)**, enter "1.8M" (1.8 MHz) as the lowest frequency of the filter passband. The highest frequency at which we want our 40 dB of attenuation, "1.6M" (1.6 MHz), is the **STOPBAND WIDTH (F_S)**.

Filter **ORDER (N)** can be thought of as the number of resonances created by the filter components. The higher the order, the more components (Ls and Cs) are required to create the circuit. Start by entering a filter order of 3 to see if we can meet our design goals.

Viewing the Response

To set up the program's calculations and display configuration, click the **ANALYSIS** tab. This is where the inductor and capacitor Q are specified (250 and 1000, respectively). Lower values of Q results in less stop band attenuation and less sharp rolloff, among other effects. Leave them at their default settings for this exercise. Specify an **ANALYSIS START FREQUENCY** of 0.5 MHz and an **ANALYSIS STOP FREQUENCY** of 10 MHz. Leave all other selections and values at their original default settings.

Before looking at the filter response plot, click the **SCHEMATIC** tab at the top of the screen. The information is summarized in Figure 1. Passband ripple (21.3 dB) and the component maximum/minimum ratios are calculated by the program.

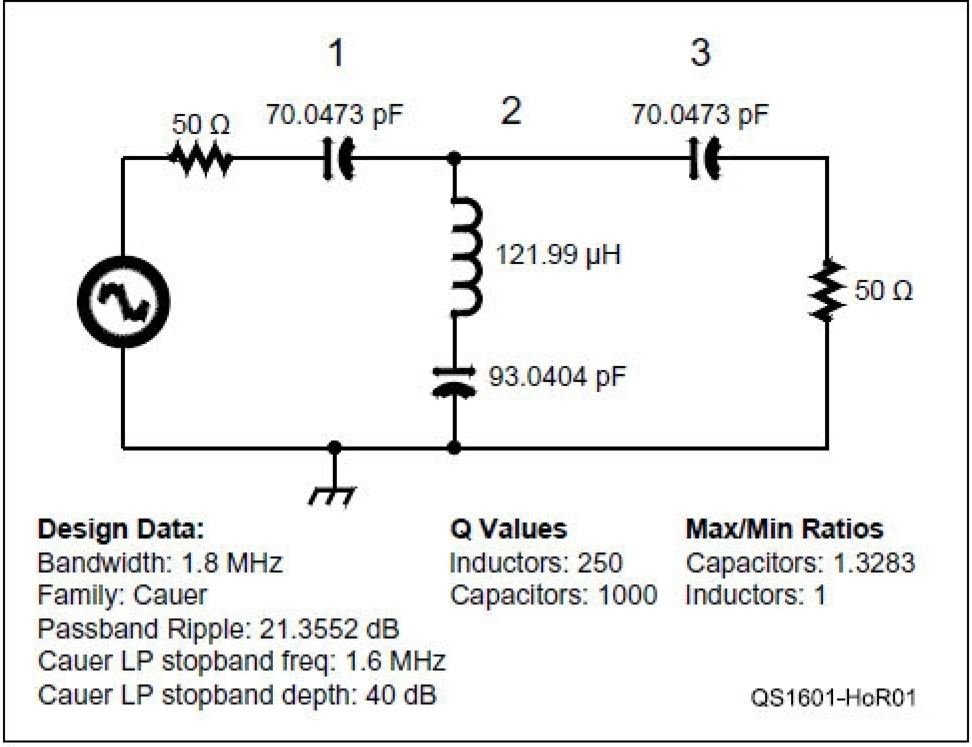


Figure 1 — The beginning third-order broadcast-reject filter schematic and the design parameters used or calculated by the *ELSIE* program.

Now click the **PLOT** tab for the frequency response graph of Figure 2. Place the cursor on the blue response line and hold down the left mouse button. At the bottom of the screen, you will see the filter performance at that frequency. The figure shows performance at the response peak of 1.93 MHz. Move the mouse to the stop band notch near 1.5 MHz to find attenuation there (–69 dB at 1.49 MHz). Attenuation varies by more than 20 dB over the 160 meter band — that's too much! The program is trying but can't meet our specifications without more components to create a higher-order filter. (While you're at it, use the **DESIGN** window to select different filter families and compare their responses.)

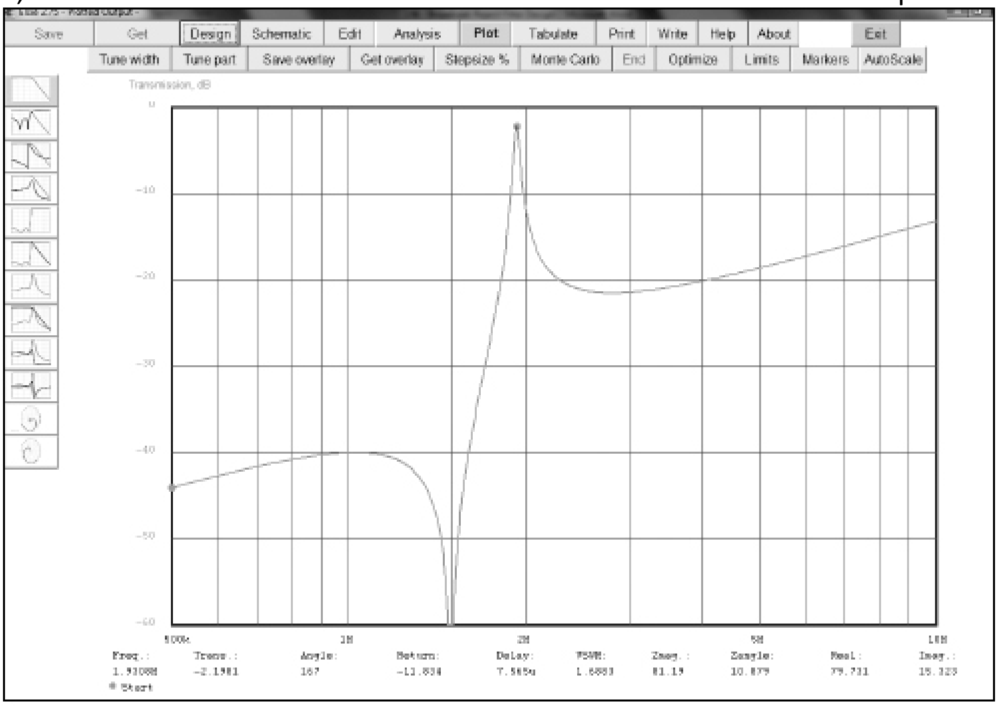


Figure 2 — Frequency response of the third-order filter shows too much variation in attenuation across the 1.8 – 2.0 MHz 160 meter band and does not meet the 40 dB attenuation requirement at 1.6 MHz.

Interactive Design

This is where the value of easy-to-use design software becomes apparent. Instead of re-starting a laborious design process, simply re-enter the new specifications and try again. Return to the **DESIGN** tab and increase the filter order from 3 to 4, then click **PLOT**. Performance is improved, but attenuation still varies by more than 10 dB across 160 meters and the rolloff isn't sharp enough; only 28 dB of attenuation at 1.6 MHz.

Increase the filter order to 5, resetting F_S to 1.6 MHz. (The program changes some values when order is changed between odd and even. Check the settings when order is changed.)

This response (see Figure 3) is much more useful. Attenuation varies by about 3 dB (1⁄2 S-unit) across the 160 meter band and we just meet our design goal with –40 dB of attenuation at 1.59 MHz. Click the **Save** tab to hold on to this design version before proceeding.

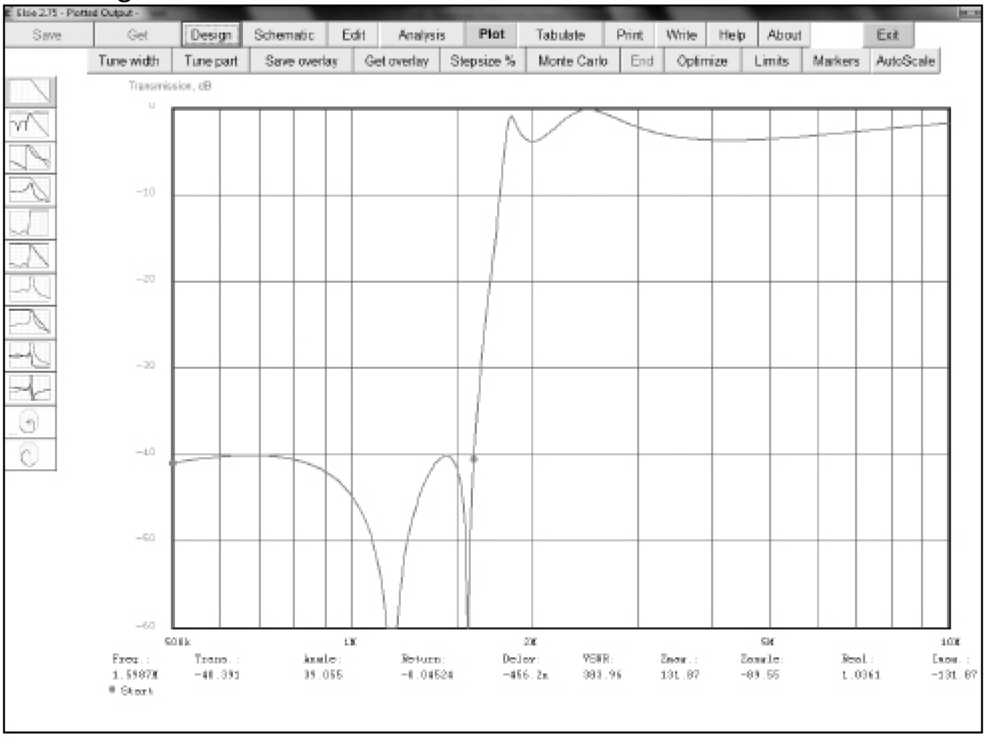


Figure 3 — Frequency response of the fifth-order filter meets the broadcast band rejection requirement with only 3 dB of variation in the 160 meter band.

Using Standard Value Components

The schematic shows all of the component values are in a reasonable range. Never-theless, I don't think your local component vendor will have, say, 538.436 pF capacitors in stock, nor do you want to have to adjust variable capacitors. Now is the time to redesign the filter using standard fixed-value parts. This will degrade filter performance a bit, but remember that we can continue to work with the design.

Return to the **DESIGN** window and click the **NEAREST 5%** tab. You'll be presented with several options, starting with changing all of the components to the nearest standard value in the 5% series. Other options include just changing the capacitors or inductors, assuming you'll wind the Ls or tune the Cs. You can also change just the capacitors or inductors and the program will re-calculate the remaining values exactly so that you can tune up the filter yourself.

Let's take the easy way out and select the first option to use all standard values. Return to the **DESIGN** tab, then check the schematic shown in Figure 4. How about performance? Viewing the frequency response, not much has changed. We have a little more variation across the band (now 4 dB) but attenuation at 1.6 MHz is still the same, only failing to reach 40 dB below 840 kHz by less than a dB.

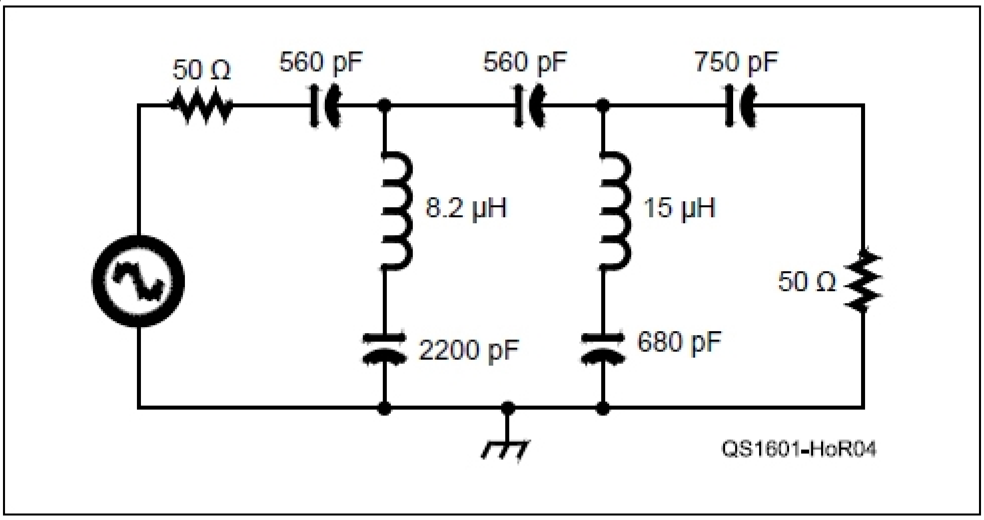


Figure 4 — The schematic of the fifth-order filter after standard 5%-series component values are substituted for exact calculated values. Filter performance is substantially the same as the exact-value version.

I think we're done! It's easy to see how you could continue to experiment with the filter, possibly trading some stop band attenuation for less passband attenuation, or smoother response in the passband, and we haven't even begun to work on the input and output impedances or delay time. Nevertheless, this design can be built with off-the-shelf components requiring no tuning to provide useful performance. Thanks to W4ENE and his terrific software.

Notes

1

A set of free student version circuit design utilities from Tonnesoft are available for download from www.arrl.org/arrl-handbook-reference.

2

All previous “Hands-On Radio” columns are available to ARRL members at www.arrl.org/hands-on-radio.

3

Above 1600 kHz in the US, AM stations are limited to 10 kW during the day and 1 kW at night. These smaller stations are less likely to cause overload problems than the full-power 50 kW transmitters.

Experiment #164 — Dividers

Dividers — they seem like such trivial things. That is, until you need one and then the mad scramble through the reference books begins. There are electronic dividers for every aspect of a signal; voltage, current, power, phase, frequency. We'll take a look at voltage and current dividers this month. You should have these circuits tucked away in your radio toolbox, ready for use wherever the need arises.

Passive Voltage and Current Dividers

The passive version of voltage and current dividers in Figures 1 and 2 are about the simplest resistive circuits around. The equation for output voltage across R_3 (the load resistor) as a fraction of the input voltage, V_{IN} is:

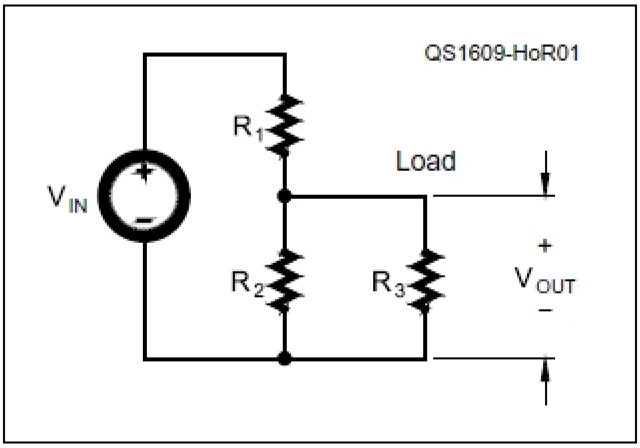


Figure 1 — A resistive voltage divider with R3 representing the output load.

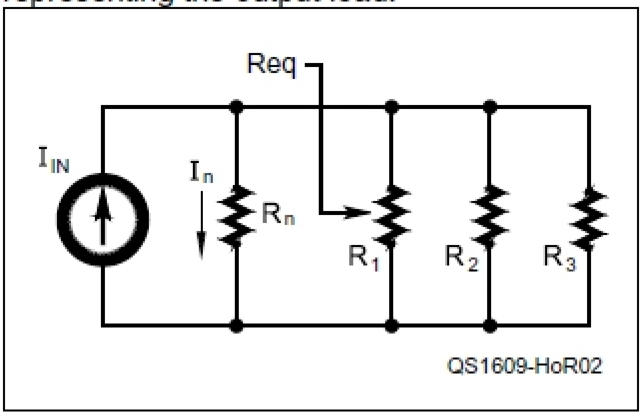


Figure 2 — A resistive current divider with four resistors.

$$V_{OUT} / V_{IN} = \frac{R_2 \parallel R_3}{R_1 + (R_2 \parallel R_3)}$$

where the symbol || represents "in parallel with." Note that you have to take into account the loading effect of R_3 in parallel with R_2 to calculate the exact division ratio.

How much does the loading effect matter? Well, that also depends on the ratio of R_1 and R_2, but let's take a typical situation with R_1 (the divider input resistor) being 10 × R_2. Varying the ratio of R_3 to R_2 from 0.1 (R_3 = 10% of R_2) to infinite (R_3 = an open circuit), the voltage divide ratio varies from 0.009 to 0.09, which is a 10:1 range! So yes, it matters. (An *Excel* spreadsheet showing how the ratios R_1:R_2 and R_3:R_2 affect V_{OUT}/V_{IN} is available on the "Hands-On Radio" web page for this experiment.

[1]

) With this ratio of R_1:R_2, R_3 must be 10×R_2 or more to cause a voltage division error of less than 10%.

Where the voltage divider is essentially a series circuit, the current divider in Figure 2 is a parallel circuit. The figure shows a four-resistor divider, but any number of resistors, *n*, can be used. The current through resistor R_n, is:

$I_n / I_{IN} = R_{eq} / (R_n + R_{eq})$

where R_{eq} is the parallel value of all resistors *other* than R_n. In Figure 2, $R_{eq} = R_1 \parallel R_2 \parallel R_3$.

Active Voltage Dividers

Op-amps are great for building circuits that perform analog math operations. In fact, that's what the "op" in "op-amp" actually means — "operational." The inexpensive IC amplifiers you buy for pennies today are the descendants of the original amplifiers used in analog computers to solve all sorts of hard problems.

The circuit in Figure 3 performs the operation of multiplication or division with the ratio determined by the ratio of the two resistors, R_i and R_f. (See Experiment #3 for an explanation of how the circuit works.)

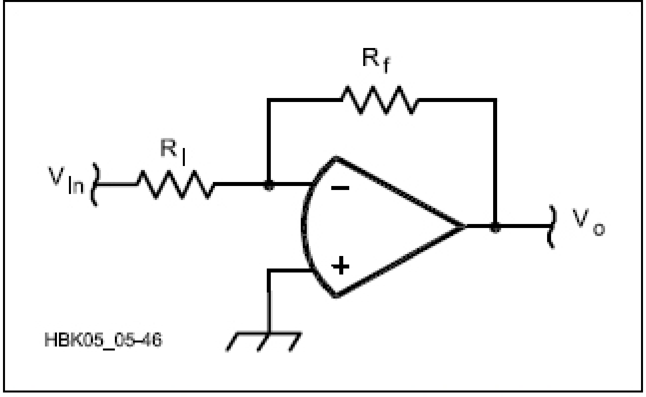


Figure 3 — An inverting amplifier circuit that can perform multiplication or division of the input voltage, V_{in}.

add an inverting buffer circuit to get rid of the minus sign and you have a multiplier or divider, depending on whether the value of R_f is larger or smaller than that of R_i.

Active Current Dividers

You may have seen the circuit in Figure 4A before — the current mirror. When the transistors Q1 and Q2 are closely matched, the collector current through Q2 will equal or "mirror" the current in Q1. Current mirrors work because matched bipolar junction transistors with the same base-to-emitter voltage will have the same collector currents. Because the bases and emitters are connected together, V_{BE} must be the same for both transistors and so are the collector currents.

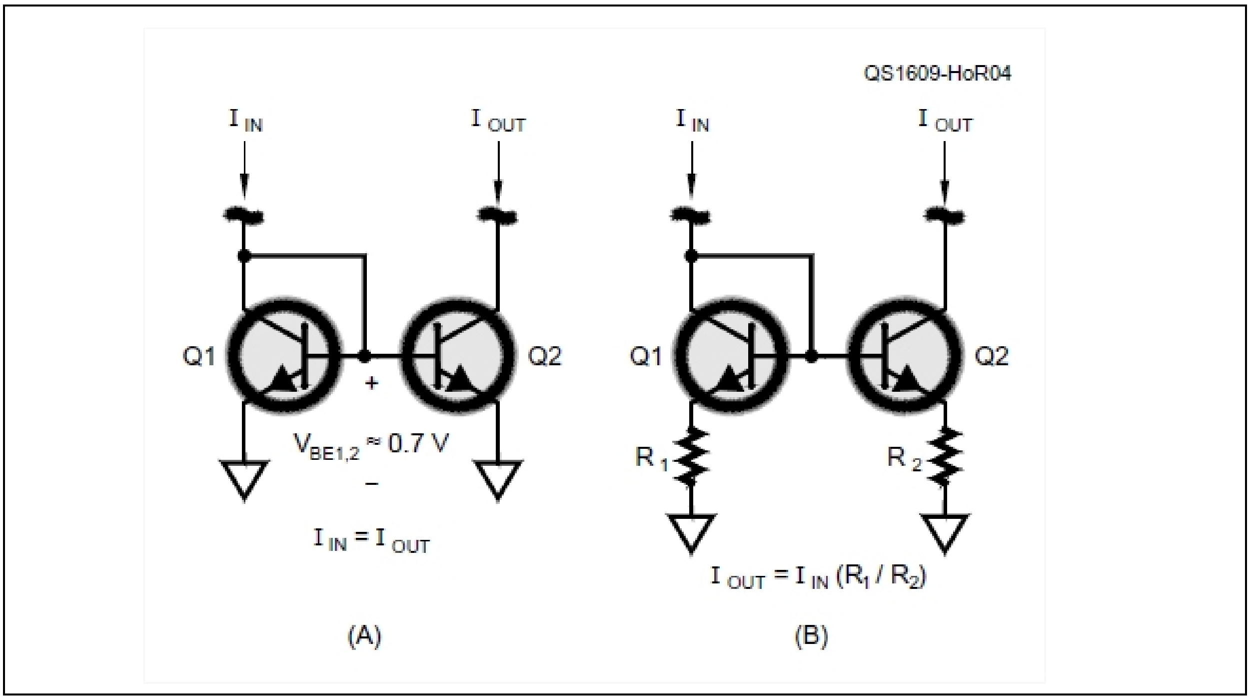


Figure 4 — A current mirror (A) can be turned into a current divider (B) by using two resistors in the transistor emitter circuits.

The two transistors are matched if they are made out of the same material (Si, Ge, GaAs, etc), have the same current gains (β), and are at the same temperature. Without careful measurements, it's hard to find two individual transistors that are precisely matched. If the transistors are part of the same IC, however, they will be very closely matched. For example, the MPQ2222 contains four 2N2222 transistors in either DIP or SMT packages.

The current mirror can be turned into a current divider by adding resistance in the emitter circuit of each transistor. With the bases connected together, the voltage from the bases to common is equal for both transistors:

$I_n \ / \ I_{IN} \ = \ R_{eq} \ / (R_n \ + \ R_{eq})$

If $V_{BE1}=V_{BE2}$, then the transistor currents are controlled by the ratio of the two resistors: $I_{IN}/I_{OUT} = R_2/R_1$ and $I_{OUT} = I_{IN}(R_1/R_2)$. In the truest sense, this circuit doesn't divide a single current into two individual currents, but it does create an output current that is a known and controllable fraction of the first.

Probing the Effect of Reactance

There is no requirement that the voltage divider in Figure 1 be constructed only from resistors. Reactances will do nicely, and the Colpitts oscillator relies on a capacitive voltage divider in its feedback circuit. (See Experiment #34, "RF Oscillators, Part 1," for a description of the Colpitts circuit.) If reactances and resistances are mixed, there will be a phase shift between the input and output voltage that depends on frequency. If inductive and capacitive reactances are used, the result is a series tuned circuit and a resonance is created at the frequency where the reactances are equal, creating equal and opposite phase shifts that cancel, as well.

In keeping with our recent theme of oscilloscope-related ideas, a voltage divider with reactance describes the combination of a scope input connected to a 10× probe. Figure 5 shows the scope's input has approximately 20 pF of parallel capacitance and the probe has a 10× smaller adjustable capacitance of 2 pF in parallel with its series resistance.

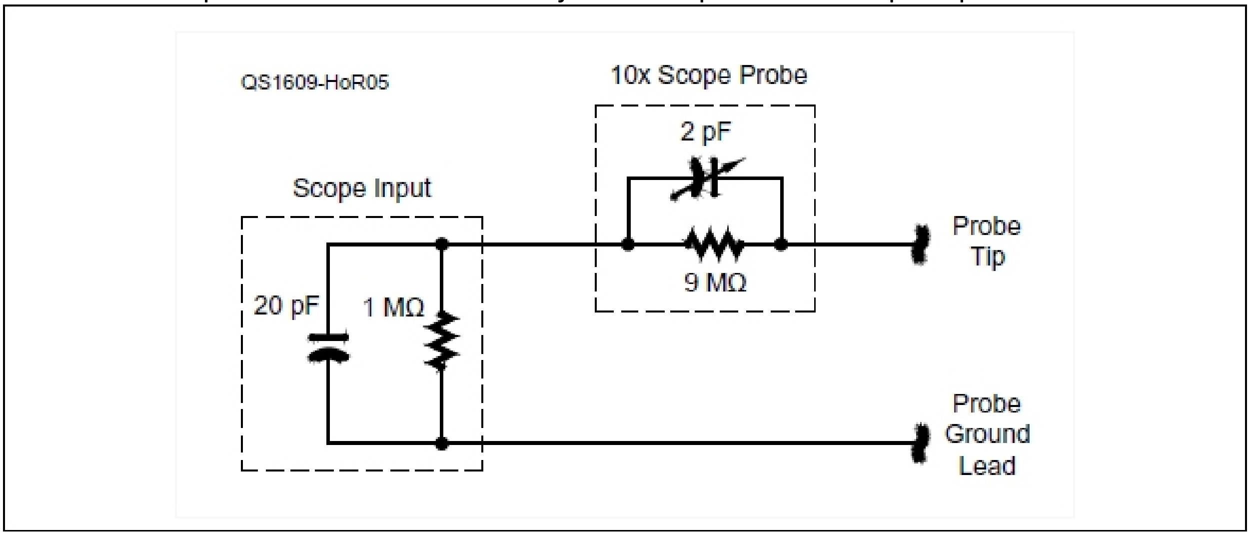


Figure 5 — The equivalent circuits of a typical oscilloscope input and probe. The probe's adjustable compensation capacitor allows the low-pass effect of the input capacitance to be balanced out by the probe capacitance's high-pass effect.

While it is expected that the scope's high-impedance input might include some capacitance, what is the function of the small adjustable capacitance in the probe? At dc, both capacitors have infinite reactance and can be ignored, leaving a 10:1 voltage divider circuit. As the signal frequency increases however, the reactance of the scope's input capacitance goes down. This would act as a low-pass filter and higher frequency signals and signal components would be attenuated, distorting the input signal.

The scope probe capacitance acts as a compensating high-pass filter. As frequency increases, its reactance goes down, creating a larger signal at the scope input. The effect — when the probe capacitor is properly adjusted — exactly balances the low-pass effect of the scope's input capacitance.

In your scope's manual, you will find instructions for adjusting a probe's compensating capacitor. The probe is connected to a square wave, usually provided at a test point right on the scope's front panel. This square wave is rich in harmonics that give it the sharp corners. If the probe's capacitance is too high, the high-pass effect allows too much harmonic energy into the scope and the square wave overshoots on each transition. If the probe's capacitance is too low, the low-pass effect dominates and the square wave's corners are rounded. By adjusting the probe compensating capacitor until the square wave's corners are as sharp and close to right angles as possible, the low- and high-pass effects balance and the frequency response of the combined probe and scope input is flat across the bandwidth of the scope.

Don't take my word for it — fire up your scope and give it a try! If you don't have a manual for your scope, the online probe compensation tutorial from Pico Technology will fill in the blanks for you.
[2]

Carefully adjust the compensating capacitor while watching the effect on the test square wave's corners. You will clearly see the high- and low-pass effects.

Notes

[1]

All previous "Hands-On Radio" experiments are available to ARRL members at
www.arrl.org/hands-on-radio
.

[2]

www.picotech.com/library/application-note/how-to-tune-x10-oscilloscope-probes

Experiment #167 — Clean Audio for Comfortable Listening

Throughout my years in ham radio, I've frequently heard something like, "It's a good receiver, but it just sounds harsh (or hissy)." Also popular is, "That radio just wears out my ears." And the ever-present, "The signals all sound like mush." Some of those might be symptoms of problems elsewhere in the receive chain, like in the Automatic Gain Control (AGC) sub-system. Quite often, though, the problems stem from a weak audio output stage.

The latest radios cost hundreds (if not thousands) of dollars and have sophisticated RF electronics and processing power. There should be an equal amount of attention paid to the actual interface with you, the user, and that is the audio output stage.

Frequency Response

In bygone days when "hi-fi sets" used vacuum tubes, frequency response was a serious part of "specs-man-ship." Bigger (heavier) audio output transformers helped extend the bottom end below 100 Hz. Higher-gain tubes and circuits extended the high end above 10 kHz. There were actually meaningful differences between the frequency response of different models. Today's integrated circuits (ICs) offer responses from well below 10 Hz to 20 kHz and beyond human hearing. Problem solved?

Not really. A communications receiver can actually have a frequency response that's *too* wide. Maybe for broadcast-quality AM, a wide response is necessary for that "warm" sound AM operators tout, but for HF SSB/CW, it just allows a lot of noise and interference into your "channel." On VHF+ FM, too much bass response means you'll begin to hear those no-longer-sub-audible tones. Too much high-frequency audio and noise will drive you crazy. You need a response appropriate to what you're trying to copy and no more; 200 to 5000 Hz is more than adequate, because IF filtering will limit signal audio response to much less than that.

Distortion

Another gremlin of audio stages is distortion, specified as *total harmonic distortion* (THD). Measured for a single-tone sine wave at a specific power level, THD is the sum of the power of all distortion products divided by the power of the fundamental frequency. Home entertainment audio gear sports THD of far less than 1%, but a typical ham transceiver specifies THD as high as several percent for a specific power level and load impedance. Is this a bad thing?

A lot of THD is definitely a bad thing, as it makes the receiver sound harsh or shrill even if the received signal is clean. You can see distortion by viewing the audio output of your rig with PC-based spectrum analyzer software such as DL4YHF's *Spectrum Lab*.[1] (Use the speaker output for lowest output impedance.) Tune in a steady carrier, such as WWV received in USB or LSB mode, and experiment with different audio levels, watching the relative amount of harmonics as you change the volume. Place different loads on the output by using a splitter to connect one or more paralleled speakers to the output. You'll see the harmonic content increasing with heavier loads and higher volume.

How can you tell if your radio has an audio problem? The Hands-On Radio web supplement for this column (www.arrl.org/hands-on-radio) has a short set of diagnostics.

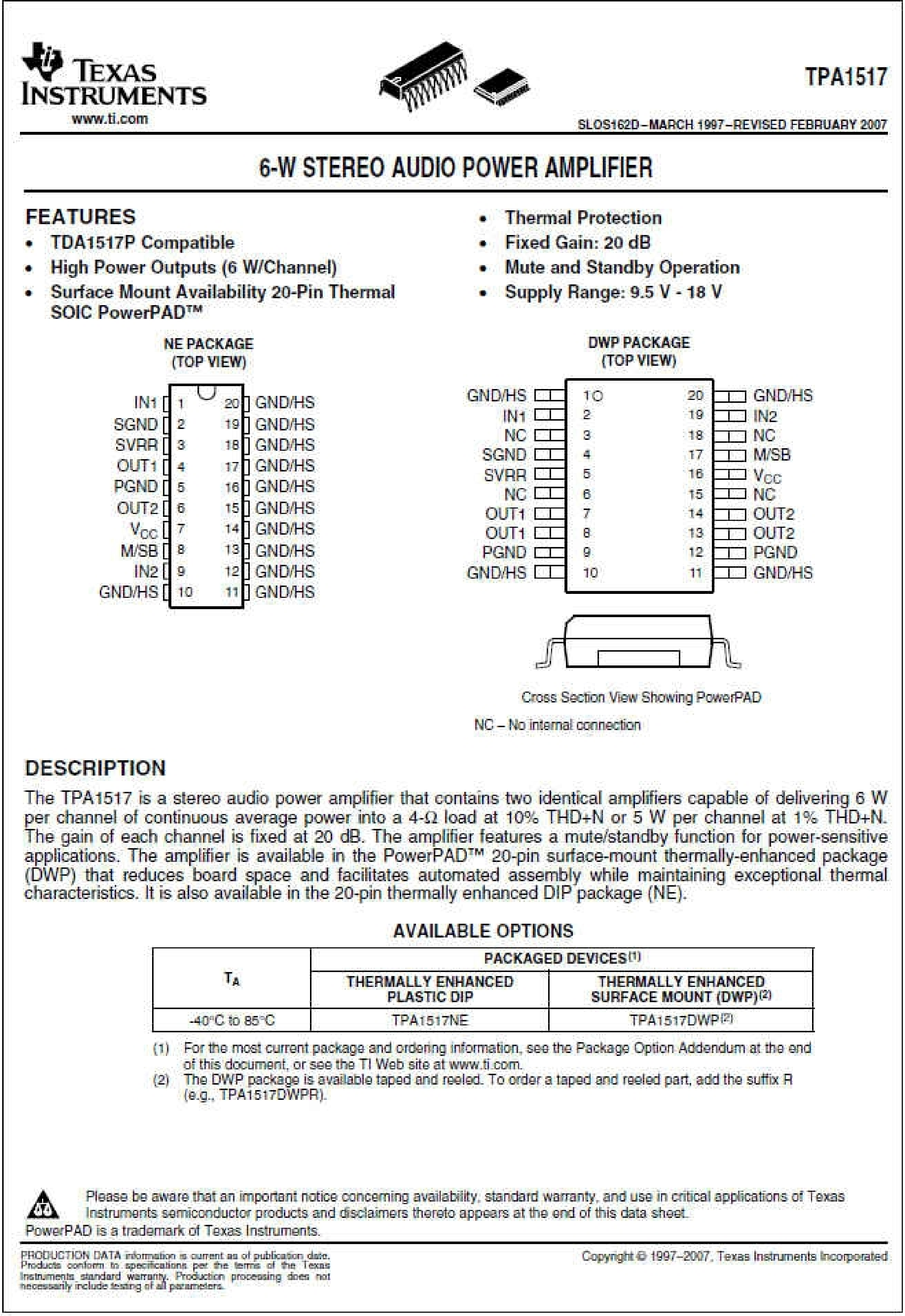
Hiss

Receiver hiss is a high-frequency "white" or wide-spectrum noise generated by the internal electronic devices. It is present whenever the receiver is turned on, regardless of what external noise is coming in through the antenna connector. While the hiss may not be loud, over long periods, it is fatiguing as it continually stimulates your inner ear and the auditory system. Rather than apply external filters, the best way to remove the hiss is by rolling off gain above a few kHz in the audio circuitry.

Audio Power Output Stages

For simple construction, it's been hard to beat the venerable LM386 audio power amplifier IC.[2] It has been used for the audio output stage of simple radios for many years, and the price is certainly reasonable. But it can be hissy with a frequency response of 300 kHz and has a ho-hum THD of 0.2% at only 125 mW of output power. This is okay for driving light headphones or a small speaker, but we can do better.

The proliferation of portable music players drives many improvements in audio output electronics. For example, the TPA1517 is made specifically to drive stereo headphones or speakers.[3] Its THD is still 0.1% at an output power of 1 W, which is plenty to drive even full-size headphones and small speakers. (This comes at the price of higher supply current over a pair of LM386 ICs.) Figure 1 shows the schematic for a bare-bones amplifier using this chip. The datasheet has additional circuit ideas.


TEXAS INSTRUMENTS
www.ti.com

TPA1517

SLOS162D–MARCH 1997–REVISED FEBRUARY 2007

6-W STEREO AUDIO POWER AMPLIFIER

FEATURES

- TDA1517P Compatible
- High Power Outputs (6 W/Channel)
- Surface Mount Availability 20-Pin Thermal SOIC PowerPAD™
- Thermal Protection
- Fixed Gain: 20 dB
- Mute and Standby Operation
- Supply Range: 9.5 V - 18 V

DESCRIPTION

The TPA1517 is a stereo audio power amplifier that contains two identical amplifiers capable of delivering 6 W per channel of continuous average power into a 4-Ω load at 10% THD+N or 5 W per channel at 1% THD+N. The gain of each channel is fixed at 20 dB. The amplifier features a mute/standby function for power-sensitive applications. The amplifier is available in the PowerPAD™ 20-pin surface-mount thermally-enhanced package (DWP) that reduces board space and facilitates automated assembly while maintaining exceptional thermal characteristics. It is also available in the 20-pin thermally enhanced DIP package (NE).

AVAILABLE OPTIONS

T_A	PACKAGED DEVICES(1)	
	THERMALLY ENHANCED PLASTIC DIP	THERMALLY ENHANCED SURFACE MOUNT (DWP)(2)
-40°C to 85°C	TPA1517NE	TPA1517DWP(2)

(1) For the most current package and ordering information, see the Package Option Addendum at the end of this document, or see the TI Web site at www.ti.com.
(2) The DWP package is available taped and reeled. To order a taped and reeled part, add the suffix R (e.g., TPA1517DWPR).

Figure 1 — The basic configuration for using the TPA1517 as an audio output amplifier. See the IC datasheet (Note 3) for more circuit ideas and notes for using the Mute/Standby switch. [Image courtesy of Texas Instruments]

There are many other audio output ICs made for driving low (4 Ω) and medium (32 Ω) load impedances. You can find them by searching for audio amplifiers from distributors like Digi-Key Electronics (**digikey.com**) and Mouser Electronics (**mouser.com**). From the many options, optimize your selection for power consumption, power output, or some desirable set of features. For example, the TPA1517 has a mute/standby control input, perfect for use in a transceiver.

Experimenters might also want to build an audio amplifier out of discrete components to become more familiar with using transistors. In that case, Figure 2 shows an audio power stage from *Experimental Methods in RF Design*.
[4]

This is a high-gain (46 dB, x400) amplifier that can provide clean audio to headphones or small speakers. Note the parallel combination of a 100 kΩ and 220 pF capacitor for both the input buffer stage and in the feedback loop from the output to the middle op-amp. This combination rolls off signals above 7 kHz (approximately $1/2\pi RC$) to limit hiss and high-frequency noise. A low-noise op-amp, such as the NE5532, should be used, as well.

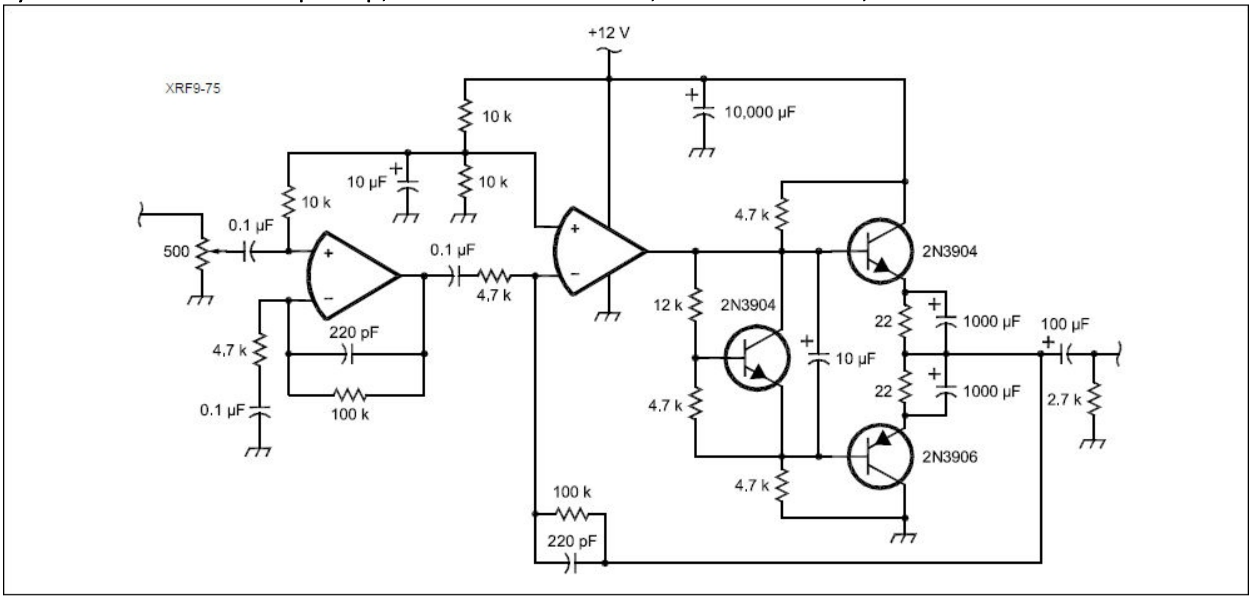


Figure 2 — A discrete audio power amplifier circuit. Use a low-noise op-amp, such as the NE5532. See text for notes on making connections to the amplifier.

Regardless of which design or IC you use, good wiring practices are required. Speaker or headphone connections should be made directly to the circuit board — do not use the chassis as the only signal return. If you are building the amplifier in a metal enclosure and using a stereo headphone jack, the body (sleeve contact) of the jack should be securely connected to the enclosure to keep common-mode RF on the cable outside. Run a lead from the jack's ground terminal to the common ground point of the amplifier.

[5]

A ferrite bead or two on the audio output lead helps keep RF out of the am
plifier, as well.

Headphones

Let's not forget about the actual final audio stage — the headphones or speaker. All this fine audio is wasted if you're not using good-quality headphones and speakers. Noise-canceling headphones are also beneficial, and they should be comfortable — with audio this clean, you'll be wearing them for many hours, enjoying what the band really sounds like!

Notes

1

www.qsl.net/dl4yhf/spectra1.html

2

www.ti.com/lit/ds/symlink/lm386.pdf

3

www.ti.com/lit/ds/symlink/tpa1517.pdf

4

W. Hayward, W7ZOI, R. Campbell, KK7B, R. Larkin, W7PUA, *Experimental Methods in RF Design,* ARRL, 2003.

5

J. Brown, K9YC, "A Ham's Guide to RFI, Ferrites, Baluns, and Audio Interfacing," Rev 5a,
audiosystemsgroup.com/RFI-Ham.pdf

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Experiment #174 — Switching Amplifiers

As part of our General and Amateur Extra class license studies, we have to learn a few things about amplifier classes. *Class* refers to how the amplifying devices (tube or transistor) operate, and how the output impedance matching and filter circuits work.

Class Background

The early Hands-On Radio experiments (#1 and #2, for example) stress the importance of *quiescent-* or *Q-point*. [1]

The Q-point is the combination of voltage and current in a circuit when no input signal is present.

The output will follow a *load line* (see Experiment #77) that describes voltages and currents possible for the load at the amplifier's output. The device's output current can be increased until it reaches *saturation*, the point at which further changes in input have no effect. Similarly, the output current can be reduced to zero at *cutoff*.

The location of the Q-point on the load line determines how "close" the amplifier is to saturation or cutoff. The amplifier circuit's configuration, Q-point location, load, gain, and output circuit characteristics all combine to determine the amplifier's output waveform, linearity, and efficiency. There are several common combinations and resulting behaviors of the amplifier — these are called *amplifier classes*.

Classes A, B, and AB

The easiest way to compare these classes is in terms of *conduction angle* — the number of degrees in one cycle during which the device is conducting current in its *linear region*, from 0° to 360°.

In *Class A* operation, the device (Figure 1A) conducts during the entire cycle; a conduction angle of 360º. If the load line is linear and the input doesn't "overdrive" the circuit out of its *linear region* (between saturation and cutoff), the output waveform is an exact reproduction of the input. This is a true "linear amplifier." Class A amplifiers have poor efficiency (about 50% maximum) because of their continuous current.

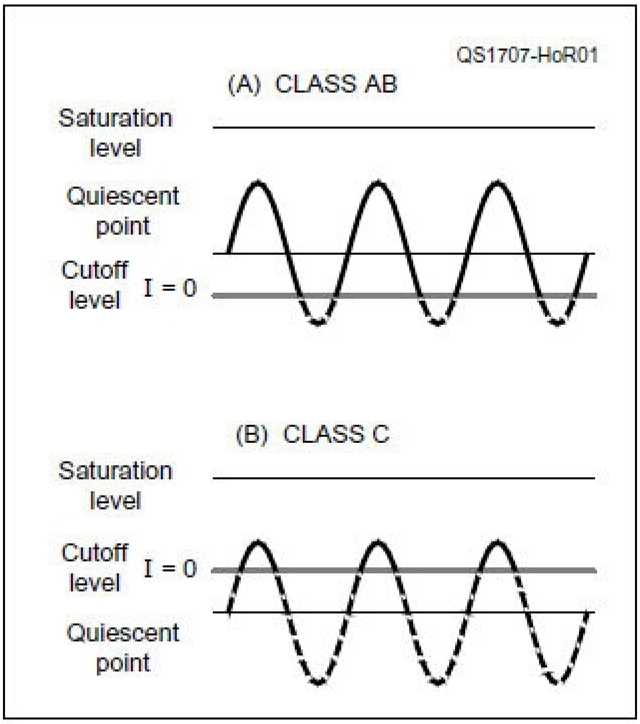

Figure 1 — Amplifier output waveforms for Class AB and Class C amplifiers with sine wave inputs. The output of Class A is a complete sine wave, and the output of Class B is either the positive or negative half-cycles.

In *Class B* operation, the conduction angle is 180º (half of the cycle). Class B amplifiers are more efficient (a maximum of 78.5%) than Class A, because they are only in their linear region half the time. But they are quite non-linear. Class B amplifiers typically use a pair of devices that conduct on alternate half cycles (*push-pull* operation). Their outputs are combined to produce a complete output signal.

Class AB operation (Figure 1A) "splits the difference" between Class A and B with a conduction angle between 180° and 360°. This is a compromise — Class AB amplifiers are more linear than Class B, and more efficient than Class A. Class AB is very popular for SSB operation, whether tube or transistor. Some readers are probably thinking, "But Class AB is still really non-linear. How can that work and still put out a clean signal?" Hold that thought.

Class C and Harmonics

The final "traditional" class is *Class C* (Figure 1B), in which the device is in its linear region for a short period, with a conduction angle of only 25 – 50% (90 – 180º). The output current consists of sine-like pulses. Class C amplifiers are very non-linear and are used to amplify signals for which only the frequency is preserved — FM and CW, for example. Because the device is only in the linear region for short periods, the efficiency of Class C amplifiers can be as high as 85%.

Clearly, Class AB, B, and C amplifiers generate a lot of harmonic distortion, because they cut off some or most of the waveform. (Remember that we are talking about an RF waveform, not audio.) There would be significant in-band distortion products as well, particularly for Class C.

Let's do an experiment. Build the common-emitter amplifier of Experiment #1 with a 50 kΩ pot instead of R1 and R2. Adjust the bias to change the Q-point, and with a 1 kHz input signal, observe the output signal spectrum by using a sound card audio analyzer (see Experiments #64 and #65). Compare the spectra for Class A, B, AB, and C operation.

To get rid of the harmonic energy requires the output circuit to act as a filter as well as provide an impedance match. When you adjust the **TUNE** function of a tube amplifier, you are bringing the output circuit to resonance where it acts as a band-pass filter. (If you have a linear amplifier, tune it up at reduced power into a dummy load, then listen for your second or third harmonic on a second receiver. Change the tuning of the amplifier while watching the S-meter to get an idea of signal level. You should be able to see (and hear) the effects of mistuning.

Class D and Switching Amplifiers

Class D goes farther than Class C — the amplifier acts as a *pulse width modulated* (PWM) switch by setting the bias and drive so that the device is either fully on (saturated) or fully off (cutoff). As Experiment #9 shows, these two states have low-power dissipation compared to the linear region. Class D is a type of *switching amplifier*.

The switch operates at f_{PWM}, which is many times the frequency of the signal to be amplified. The duration of each pulse is controlled by the amplitude of the input waveform to change the pulse's average energy. The switch is turned on and off so rapidly that it spends very little time in the linear region, and its conduction angle is very small. This creates a series of current pulses.

A low-pass filter then removes the high-frequency switching components and leaves only the desired average-energy signal. Essentially, a Class D amplifier is a switching mode power supply, with the input signal acting as the output voltage control signal.

Because the switching frequency must be many times the frequency of the signals to be amplified, Class D circuits are only used for audio. The switching frequency is typically greater than 100 kHz and easily separated from the desired audio range below 20 – 30 kHz. There are many Class D audio amplifier integrated circuits (ICs), and because they are so efficient (more than 90%), they are the circuit of choice in battery-powered gear like smartphones and music players. (The Sparkfun BOB-11044 Class D amplifier kit based on the TPA2005D1 chip — **www.sparkfun.com** — only costs a few dollars and makes a great experiment.)

Class E and F

At RF, the frequency of the switching pulses, f_0, is so high that it takes more work to keep the device out of the linear region. Remember that current should be high or voltage should be high, but not both at the same time. This is accomplished by the tuned output networks, as shown in Figure 2, described more completely by Iulian Rosu, YO3DAC/VA3IUL.[2]

For *Class E*, the output network's series-LC network is tuned to the desired RF frequency of operation. The capacitance across the switch is charged and discharged at the same rate. The tune
d circuit causes voltage at the device to go to zero just as the current is beginning to increase. This keeps switching losses low, and the tuned circuit takes care of the harmonics. A separate tuned network is required for each band of operation.

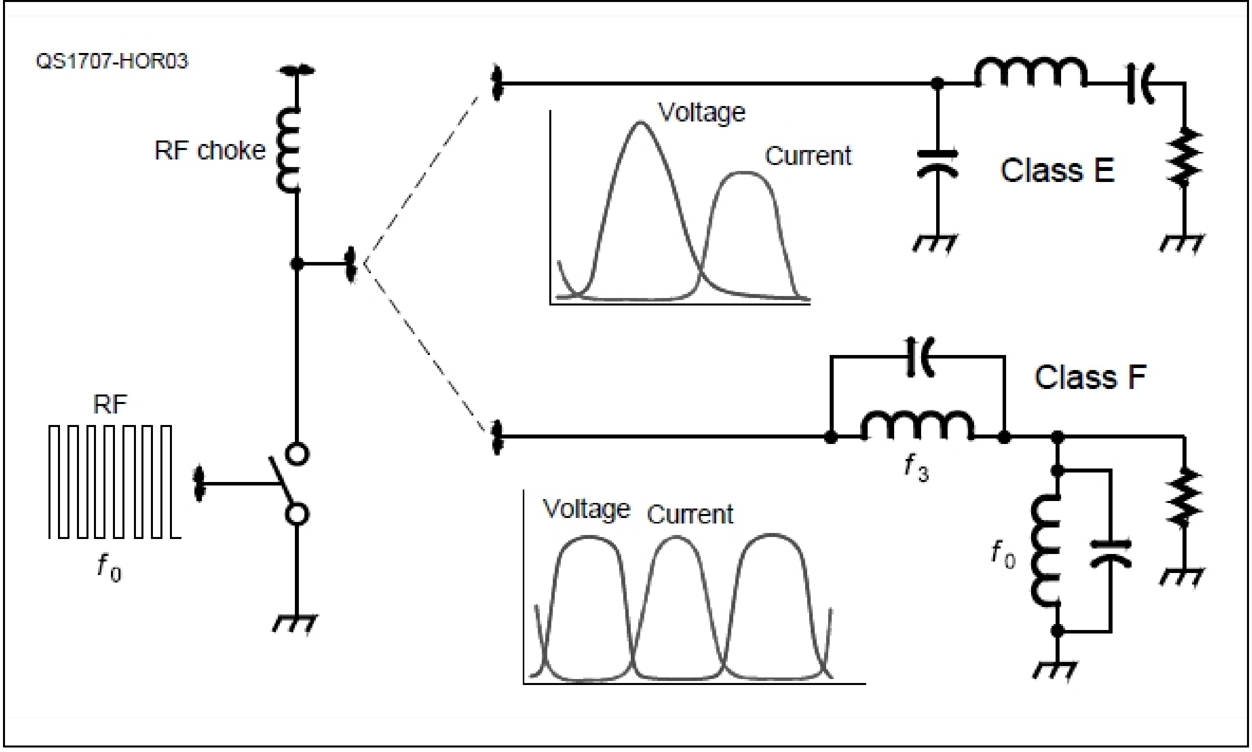

Figure 2 — Class E and Class F amplifiers use tuned output networks to reject harmonic energy while ensuring that the switch does not operate in its linear region.

The related *Class F* amplifier uses a parallel-LC trap in the output circuit to block f_3, the third and strongest harmonic of a square wave, while passing the fundamental (f_0). Again, with careful tuning, the voltage and current are never maximum at the same time, the switch stays out of its linear region, and efficiency is high.

Out-Classed

There are other classes, as well. *Class G* is similar to a Class AB amplifier, but switches between two voltage levels to reduce power dissipation at low signal levels. *Class I* uses two devices driven with complementary pulse duty cycles to cancel harmonics and follow the input waveform. *Class S* is a variation on Class D, and *Class T* uses digital signal processing (DSP) to optimize pulse widths in a Class D amplifier for better performance.[3]

Notes

1

All Hands-On Radio experiments are available online to ARRL members at www.arrl.org/hands-on-radio.

2

I. Rosu, YO3DAC/VA3IUL, “RF Power Amplifiers,” www.qsl.net/va3iul.

3

“Amplifier Classes,” Electronic Tutorials, www.electronics-tutorials.ws/amplifier/amplifier-classes.html.

Experiment #124 — The Beta Match

This month's column will show you four different types of antenna feed-point impedance matching that all work in the same way even though they look quite different. In addition, I'll introduce a new friend for your computer toolbox — *SimSmith* by Ward Harrington, AE6TY.

The basic problem — impedance matching using inductance across an antenna's feed-point — is the same, but the solution goes by several names. This makes it more difficult to understand because giving the same things a different name (or giving different things the same name) is confusing. Nevertheless, as you read the column, keep in mind that all of the techniques presented here accomplish the same task.

The Essential Problem and its Solution

While discussing impedance matching of antennas, it's natural for most hams to imagine impedances *greater* than 50 Ω. The examples we use in learning about SWR primarily use higher impedance values for the calculation Z_{LOAD} / Z_0: 100 / 50 = 2:1, for example. In actuality, it's common for Z_{LOAD} to be *less* than 50 Ω. A Yagi's driven element feed-point impedance is often in the range of 20 to 30 Ω and the natural impedance for a quarter-wave monopole (the common ground plane) is around 35 Ω.

Transforming this lower impedance to 50 Ω doesn't lend itself well to the most common techniques. The impedance ratio of 1.4 - 2:1 doesn't fall into the usual 1:2:4:9-type ratio of "easy" transformer impedance ratios, nor are there coaxial cables with a characteristic impedance of 35 to 40 Ω that would enable simple quarter-wave "Q sections" to do the job. [1]

[A pair of 75 Ω cables in parallel will get in range, though they are a bit clunky. — *Ed*.]

Nevertheless, the clever approach illustrated in Figure 1 gives the electrical schematic view of how this problem is solved. First, we have to give up the usual assumption that the feed-point impedance is resonant — that is, R + *j*0 Ω. It's part of our "ham DNA" that makes us think antennas need to be resonant to work, but in this case resonance actually makes the problem harder.

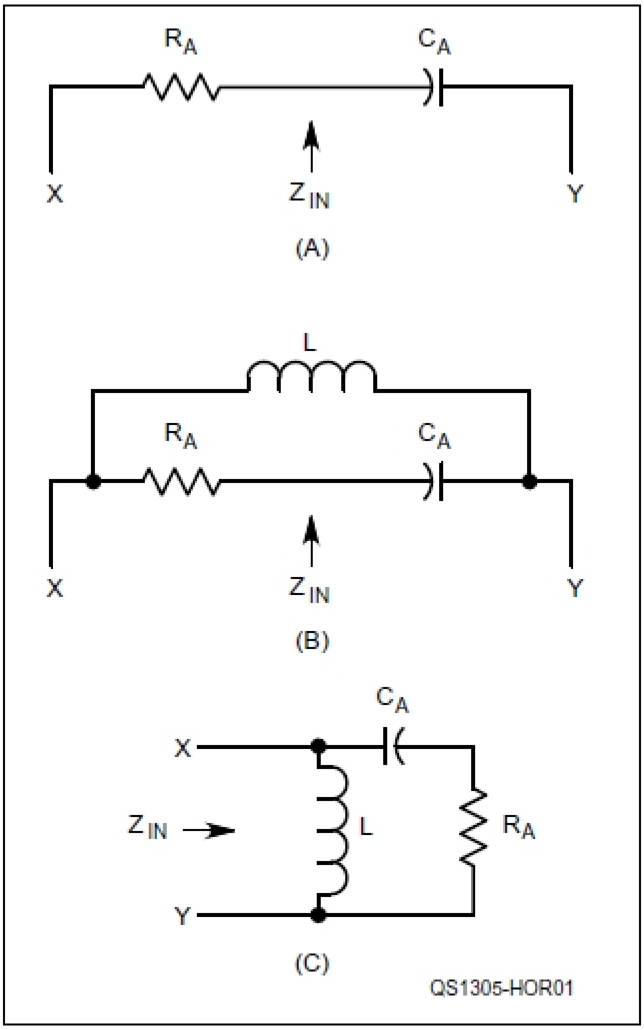

Figure 1 — A Yagi driven element or a monopole's feed-point impedance is made capacitive by shortening it below its resonant length. Adding inductance in parallel with the resulting impedance creates an L network as in (C), transforming the impedance to 50 Ω.

By making the antenna a little shorter than its resonant length, the feed-point impedance becomes slightly capacitive (A). That capacitive reactance can then be used as part of an L network by adding an inductor across the feed point as shown in (B). Redrawing the circuit (C) results in the L network's more familiar form. (See Experiment #21 for more about L networks.)

There is a bit of a trick involved. You can't have just *any* amount of resistance and capacitive reactance. The combination has to be in the right range so that adding the inductance transforms the impedance to 50 + j0 Ω. How do you figure that out?

SimSmith

The standard way of visualizing transmission line and impedance matching mechanics is by using a Smith chart. (If you are unfamiliar with the Smith chart, read the introduction in Hands-On Radio experiments #59-61. Recent editions of the *ARRL Antenna Book* include a detailed tutorial on the Smith chart, either in print or on the CD-ROM.[2]) Yesterday's compass and straightedge have been replaced by interactive computer software such as the easy-to-use *SimSmith* (**www.ae6ty.com/Smith_Charts.html**). Written in Java, AE6TY's free tool is available for a wide variety of computers. Furthermore, he has created videos and guides to explain how to use the software and the Smith chart, so there is no reason not to have a copy and begin learning the power of "seeing inside" transmission lines and matching networks. Before taking a look at our current problem using *SimSmith*, allow me to point out a few of its features that I will use here.

First, as you can see in Figure 2, the program shows the usual *constant-resistance circles* and *constant-reactance arcs* in light red. Less usual are the *constant-conductance circles* and *constant-susceptance arcs* shown in light blue. (Susceptance, B, is the reciprocal of reactance, X.) The normalized 1.0 + j0 point is at the center. (In our 50 Ω world, that represents an impedance of 50 + j0 Ω or an admittance of 0.02 + j0 S, where S is the symbol for siemens, the unit of conductance.) From any impedance or admittance point on the chart, adding resistance or reactance in series "moves" along the red circles and arcs. Adding resistance or reactance in parallel or *shunt* "moves" along the blue circles and arcs. (Instead of move, I'll use the correct term *transform* from here on.)

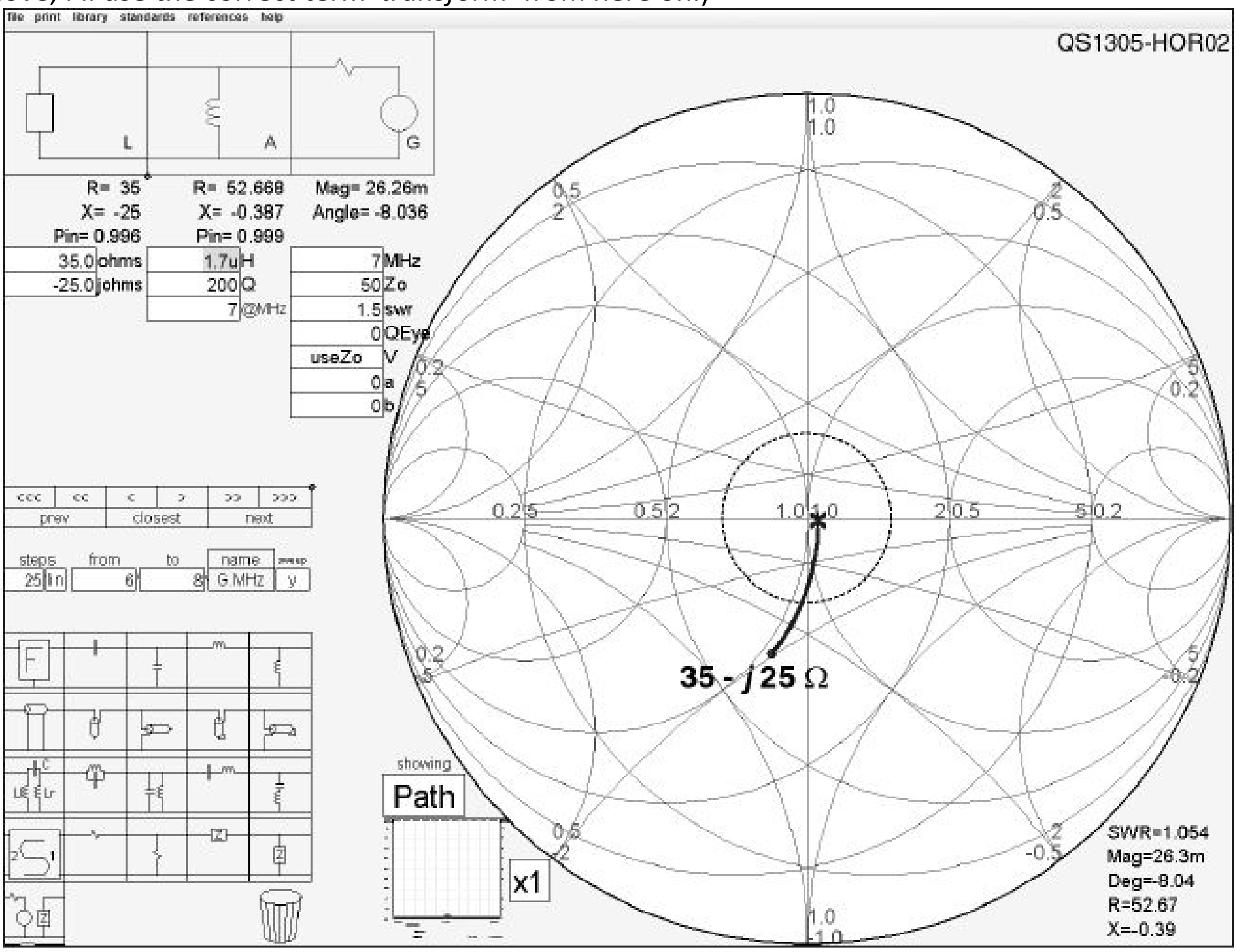

Figure 2 — *SimSmith* screen showing the shunt inductance sub-circuit matching a load of 35 – j25 Ω.

At the upper left, *SimSmith* shows the transmission line circuit you've constructed, including a load at the left and the source (*generator* in Smith chart speak) at the right. A collection of subcircuits is available at the lower left — there is everything from series and shunt components, to stubs, to tuned circuits, to general-purpose blocks that perform specific math functions. Add a subcircuit by drag-and-dropping it onto the transmission line circuit at the desired point. Then fill in the values (too small to reproduce in the figure) underneath the subcircuit. *SimSmith* does the rest. Let's try it.

Using a Coil

Figure 2 is a screen shot from *SimSmith* showing the equivalent of Figure 1C. An easier to read version is on the Hands-On Radio web page. By entering 1.5 in the generator's SWR value window, *SimSmith* drew a *constant-SWR circle* around the center — all points within this circle represent SWR values of 1.5:1 or less — for reference. I selected a frequency of 7 MHz because I use a quarter-wave vertical on 40 meters at my station. Game on!

While designing an antenna, achieving the goal of being able to use a single shunt inductor as your matching network requires the right feed-point impedance, shown as the load on *SimSmith*. Adding the shunt inductor will transform the impedance counter-clockwise parallel to one of the light blue constant-conductance circles as shown by a heavy blue line. (The inductor adds susceptance but does not affect the conductance. Parallel capacitance transforms clockwise.) The feed-point impedance should be designed such that adding inductive susceptance can transform the impedance to within the desired constant-SWR circle.

Starting with the antenna's feed-point impedance at the left, I've entered a value of 35 – *j*25 Ω. This is a reasonable value for an aluminum tubing vertical over a good ground system, adjusted to a bit less than its natural resonant length. By "fiddling with" (technical term) the value of inductance, I found that a value of 1.7 µH presented a resulting impedance of 52.3 – *j* 0.4 Ω to the feed line for an SWR value of 1.05:1. I'd say that works. In fact, this is quite close to the size of inductor I use to match my vertical antenna on 40 meters.

Try a Shorted Stub

Assuming you've downloaded *SimSmith* and are running it, enter the same values for load impedance but replace the parallel inductor with a shorted stub — the sub-circuit directly below the parallel inductor. Drag-and-drop the parallel inductor sub-circuit into the trash can symbol. Then drag the shorted-stub subcircuit to the transmission line circuit.

From the values *SimSmith* assumes about the stub (such as it being made of 50 Ω coax), adjust the length until you get about the same match as with the inductor (≈55°). For fun, increase stub length to 90° — the stub now presents an open circuit so that it does nothing. (See Experiment #22 for more about stubs.)

Can you use a longer length to create an inductive feed-point impedance and match it with a parallel capacitance? Change the feed-point impedance to 35 + *j*25 Ω and find out. (300 pF should get you close.)

Using a Hairpin

Along with shunt inductance and shorted stubs, the third name and fourth idea covered here is the *hairpin* or *beta match* shown in Figure 3. You should recognize the matching device as a shorted stub of open-wire transmission line. Typical hairpins are made of heavy wire with wide spacing that results in a high characteristic impedance. What length of hairpin is required to match our original load if its characteristic impedance is 300 Ω? (Roughly15° or 2.8 feet at 7 MHz for a 95% velocity factor.)

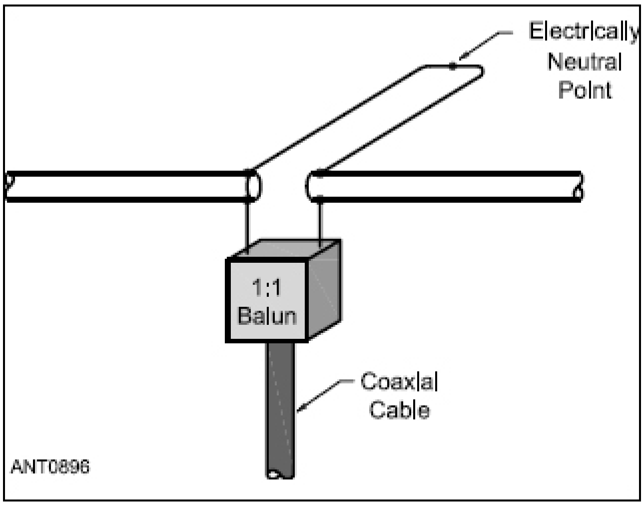

Figure 3 — The hairpin match across a balanced, insulated driven element. Hy-Gain antennas attach the electrically neutral center of the hairpin to the antenna boom, creating a beta match.

If the driven element is insulated and electrically balanced, the very center of the feed-point and the matching hairpin are electrically neutral. Hy-Gain antennas add mechanical stability to the design by attaching that point of the hairpin to the antenna boom — also electrically neutral with respect to the feed line — creating the *beta match*.

A Common Theme

You should now see the common theme of all four matching designs. By creating capacitive reactance in the feed-point impedance and applying a shunt inductance across the feed point, the ratio and phase of voltage and current can be altered to create a purely resistive impedance of the desired value.

[1]

See Hands-On Radio experiment #81, "Synchronous Transformers." All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

[2]

The ARRL Antenna Book, 22nd Edition. Available from your ARRL dealer or the ARRL Bookstore, ARRL order no. 6948. Telephone 860-594-0355, or toll-free in the US 888-277-5289; **www.arrl.org/shop**; **pubsales@arrl.org**.

Experiment #133 — Extended Double Zepp Antenna

The reference to the Extended Double Zepp (EDZ) antennas in Experiment #131 (on the coax to open-wire balun) certainly generated some interest!

[1]

The Zepp is one of the oldest antennas. Patented in 1909 by Hans Beggerow (German patent 225204, www.aktuellum.com/circuits/antenna-patent) the antenna is shown suspended vertically from a balloon (naming it after the Zeppelin airship came later) looking for all the world like an upside-down J-pole, which, in fact, it is!

The Zepp is usually imagined as horizontal, as in Figure 1A, and the J-pole as vertical, but electrically they are essentially the same antenna. Both use a quarter-wave section of transmission line to convert the high impedance at the end of a half-wave radiating element to a lower impedance suitable for attaching to feed line.

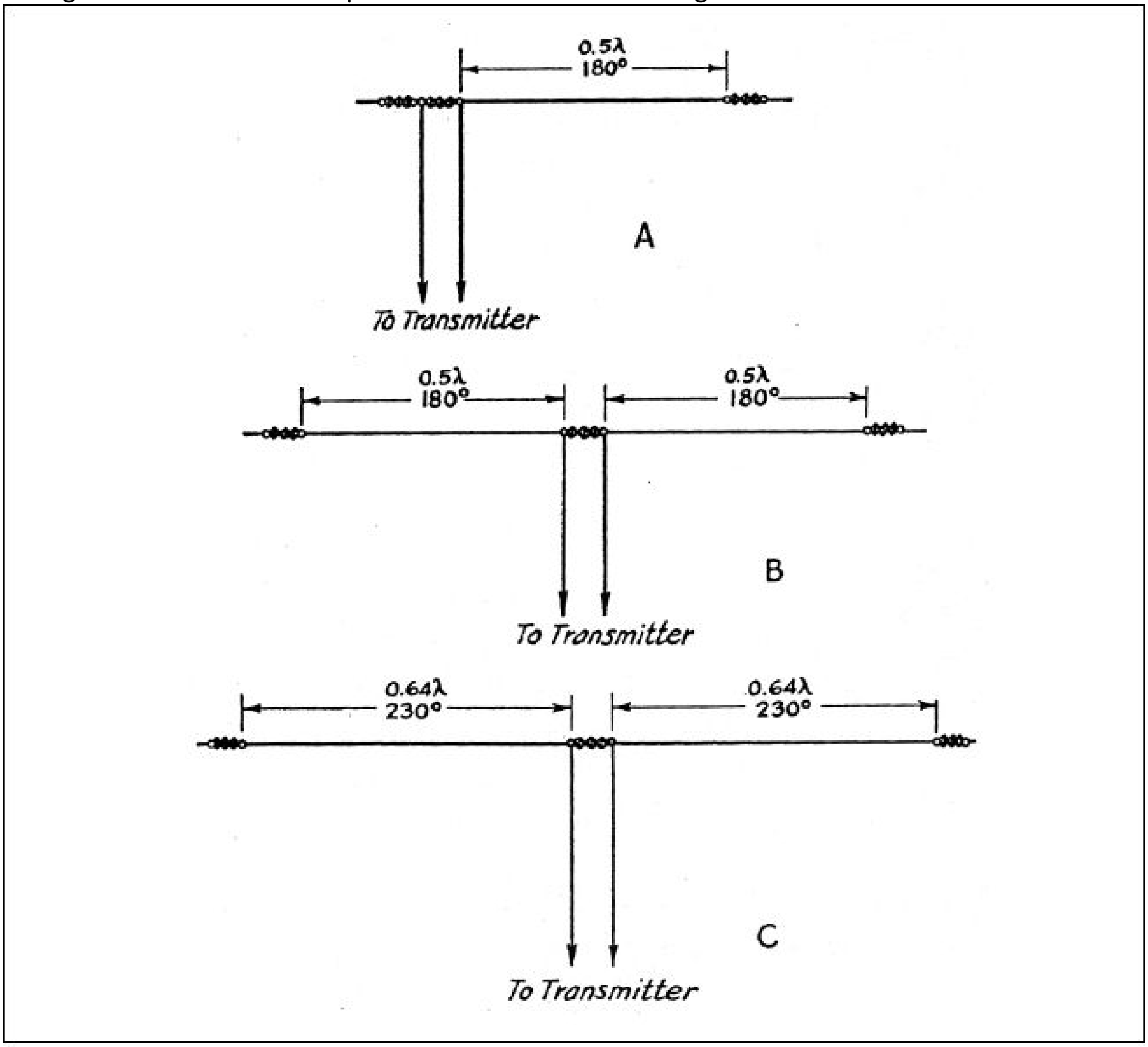

Figure 1 — The evolution of the Zepp antenna from a single half-wave antenna (A) to a pair of half-waves in phase (B) through the Extended Double Zepp antenna at C. [*QST*, June 1938]

Pepping Up the Zepp

The half-wave Zepp offers no advantage over a center-fed dipole in terms of gain or directivity. The only difference is the feed point being located at the end of the Zepp and in the middle of the dipole. As a result, the basic Zepp has a gain of 0 dBd.

[2]

Like arranging a pair of dipoles in an array to focus the radiated energy in a desired direction, creating gain, a pair of Zepps can be connected end to end as shown in Figure 1B. This creates the "two half-waves in phase" antenna that narrows the broadside pattern a bit and has gain of 1.9 dB over a single Zepp. (The net gain is less than 3 dB due to coupling between the separate elements.) The "double Zepp" is a basic collinear array with both elements lying along the same line.

The "missing" 1.1 dB of gain would be available if the radiation patterns of the two half-wave antennas could be added together independently. Coupling between the two antenna halves can be reduced by moving the elements farther apart, but feeding them would then become complicated. This problem was solved in a 1936 IRE paper by GH Brown who lengthened or extended each element from 180 degrees (half-wavelength) to 230 degrees, as seen in Figure 1C.

This antenna was introduced to amateurs in the June 1938 issue of *QST* by W2NB as the Extended Double Zepp or EDZ.

Not only does the antenna "recover" the missing gain to a full 3 dBd but can also be easily matched to either open-wire or coaxial feed line through the use of transmission line techniques. A more recent *QST* article by W5JH gives the EDZ design information shown in Table 1 for HF bands from 40 through 10 meters. [3]

Table 1
EDZ Antenna Dimensions for Different Bands

Freq (MHz)	L_d (ft)	Min Height (ft)	Antenna Z	Feed point Z	L_f (ft)	SWR
7.075	175.2	66	170.5 –*j*976.1	47.10 +*j*0.25	21.65	1.062:1
10.11	122.6	47	163.0 –*j*934.1	47.80 –*j*0.07	14.85	1.046:1
14.175	87.5	34	155.3 –*j*889.6	48.62 –*j*0.36	10.35	1.029:1
18.1	68.5	30	133.4 –*j*848.9	44.63 +*j*0.38	7.92	1.121:1
21.2	58.5	30	132.1 –*j*799.9	47.65 –*j*0.28	6.56	1.050:1
24.9	49.8	30	156.7 –*j*772.3	58.60 –*j*0.10	5.51	1.172:1
28.2	44	30	169.8 –*j*772.4	63.24 +*j*0.26	4.88	1.265:1

L_d is the antenna length, L_f is the length of the matching feed line
Feed point Z refers to the impedance at the end of the open-wire feed line where coax can be attached.

Feeding the Zepp

Matching the EDZ to a feed line is an interesting story. With each element of the array being longer than one half-wavelength, the feed point impedance is quite reactive. For example, W5JH gives the feed point impedance of a 20 meter EDZ as 155.3 – *j*889.6 Ω for an SWR of about 15.5:1. If connected directly to 100 feet of 50 Ω RG-8X coax that has 0.9 dB of matched loss at 20 meters, *TLW* calculates that a 15.5:1 SWR would result in 9.8 dB of additional loss for a total of 10.7 dB in the feed line. [4]

Obviously, it is a good idea to lower the SWR in some way!

Luckily, the impedance transforming properties of transmission lines can be used to change the impedance. (See Experiments #59 – 61 on the Smith Chart.) In this particular case, using a short length of high-impedance open-wire line (also called a ladder or window line) transforms the high feed point impedance (the "Antenna Z" column in Table 1) to something quite close to 50 Ω (the "Feed Point Z" column in Table 1). Figure 2 shows how a 10.3 foot section of 600 Ω line transforms the antenna feed point impedance point to very nearly 50 + *j*0 Ω at the center of the chart.

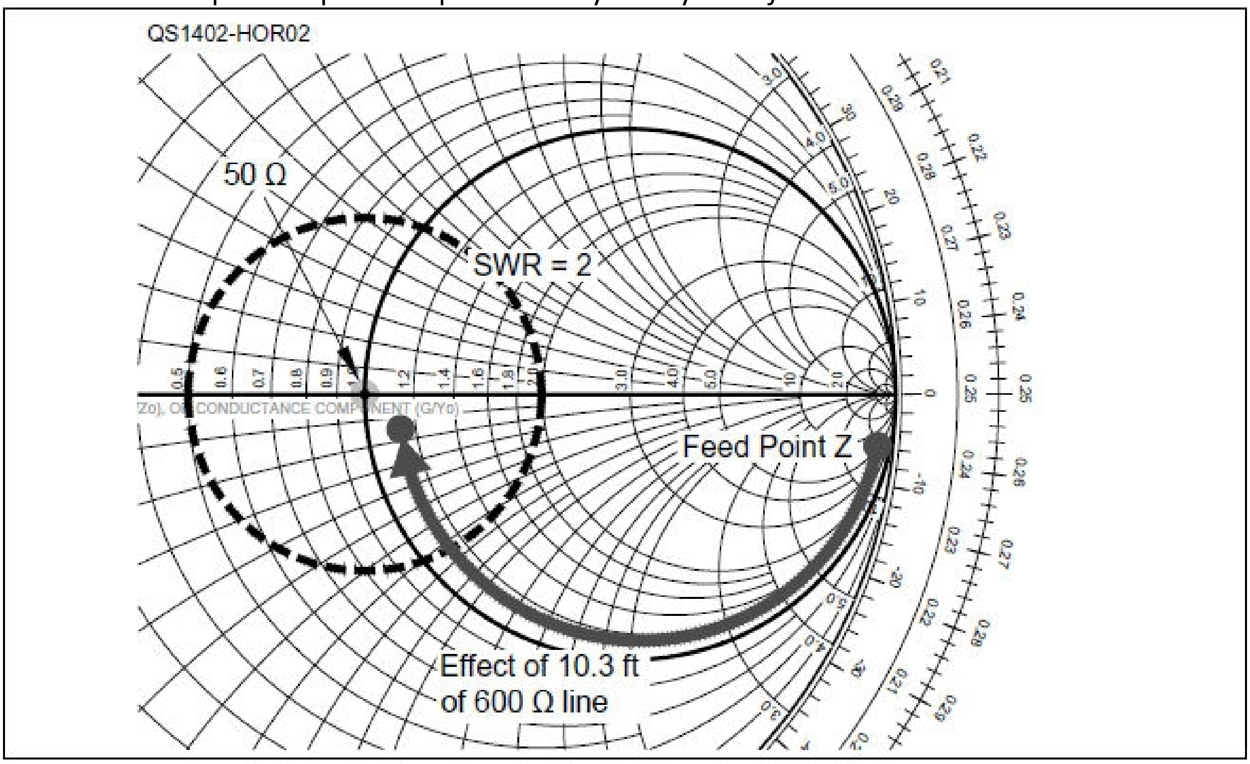

Figure 2 — By attaching a 10.3 foot section of 600 Ω transmission line, the highly reactive feed point impedance of 155.3 – *j*889.6 Ω can be transformed to very nearly 50 + *j*0 Ω.

At the end of this short feed line section, you have two options — connect a 50 Ω feed line or extend the open-wire line by some multiple of half-wavelengths so that the 50 Ω point is reached again and attach the 50 Ω feed line (or transmitter) there. By attaching 50 Ω coax at a point where the high-impedance open-wire line presents a 50 Ω impedance the SWR will then remain low in the coax all the way to the transmitter.

If your antenna is up in the air, the end of the matching section of 600 Ω line will be dangling well off the ground and you might not want to lift that much coax. Furthermore, the solid conductors of most open-wire feed lines will break from flexing in the wind with this load attached. Since the impedances in a transmission line repeat every half-wavelength along the line, you can add feed line in half-wavelength sections and reach another 50 Ω point, hopefully near the ground or a mechanical support.

In our case, for 600 Ω open-wire line with a velocity factor of 0.92, one half-wavelength at 14.175 MHz is 32 feet. By adding multiples of 32 feet to the overall length, you can bring the 50 Ω point to a location where it is more convenient to attach a coaxial feed line, perhaps using a balun, as was shown in Experiment #131.

450 Ω window line also works but not quite as well. I found that *TLW* indicated that a length of 11.6 feet transformed the impedance to about 27 Ω for a minimum SWR of nearly 2:1. This is a lot better than 15.5:1 and will be lower at the end of the coax due to its loss but it might be worth buying or making your own 600 Ω line if you want to use this antenna system design.

I decided to go further by using *EZNEC* antenna modeling software (www.eznec.com), creating a design with a low SWR point on two bands, 20 and 15 meters while using 450 Ω window line. I started with the dimensions of the 20 meter EDZ in Table 1, and dug in. You can do this too — start with the W5JH dimensions then optimize for your needs.

By lengthening the antenna to 89 feet and using a transmission line length of 11.1 feet, an SWR of 1.6 was obtained at 14.05 MHz and 1.4 at 21.1 MHz. Bringing the low SWR point to ground level required more window line. Since I needed an integer number of half-wavelengths at both 14.05 and 21.1 MHz, I added 2 half-wavelengths at 14.05 MHz (63.8 feet according to *TLW*) for a total of 74.9 feet which is also 3 half-wavelengths at 21.1 MHz. A bit more optimizing gave a line length of 74.7 feet and an SWR of 1.5 and 1.3:1 on 20 and 15 meters, respectively. Including ground reflections, gain is around 8.7 dBi on both bands with no tuner required.

I've installed a trio of these antennas in a triangle so the three 20 meter patterns cover all the main DX azimuths. Being electrically long on 15 meters, the antenna generates "four-leaf clover" patterns so my next project is redesigning the antenna for a single main lobe on both bands. Luckily, W7SX tackled that question in a July 1999 *QEX* article so I will be putting *EZNEC* to work once again![5]

[1] All previous Hands-On Radio experiments are available to ARRL members at www.arrl.org/hands-on-radio.\

[2] dBd specifies gain with respect to a dipole, usually in free space. Add 2.15 dB to obtain dBi, gain with respect to an isotropic radiator.

[3] J. Haigwood, W5JH, "The Extended Double Zepp Revisited," Sep 2006, *QST*, pp 35 – 36.

[4] *TLW or Transmission Line Program for Windows*, by Dean Straw, N6BV, is a transmission line calculator program included with the ARRL Antenna Book available from your ARRL dealer, or from the ARRL Store, ARRL order no. 6948. Telephone toll-free in the US 888-277-5289, or 860-594-0355; fax 860-594-0303; www.arrl.org/shop/; pubsales@arrl.org.

[5] R. Zavrel, Jr., W7SX, "The Multiband Extended Double Zepp and Derivative Designs," July 1999 *QEX*, pp 34 – 39.

Experiment #136 — End-Fed Antennas

An HF antenna currently enjoying some popularity, especially with backpackers such as those active in the Summits On the Air (

www.sota.org.uk

) program, is the End-Fed Half-Wave (EFHW). One of the oldest antennas, it was originally known as the "Zepp" and is widely used today in its VHF/UHF disguise as the J-pole (see Experiment #133

[1]

). Mechanically, it can be convenient to attach the feed line at one end, which is also a support point. It is also easy to toss one support rope over a high point and let the EFHW slope to a lower point where the feed line connection is made. Electrically, however, there is more to this antenna than meets the eye, and that can lead to some unexpected results.

At the End

The impedance of a half-wavelength piece of wire varies from a minimum of about 73 Ω when the feed point is at the middle to a much higher value at the end. Off-center-fed (OCF) antennas take advantage of this by locating the feed point somewhere at which a medium-sized impedance occurs on several bands. A fixed-ratio impedance transformer then creates a coax-friendly impedance on several bands.

At first glance, the impedance at the end of the wire should be infinite because the current has to be zero. In the real world, however, there is a fair amount of capacitance between the antenna and anything close to it that conducts electricity. This lowers impedance, especially at the end of the antenna. For example, at 14 MHz, 10 pF of capacitance is approximately 1.3 kΩ of reactance. It doesn't take many pF of capacitance in parallel with the impedance at the end of a wire to lower the resulting impedance dramatically.

Yes, but capacitance to what? Conductive material within a quarter to a half-wavelength of the antenna. That includes capacitance to ground which in a center-fed half-wave is responsible for some current flow at the end of the antenna and partly responsible for making the wire seem longer electrically than it is physically, helping to create the familiar formula for dipole length: $l = 468 / f$ (see Experiment #92).

The Whole Enchilada

What else is nearby that conducts electricity? The feed line, of course! In the absence of stern measures to prevent it, there will be a hefty amount of common-mode current flowing on the feed line, whether coaxial or parallel-conductor. But wait, isn't parallel-conductor feed line balanced? Yes, but only to the differential-mode currents (see Experiment #91). Along with the equal-and-opposite currents carrying power in the line, each of the conductors can pick up common-mode current just like any other wire and re-radiate signals just like any other antenna. For coax, the center conductor and the inside surface of the shield may be isolated from the antenna's radiated field but the outer surface is completely exposed and it, too, picks up common-mode current and re-radiates a signal. The combination of the antenna wire and the feed line's common-mode current path form the entire *antenna system*.

Based on the end-fed antenna model in W7EL's *EZNEC* User Manual, Figure 1 illustrates why considering the entire system is important.

[2]

,

[3]

You can see that antenna current varies from zero (this is a simulation) at the un-fed end to a small but non-zero value at the feed point end. The approximately half-wavelength feed line, in this case parallel-conductor, is terminated at the bottom by the power source and has one conductor attached to the antenna at the top. The remaining conductor is left open.

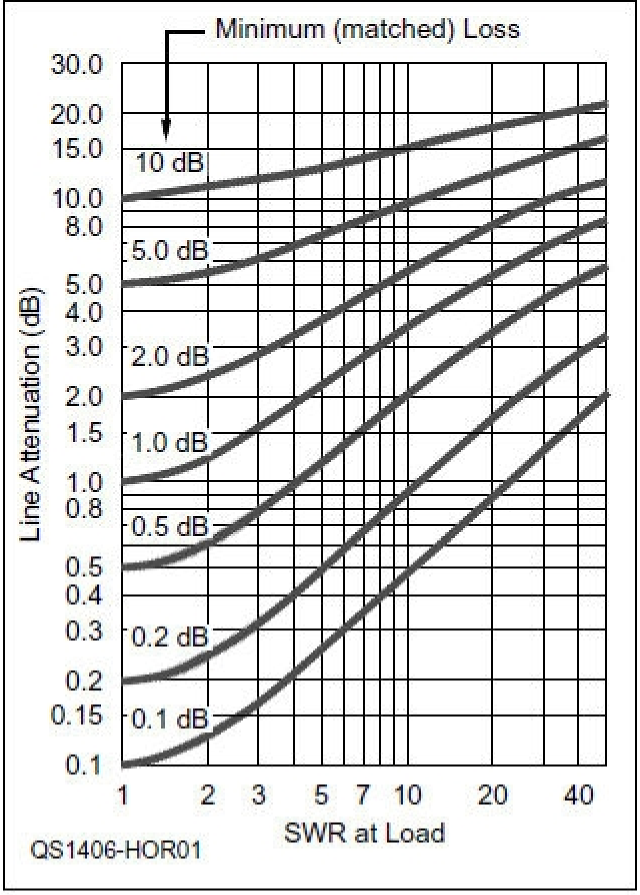

Figure 1 — An *EZNEC* simulation of an end-fed half-wave antenna and its feed line. Common-mode current on the feed line creates a non-zero current at the feed end of the antenna.

There are two currents in the feed line (phase is not shown in this drawing, only magnitude) that are *almost* the same magnitude. The currents differ by the amount of common-mode current, I_{CM}, that increases I_{Line1} and reduces I_{Line2} so that $I_{CM} = I_{Line1} - I_{Line2}$. I_{CM} is also the value of current at the end of the antenna wire. From the standpoint of a radiated signal, the EFHW antenna is not really "end-fed" at all!

Radiating currents in the EFHW antenna system consist of the current on the antenna wire *and* the common-mode current on the feed line. The EFHW is in reality somewhat off-center-fed. As you might imagine, the resulting radiation pattern is nearly omnidirectional and not very much like a classic dipole. That may be better for a portable station than an antenna with nulls along its axis. Nevertheless, the EFHW user should be aware of where the antenna system current is flowing!

Another thing the EFHW user should be aware of is that when an antenna system is unbalanced, *everything* that is connected to it and is not isolated by chokes or other methods (such as detuning) should be assumed to be part of the antenna system. A typical EFHW system includes not only the antenna and feed line but the ground connection, all of the radio equipment, and the operator when touching anything.

When running QRP, the resulting common-mode RF current and voltage on the equipment may not be very noticeable. But, at and above 100 W, RF voltages high enough to cause RF burns can be present at different points in the system — such as on the microphone or key! When this happens, you can sometimes "move" the hot spots around in the system by attaching quarter-wavelength wires to the high-voltage point. Because the high-voltage point moves to the open end of the wire, secure it where it won't be accidentally touched while in use.

Experiencing the System

You can demonstrate these effects with an SWR analyzer covering 2 meters and a simple experimental test antenna. Cut 38 inches of stiff wire or rod into two equal pieces (coat hanger wire will do nicely), creating a dipole antenna for 2 meters. Include about ¾-inch on each wire to form loops or use terminals as in Figure 2. To support the antenna and hold it steady for testing, I used a piece of ½-inch PVC pipe and a T fitting as shown in Figure 2. 8-32 screws through the fitting hold the antenna, as shown in Figure 2.

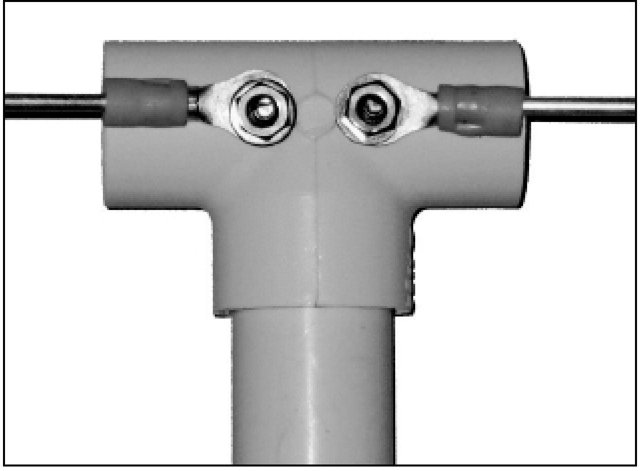

Figure 2 — A PVC T fitting becomes a dipole center insulator for experimenting with the effect of common-mode current in an antenna system.

Next, create an RG-8/X or RG-58 coaxial cable with an RF connector on one end and alligator clips on the other. (This is a handy antenna test accessory.) Any length from 6 to 10 feet is fine. Attach the SWR analyzer to the antenna with the coax and tape the coax along the vertical support for a distance of 4 or 5 feet. This stabilizes the antenna system and keeps it in the same configuration as you experiment — very important for antenna testing.

Locate an open spot where you can secure the antenna support to hold the antenna horizontally a couple of feet over your head. There should be a wavelength or so (6 feet) of clearance between the antenna and any conducting wires or metal surfaces.

Use the SWR analyzer to find the frequency at which the antenna is closest to resonance — mine was about 145.15 MHz. (You may not be able to find the X = 0 resonance due to the setup or analyzer performance.) Record both the resistive (R) and reactive (X) values, whichever are available from the analyzer. In my case, at 145.15 MHz the feed point impedance was measured to be 43 ± *j*20 Ω through 8 feet of RG-8/X cable. (Most analyzers don't show the sign of the reactance and that's not important for this experiment.)

Touch the fingers of one hand to the outside of the feed line and watch the impedance measurement as you move your hand up and down the feed line within a couple of feet of the feed point. You won't see a lot of variation (a few ohms for R and X) because the low impedance of the antenna feed point masks the effect of the coax shield's outer surface.

To create a high feed point impedance, tune the analyzer 20 or 30 MHz lower so that the feed point R is 300 to 400 Ω. Perform the same experiment of touching the coax jacket while observing the effect on feed point impedance. At 115 MHz, R varied from 290 – 390 Ω and X from 0 to 115 Ω. This wide variation showed the antenna system included the common-mode current path and any impedance added to it, such as fingers!

If you have some clamp-on ferrite cores, snap them on to the feed line just below the feed point, as shown in Figure 3. The common #43 mix used for suppressing VHF/UHF EMI is just right although other mixes will have some effect, as well. As you add each core, repeat the touch-and-observe experiment. By adding five cores, as shown in the photo, variation in impedance for my antenna was reduced to ±10 Ω of resistance and ±20 Ω of reactance.

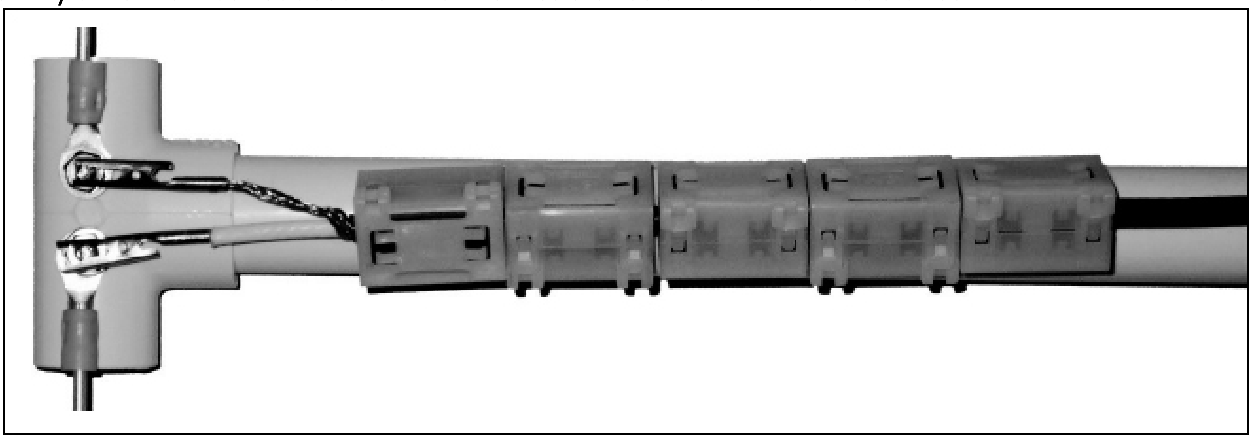

Figure 3 — Ferrite cores (#43 mix) added to the feed line increase the impedance of the common-mode current path, reducing its effect on antenna performance and feed point impedance.

The wide variations illustrate the sensitivity of a high impedance feed point to the presence of other conductors connected to and in the near field (within a few wavelengths) of the antenna. You can also see the effect of adding impedance to the feed line's common-mode current path with ferrite cores. By isolating the shield's outer surface with ferrite cores, the effect of common-mode current paths on the system can be reduced.

Be careful, however! Your EFHW system may depend on that common-mode current to reduce the feed point impedance of the antenna system. Without it, the EFHW feed point impedance will generally be much higher and difficult to match. It may be better to move any feed line choke closer to the transmitter and let the feed line between the choke and antenna radiate.

For another look at the EFHW through the eyes of an antenna designer, read AE6TY's *QRP Quarterly* article, "Refining an End Fed Antenna." [4]

By understanding the EFHW as an antenna system, you'll start to look at your other antennas as systems, too. This will help you plan, install, understand, and use your growing antenna farm more effectively, at home or in the field.

1

All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

EZNEC User Manual, version 5.0, **eznec.com/misc/EZNEC_Printable_Manual/5.0/EZW50_User_Manual.pdf**.

3

Lewallen, Roy W7EL, "Baluns, What They Do and How They Do It," **www.eznec.com/.Amateur/Articles/Baluns.pdf**

4

www.ae6ty.com/Papers_files/Refining%20an%20End%20Fed%20Vertical%20Dipole.pdf

Experiment #150 — Log-periodic Basics

The standard design for rotatable directional ham antennas has been the Yagi-Uda array, known today just as "the Yagi," nearly since its introduction in the late 1920s.
[1]

Chester Buchanan, W3DZZ, added parallel LC circuits, aka "traps," to dipoles and Yagis in 1955, putting rotatable directivity on 14 Mc and up within reach of the average station builder.
[2]

Then came 30, 17, and 12 meters. Hams wanted "pointable gain" on these bands, and that changed their antenna requirements dramatically. Suddenly, the *log-periodic* became an all-bands-on-one-boom solution.

Frequency Independence

One of the common claims for log-periodic antennas, most commonly a Log-Periodic Dipole Array (LPDA), is that they are frequency-independent. Why so? The fundamental idea (somewhat oversimplified) is that by defining an antenna entirely in terms of angles and ratios, it will behave consistently when scaled to any frequency. This is related to the notions of *self-similarity* and *scale-invariance.* If the antenna's structure remains consistent when scaled by some factor, the antenna's behavior with frequency turns out to be periodic (repeating) according to the logarithm of that factor. Thus the name, log-periodic.

In its most common amateur form, the LPDA consists of a set of linear $\lambda/2$ dipoles covering the lowest to highest frequency of the antenna's range, which is usually one octave at HF from 14 to 30 MHz. (Three octave ranges of 3 – 30 MHz are real monsters!) Tennadyne (
www.tennadyne.com
) makes HF logs and a five-octave 50 – 1300 MHz model.

So, where are the ratios? The three primary parameters of LPDA design are:

- Apex angle, α (alpha), which controls the shape of the triangular LPDA outline
- Scale factor, τ (tau), which controls the ratio between spacing and length of adjacent elements
- Relative spacing, σ (sigma), which controls how many elements fill the triangular outline

These three parameters completely define the shape and internal structure of an LPDA, whether it is intended for use at HF, VHF, microwave, or light. An LPDA designed from the same three parameter values will look the same at any scale.

The scale factor, τ, captures the relationship of L, R, and D for the antenna elements as illustrated in Figure 1:

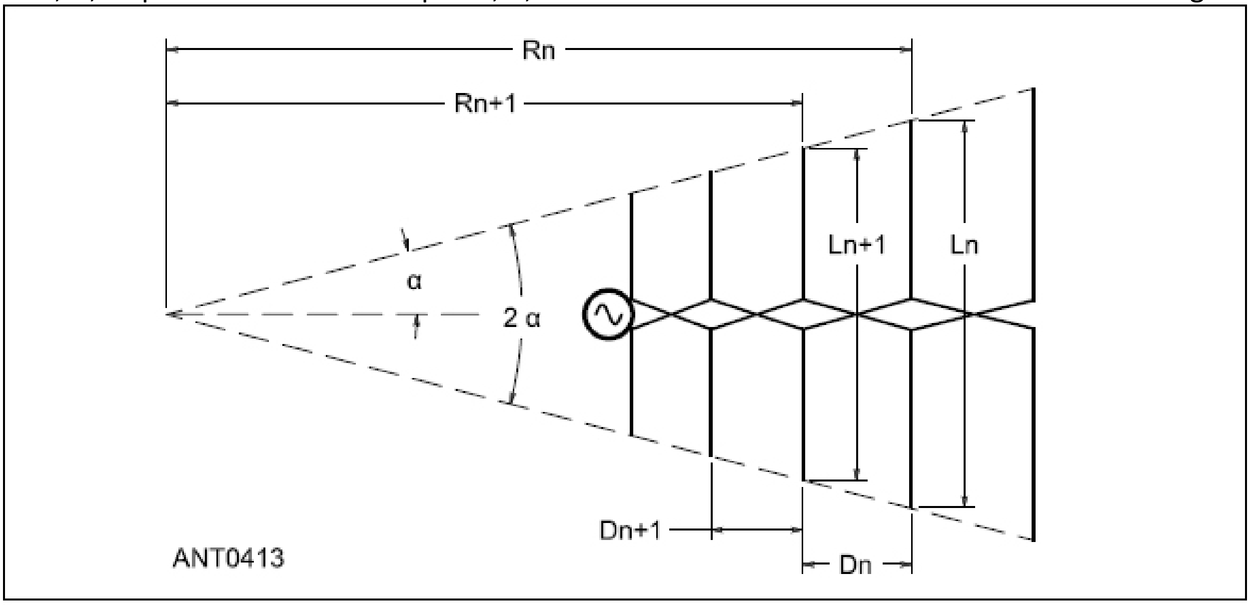

Figure 1 — The fundamental relationships which control the design and performance of a log-periodic dipole array (LPDA).

$$\tau = \frac{R_{n+1}}{R_n} = \frac{D_{n+1}}{D_n} = \frac{L_{n+1}}{L_n}$$

As τ increases, the elements get farther apart and the lengths of adjacent elements differ more.

The three parameters are related by the following equation:

$$\sigma = \frac{1-\tau}{4\tan(\alpha)}$$

By picking values for two of the parameters, the third can be determined — a lot like Ohm's Law.

Building a Log

First, we define the frequency range to be covered to determine longest and shortest dipole lengths (typically resonant a few percent outside the desired range). Then, we pick a boom length. This sets the overall size of the triangle and determines τ. Next, we have to specify how much “ripple” we can tolerate in the antenna’s behavior over that range. That determines how many elements will fill the triangle. We do this by choosing a value for σ or by specifying the number of elements. Figure 2 shows some examples of antennas which all have the same gain but cover different frequency ranges, or that have the same frequency ranges and different element spacings.

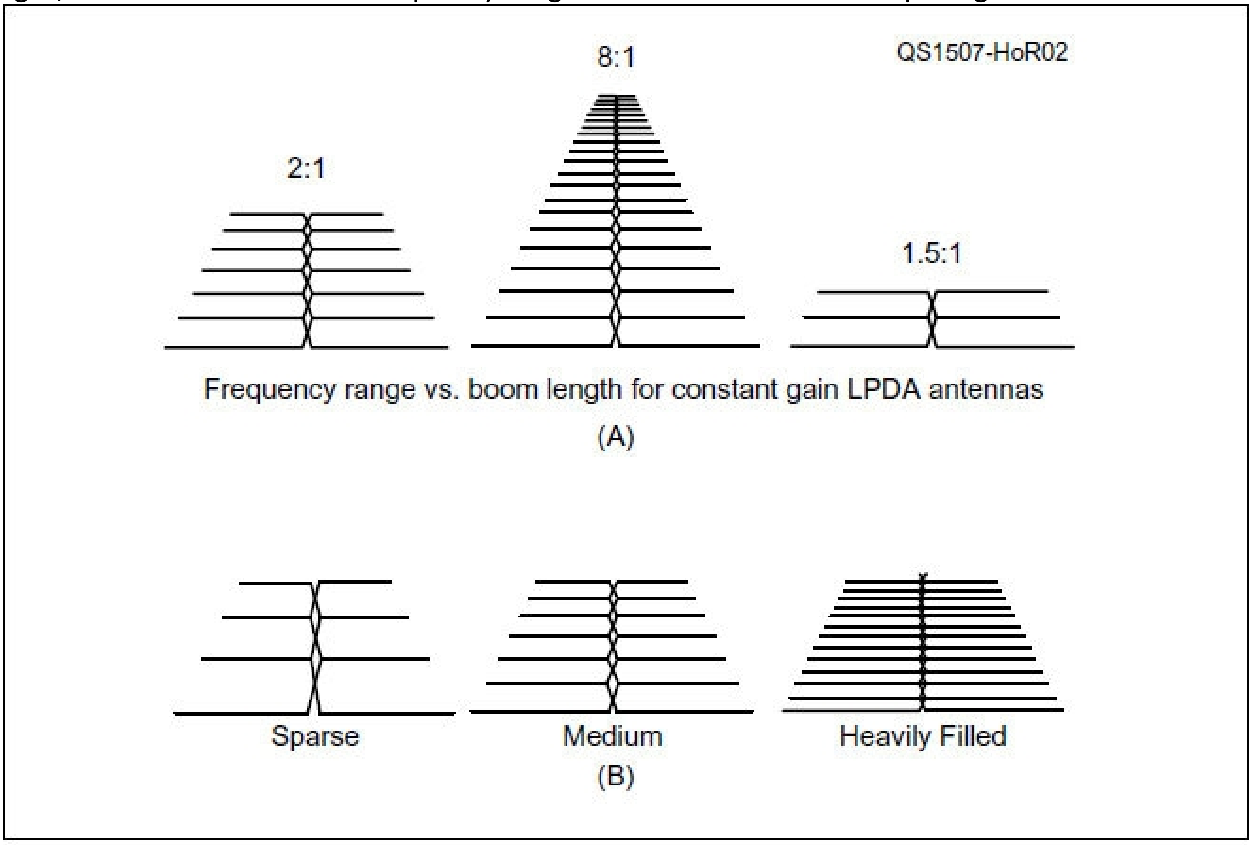

Figure 2 — At A are shown designs with equivalent gain but different frequency ranges. At B are shown designs with the same frequency ranges but different relative spacing (σ).

As a practical matter, we use charts or software to design the antenna. Figure 3 shows the most common chart used for log-periodic design. On the horizontal axis for τ, the triangle gets “pointier” toward the right. On the vertical axis for σ, toward the top there are more and more elements packed into the triangle. The slanted straight lines show different values for the apex angle, α.

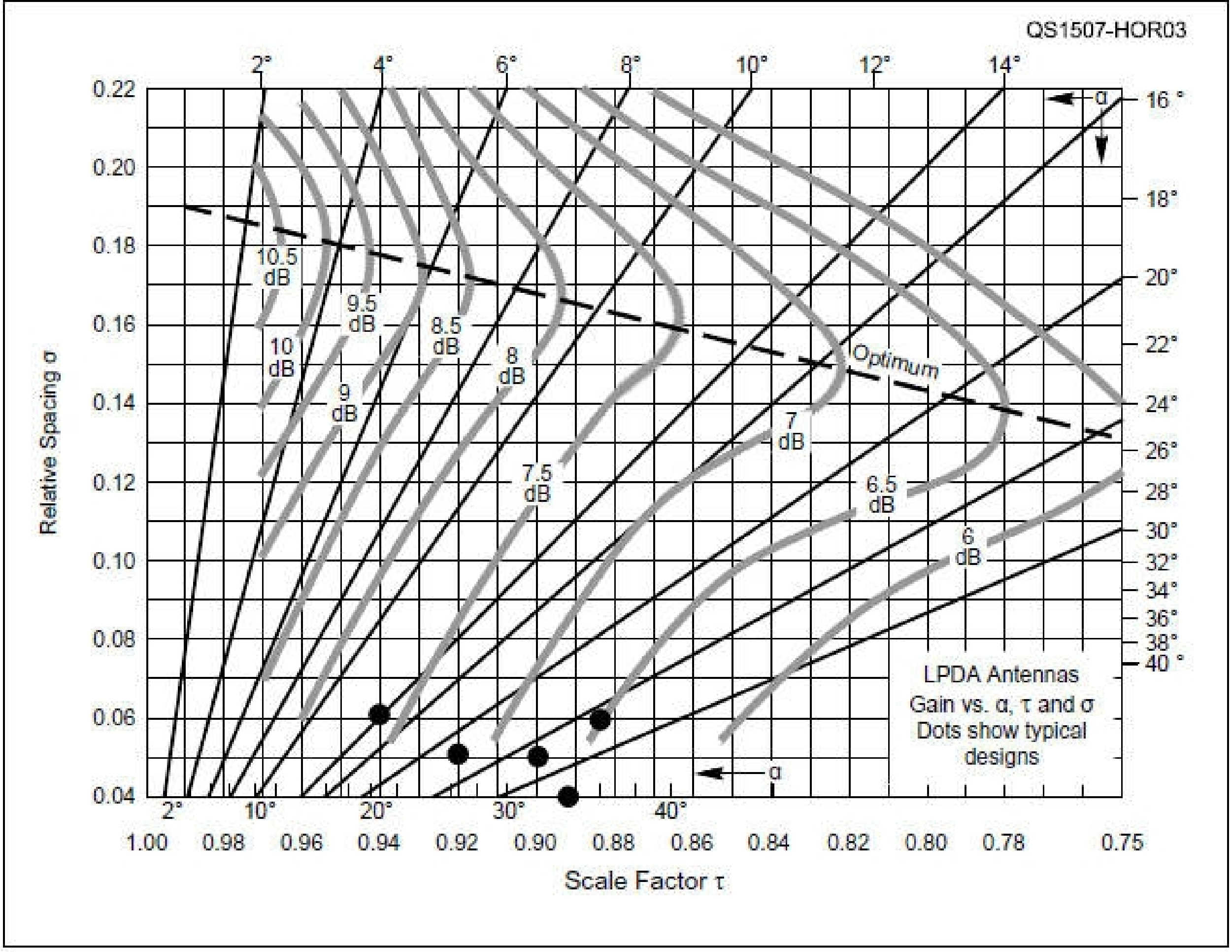

Figure 3 — Chart showing relationship between τ, σ, α, and gain. The dashed line gives the value of σ resulting in maximum gain for a value of τ. The black dots represent practical HF log period designs. The grey lines represent combinations of parameters yielding a constant gain. A full-size version of this chart is available online.[3]

Overlaid on top of the parameter scales are the curved grey lines. As on a topographic map, these represent combinations of the three parameters which result in the same values of gain, labeled for each line. The dotted line running across all of the curves shows the value of σ required to obtain the maximum gain for a particular value of τ. (This chart is based on a particular length-to-diameter ratio for the dipoles and characteristic impedance of the antenna.) Using the optimum value usually results in an antenna too large to be practical but typical designs (represented by black dots at the bottom of Figure 3) have acceptable performance.

The program *LPCAD* by Roger Cox, WBØDGF, is a more practical method of designing your own antenna (**wb0dgf.com/LPCAD.htm**). Using this software, after establishing the antenna's frequency range, you can enter values for τ and σ directly to see the results. (This is an easy way to find out why the optimum value of σ is impractical.) Or you can enter boom length and number of elements, which is a much more practical way of designing a log-periodic!

Feeding the Log

The LPDA is fed from the forward end of the array at the triangle's apex. For frequencies toward the middle or low end of the antenna's range, the "front" dipoles are very short compared to λ/2, and so will have a high impedance. A traveling wave develops as the signal moves along the transmission line toward the longer dipoles until it encounters the dipoles close to resonance, which are excited by the wave and radiate its energy. This *active region* moves back and forth with the operating frequency.

Phase reversal is key to the antenna's performance. If each successive dipole is fed out of phase with the adjacent dipoles, the array develops a *back-fire* pattern to the front of the array to the left in Figure 1. If phase is not reversed — a common error made by first-time log-periodic assemblers, such as myself — radiation is in the *end-fire* direction to the back of the array, resulting in poor SWR and gain.

Feeding successive elements out of phase can be accomplished by constructing the boom from a pair of conductive tubes insulated from each other forming a parallel conductor line. The feed line runs through one tube and is connected to the parallel conductors at the front of the array. Another method, more common in smaller TV antennas and in really large military or commercial LPDAs, is to insulate all of the elements from a single supporting boom and use crisscrossing straps to connect all of the elements.

Filling Your Log Book

This column just scratches the surface of log-periodic and frequency-independent antenna design. There are dozens of designs in common use from MF through mm-wave as described in *The ARRL Antenna Book* and numerous other references.[4],[5]

Try downloading a log-periodic manual online and entering the element lengths and spacings to see what value you obtain for σ, τ, and α, then find where that design falls on the chart. You'll never look at a TV antenna the same way again!

References

[1] The first amateur to use a Yagi was 1CCZ in 1928. His neighbors thought it was either a Ferris Wheel, a dirigible, or a ship ("Strays," *QST*, Oct 1928). The Yagi was described a few months earlier: H. Yagi, "Beam Transmission of Ultra Short Waves," Proc. IRE, Jun 1928, Vol 26, pp 715 – 741.

[2] Buchanan, C., W3DZZ, "The Multimatch Antenna System," *QST,* March 1955, p 22.

[3] All Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

[4] *The ARRL Antenna Book*, 22nd edition, Chapter 7, ARRL.

[5] Johnson & Jasik, *Antenna Engineering Handbook*, 2nd edition, Chapter 14, McGraw-Hill.

Experiment #165 — Propagation Prediction

While we all knew Solar Cycle 24 was headed to a solar minimum sooner or later, over the past few months there has been a lot of hope that maybe there would be a few more months of sunspots. Well, the reality is that there have been quite a few spotless days lately, and we are assured of more to come. Luckily, there are great tools and more data than ever before to help you make your on-the-air minutes count. This month, we're going to learn about an excellent and free resource, *VOACAP Online*.

Propagation Prediction — Then and Now

Before the Internet, the only information available to most hams were the hourly announcements from WWV and WWVH, of the solar flux and K indices along with the condition of the geomagnetic field.
[1]

A simple chart gave you some idea of whether conditions were better or worse than "normal," whatever "normal" meant. Numeric tables of monthly estimates generated by programs like *IONCAP* (Ionospheric Communications Analysis and Prediction Program) and the early coverage maps by pioneering PC programs like *W6ELProp* have been replaced with sophisticated graphic presentations a far more nuanced view.

The network of NCDXF beacon stations (
www.ncdxf.org/pages/beacons.html
) is as useful as ever, but is supplemented by the worldwide Reverse Beacon Network (
www.reversebeacon.net
) of automated receivers, including signal reports of beacon signals. Reception reports ("spots") once distributed by packet radio bulletin-board systems using *PacketCluster* software are now available to all via Telnet connections and on websites like
dxmaps.com
,
dxsummit.fi
, and
dxheat.com
. Get on the air, call CQ on CW or RTTY, and within a few seconds, your presence will be made known worldwide!

With all this information around, who needs predictions? Unless you have 24 hours a day to spend watching a computer screen (or watching for DX alerts coming in by text message) you need to decide when you'll be at the rig. With sunspots on the decline, band openings above 10 MHz will decline too. While the lower-frequency bands improve for long-haul contacts with declining solar flux, you still need to plan for the short openings between daytime absorption and the MUF (maximum usable frequency) falling at night. Good planning makes for happy hams!

VOACAP Online

VOACAP (Voice of America Coverage Analysis Program) was developed to predict broadcast coverage using detailed ionospheric models and the continually improving understanding of interactions between the Sun and Earth's geomagnetic environment.

While you can download *VOACAP* and run it on your PC, the software has been made available with an online interface (
www.voacap.com
) by Jari Perkiömäki, OH6BG/OG6G. It manages a lot of the setup and configuration so that the program is easily usable by beginners. (Once you are familiar with the online version, the PC-based version will allow you additional flexibility and customizing to suit your station more exactly. A user's manual is available on the *VOACAP Online* home page.)

Coverage Area Maps

Let's start with a map of locations for which a particular band is expected to support contacts. Browse to
www.voacap.com/coverage.html
or click COVERAGE AREA MAP on the *VOACAP* home page. You'll see a screen like that shown in Figure 1. For first-time visitors, it will be centered on the "East Pole" (the intersection of 0º E and 0º N). Start by selecting a QTH in the Transmitter Site panel: I selected Jefferson City, Missouri, the closest menu choice to my home. I then selected the transmitter parameters — antenna (dipole at 33 feet), power (100 W), and mode (CW) — and a band (14.1 MHz). The website automatically loads the SSN (smoothed sunspot number) from a solar observatory (35 on the day the map was generated) and uses the date and time of the PC's clock (2100 UTC on 31 July 2016). The receiving antenna is assumed to be a dipole at 33 feet, as well. Accept those defaults for now.

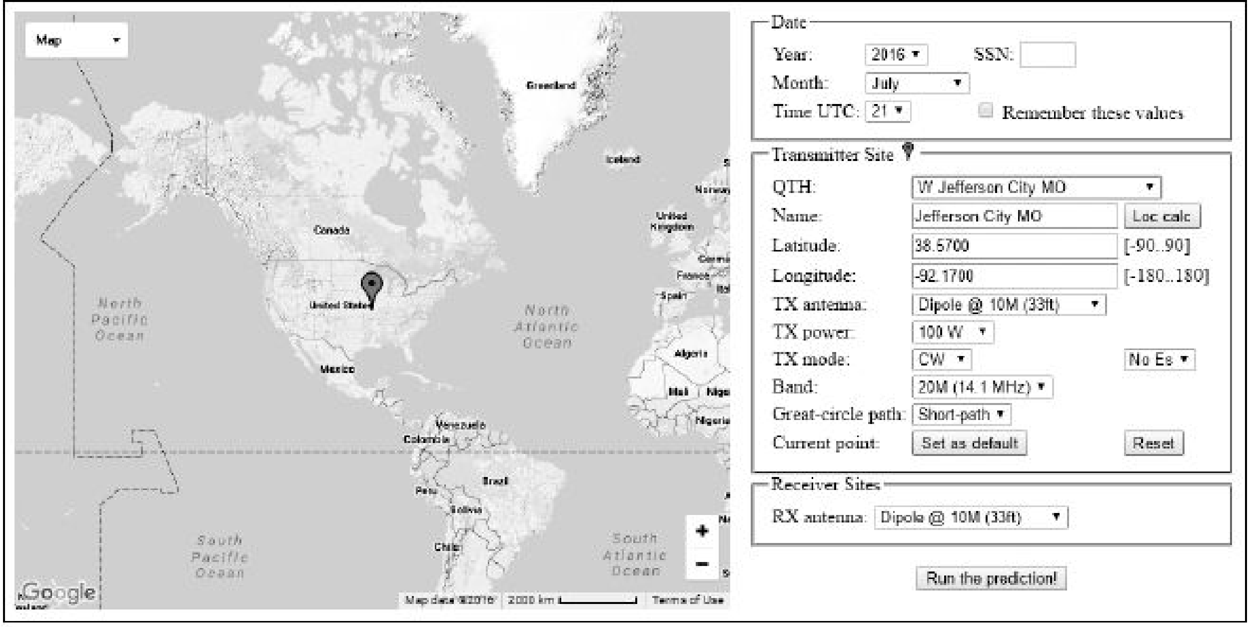

Figure 1 — This *VOACAP Online* configuration screen allows you to select your transmitting location and basic station parameters, and automatically obtains solar data online.

Click the RUN THE PREDICTION! button to see where 20 meter CW might be open between these two types of stations, and after about 5 seconds of computation you will see a map like that in Figure 2. (Note the skip zone [dark area] around the transmitting location indicated by the bright red dot.) The brighter the color, the more likely it is that you'll be able to make contact with the signal qualities built in to the online software. (You can configure "quality of service" values in the PC-based version.) Even on a summer afternoon with solar activity the minimum and using low dipoles, there should be opportunities to make contacts galore! Why not get on and call CQ or tune around?

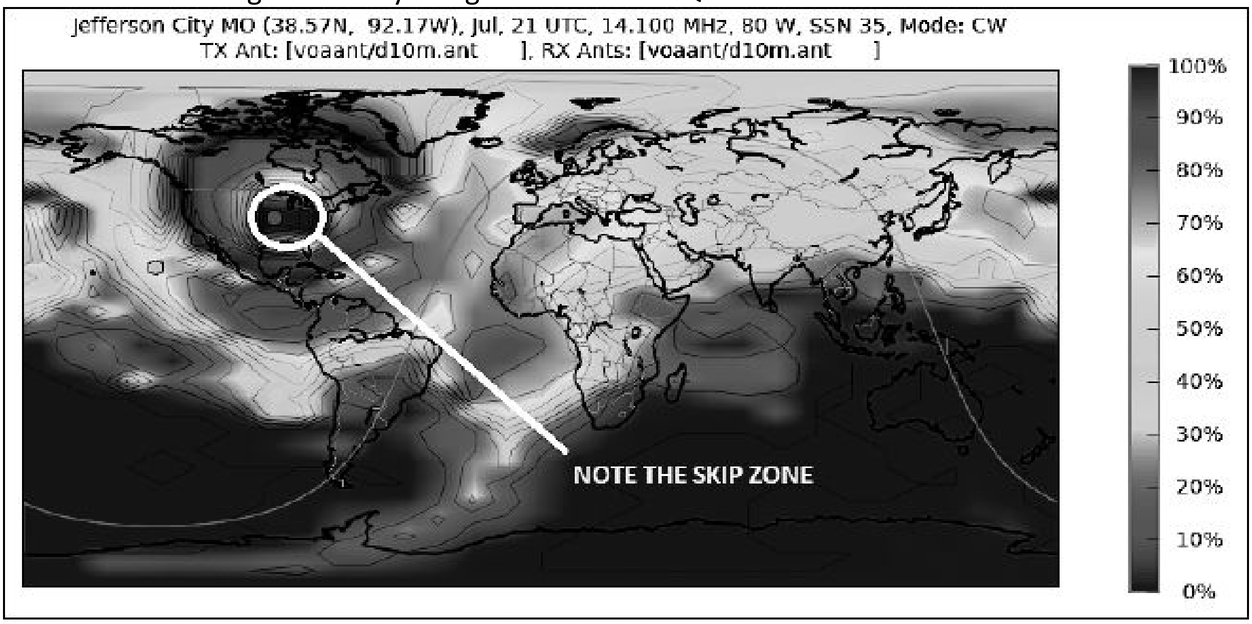

Figure 2 — A coverage map for 14 MHz at 2100 UTC on 31 July 2016 with a SSN of 35. Even with a simple dipole antenna on the receiving end, the band is open to North and South America, Africa, and much of Europe if the geomagnetic field is stable.

Here are your first assignments: Vary the transmitting parameters to see what effect they have on your coverage map. (All of the antennas are assumed to be oriented in the preferred direction when calculating signal strength.)

- What happens if you switch from CW to SSB? (Spreading your signal over a 3 kHz bandwidth instead of 300 Hz certainly reduces your available coverage!)
- What happens if you switch from the low dipole to a ¼-wave vertical with a good ground system? (This could be a significant improvement.)
- Experiment with raising the dipole in 10-meter increments. At greater heights, why do secondary "holes" in coverage appear? (The elevation pattern of the dipole breaks up into lobes and nulls between them.)
- Try small Yagis at different heights, too. Extra credit for experimenting with long path (selectable in the GREAT-CIRCLE PATH menu) to see how much power and antenna it might take to work, say, Japan the long way around!
- Now change the time through the day from around sunrise at your location to night and watch the effects of the Earth's rotation. (The Sun's position is shown as a yellow dot on the map.)

Point-to-Point Predictions

Let's say you do want to see when the bands might be open to a particular location. Maybe there is a DXpedition on, or you might have a friend with whom you make regular contacts. *VOACAP Online* will "run the numbers" between two points as well. Browse to

www.voacap.com/prediction.html

and enter your transmitter information as before. This time, select a receiving location such as PY1 — Rio de Janeiro, and you'll see the hour-by-hour band availability chart update to that in Figure 3.

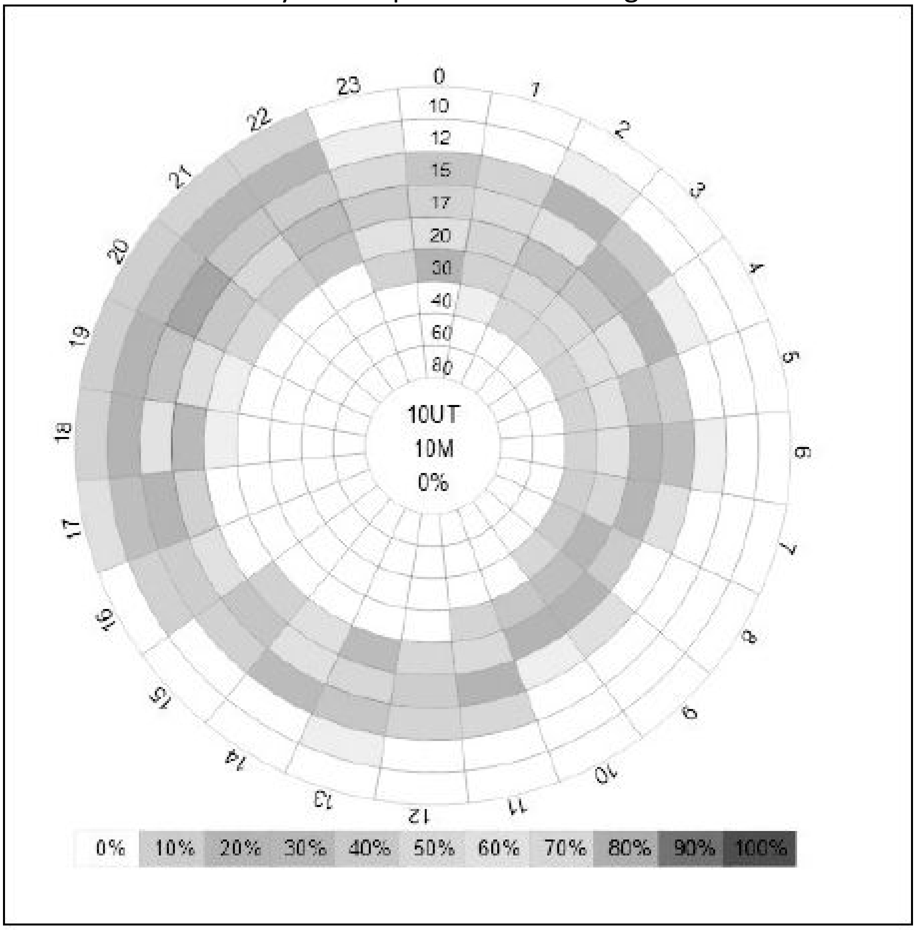

Figure 3 — Hour-by-hour band chart for the path between Jefferson City, Missouri and PY1, Rio de Janeiro, with the same transmitting parameters and date/time as Figure 2. While 15 meters has the best opportunity for a contact at 2000 UTC, there are many other bands and times when contacts are quite possible.

The bands from 80 through 10 meters each have their own ring showing the expected communication probability for each hour throughout the entire 24-hour day. At least one of the HF bands is open between WØ and PY1 at all times!

Now click the button at the lower right, labeled RUN PREDICTION! to see a bands-by-the-hour chart of propgation reliability.

Planning for propagation is a lot of fun — not as much fun as you'll have getting on the air and making QSOs you've investigated online, but nothing whets one's DX appetite like a chart saying, "Come and get it!" Once you're up to speed on *VOACAP Online*, give the PC-based version a try and you'll find it to be a very valuable tool in your hands-on repertoire.

[1] See tf.nist.gov/stations/iform.html for the schedule of information transmitted by WWV and WWVH.

Experiment #176 — Dipole Feed Points

The ordinary half-wavelength dipole is one of the best (and oldest) antennas — hard to beat for good performance, ease of construction, and bang for the builder's buck.

Dipole Fundamentals

When the dipole is 1⁄2-wavelength long it acts just like a vibrating string at its *fundamental* frequency. Find a string, stretch it tight between a pair of sturdy supports a foot or two apart, and pluck the string with your finger. Look closely at it from the side. You will see the string's maximum displacement occurs in the middle, gradually reducing to zero at each end. Imagine that displacement represents electrical current, and you have a mental picture of a dipole at its half-wave resonant frequency. Current flows in one direction (just as the string is displaced in one direction) for half of a cycle, goes to zero, then reverses and builds to a maximum in the other direction for the second half cycle.

Figure 1 shows the dipole's current and voltage along the antenna. When current is flowing from left to right, voltage on the left side is higher than on the right, and vice versa. This is the root of the dipole's name; di– (meaning "two") and –pole (meaning "electrical polarity"). The two "poles" of the dipole (the left and right halves) have opposite electrical polarities so that current is always flowing from one to the other. The polarity reverses with every half cycle, so the direction of current flow also reverses.

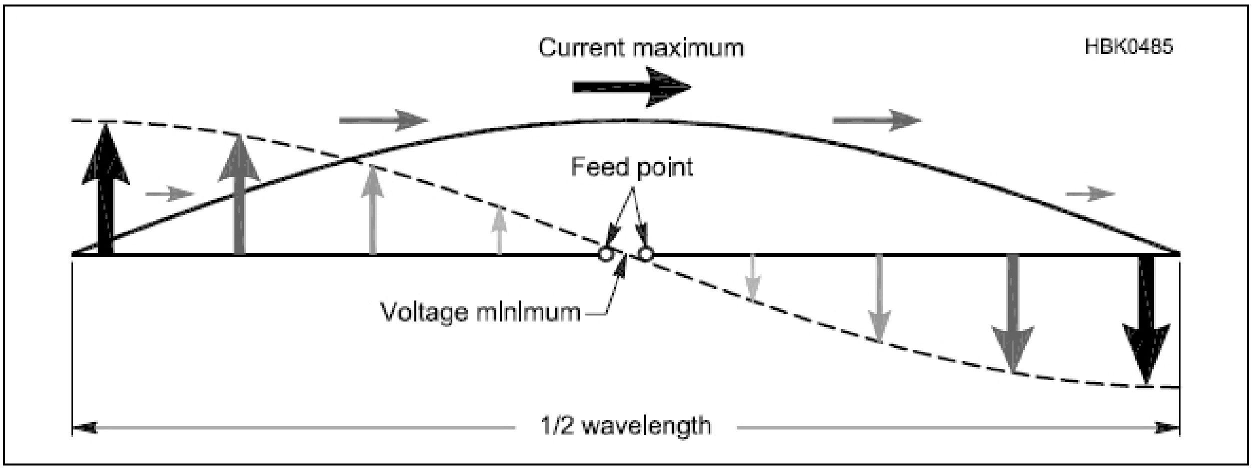

Figure 1 — Current and voltage distribution along a half-wave dipole. Impedance at any point is the ratio of voltage to current, so it is the highest at the ends and lowest in the middle.

The Feed Point

Note that I haven't mentioned the feed point yet — that's because it doesn't affect the basics of how the dipole "works." Just as you can pluck an instrument string anywhere along its length, you can feed a dipole anywhere as well. Plucking the string in various places may affect how loud it sounds as a result, but it doesn't change the fundamental frequency. Different points along the string are more effective for making the string sound louder or softer — but still at the same frequency.

For a dipole, the feed point is simply where you attach the feed line. It can be anywhere along the dipole — centered, off-center, or at the end. The feed point is just the place where you "pluck the string" and apply energy to the antenna. The dipole itself doesn't care. Changing the feed point of a resonant, half-wave dipole doesn't change the distribution of current and voltage along the dipole or its radiation pattern or its resonant frequency. (This assumes no current can flow back down the feed line as common-mode current.)

Impedance (Z) is the ratio of voltage (V) to current (I), whether at an antenna feed point or in a circuit. If the feed point is at the center, voltage is minimum and current is maximum, so we expect the impedance to be low, as well. And it is — the impedance of a resonant half-wave dipole at the center is approximately 72 Ω in free space.

If you move the feed point away from the center of the dipole, however, the situation changes. Voltage begins to increase and current to decrease, so the farther away from the center of the dipole, the higher impedance gets. If the feed line is attached somewhere off-center, the higher feed point impedance means that the feed line should have a higher characteristic impedance, if energy is to be efficiently transferred to the dipole. Toward the end of the dipole, impedance gets very high, indeed, perhaps a few thousand ohms. This can make it challenging to transfer energy to the dipole near its end. (See Hands-On Radio Experiment #136: "End-Fed Antennas."[1])

That said, the half-wave dipole simply doesn't care where you attach the feed line. It will have the same voltage and current distribution if excited with the same amount of power no matter where that power source is attached.

Multiband Operation

For single-band, coax-fed operation, there's no electrical reason to feed the dipole anywhere else but the center; the impedance is a good match to 50 or 75 Ω feed line, and it's mechanically simple. Moving the feed point away from the center makes the antenna system asymmetrical. This results in the feed line picking up a lot of common-mode current, requiring choke baluns and other techniques to isolate the line. So for a single-band dipole, leave the feed point centered.

For multi-band operation, the situation is quite different. Figure 2 shows the current and voltage distribution along the same dipole on its second (blue) and third (red) harmonic. At the second harmonic, the central feed point impedance is now very high, because voltage is high and current is low. You can see a number of high- and low-impedance points along the dipole for the different frequencies. Feeding the dipole at the center means low feed point impedance on the odd harmonics, beginning with the fundamental and high impedances on the even harmonics. The result is a severe impedance mismatch on at least half the bands. If the feed line is low-loss (such as heavy open-wire line [2]), that may be acceptable, and an antenna tuner can be used at the transmitter.

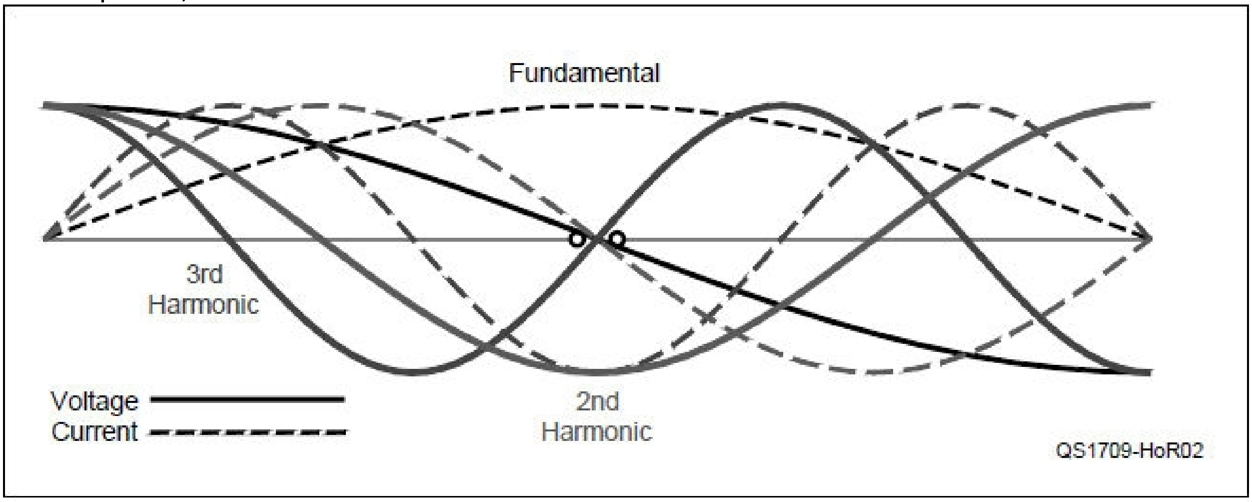

Figure 2 — Current (dashed line) and voltage (solid line) along the dipole at the second (blue) and third (red) harmonics. Where voltage is high and current low, that is a high-impedance point and vice versa.

Hitting the Spot

More often, though, a high standing-wave ratio (SWR) is not acceptable due to the high losses it causes in coaxial cable. The compromise is to find one spot along the dipole where feed line impedance is similar on different bands. An impedance transformer converts the impedances to a value that does not create a high SWR in coaxial cable. The SWR will not be 1:1, but the feed line loss will be modest and the impedance will be within the range of most tuners.

The most common such design places the feed point at 1⁄3 of the total dipole length from one end and uses a 4:1 impedance transformer. This results in SWR of less than 2:1 on the fundamental, second, and fourth harmonic, e.g. 40, 20, and 10 meters. (The feed point impedance depends on height above ground, as well.) A current choke should be used at the feed point to block common-mode current and decouple the feed line from interacting with the antenna. Several commercial antennas are available in this configuration and the antenna is popularly known as the off-center-fed dipole (OCFD).

Another popular configuration is to feed the antenna with open-wire line and use a tuner to match the feed line impedance to 50 Ω. There will be more feed line loss than matching at the antenna, but the convenience of using a single feed line outweighs the tolerable performance loss.

In 1996, K1POO analyzed the OCFD and came up with an alternative feed point position at approximately 1⁄6 of the length from one end. [3] Figure 3 shows a conventional OCFD and the K1POO design.

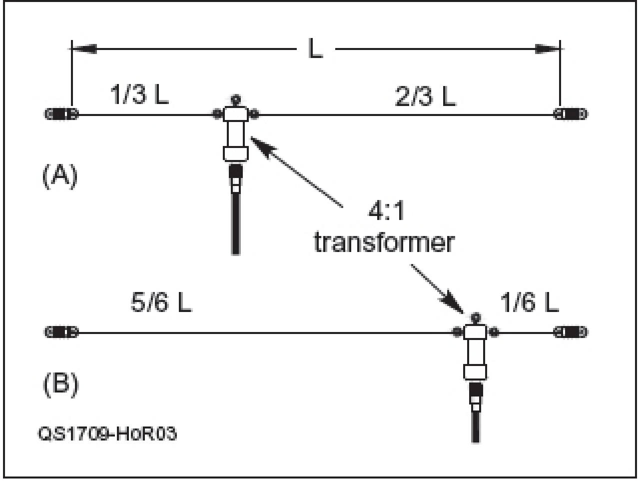

Figure 3 —The convention ⅓ – ⅔ off-center-fed dipole design (A) and the K1POO ⅙ – ⅚ design (B). Each uses a 4:1 impedance transformer. A current choke on the feed line (not shown) is required at the output of the impedance transformer to decouple the antenna and feed line, minimizing interaction.

Children of a Common Mother

What's particularly important to realize, and the reason for this discussion, is that all three of the antennas we've discussed — the half-wave dipole, the end-fed half-wave (EFHW), and the off-center-fed dipole (OCFD) — are the *same antenna!* The only difference is where the feed line is attached. Assuming they have the same amount of power applied to them, they all have the same radiation pattern and gain. Certainly, you may prefer one design over the other for mechanical or aesthetic convenience. Our oldest antenna friend, the dipole, however, will work just as it always has.

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at www.arrl.org/hands-on-radio.

2

B. Shackleford, W6YE, "Custom Open-wire Line — It's a Snap," *QST*, July 2011, pp. 33 – 36.

3

rsars.files.wordpress.com/2013/01/k1poo-4-band-ocfd-40-20-15-10m-richard-formato-iss-1-3.pdf

Experiment #131 — Coax-to-Open Wire Balun

I recently designed some Extended Double Zepp (EDZ) antennas that present a reasonable SWR on 14 and 21 MHz. The design uses a specific length of 450 Ω ladder line, resulting in an SWR of less than 2:1 at the end of the ladder line on both bands. Since that length was too short to reach the shack, I chose to transition from the 450 Ω line to 50 Ω coaxial cable. (The EDZ design will be presented in a future column or article.)

One can just connect the coax to the ladder line and hope for the best — it might work, as some designs for multiband antennas will function that way. Unfortunately, the *outside* of the coax shield is also connected at the junction of the two feed lines, creating a *common-mode current path* with impedance depending on the length of the coax and the operating frequency.

The basic idea is explained in Roy Lewallen's, W7EL, classic article "Baluns & What They Do," at **www.eznec.com/Amateur/Articles/Baluns.pdf**. If you haven't read it, this would be a good time to do so.

The Case for Using a Balun

Roy's article shows why a current or choke balun is needed at the transition from the coax to a dipole with the wires at right angles to the coax. What if instead of a dipole, the coax is connected to ladder line? Is a balun still necessary? In transmission lines, the conductors are tightly coupled so that the currents are equal and in opposite directions.[1] That means the same current should flow on the inside of the coax shield and the conductor of the ladder line to which the shield is connected. If any of the current escaped on the outside of the coax feed line as common-mode current, then the balanced current rule would be violated, upsetting the impedance presented at the junction of the two feed lines.

While the coupling of the two conductors in the feed line *should* be sufficient to guarantee balanced currents in each, it's a good idea to raise the impedance of the common-mode current path, especially because you don't know the impedance of that path. Common-mode current on feed lines can cause the antenna system to behave unpredictably.

There is another reason to add some common-mode impedance to the feed line — preserving the symmetry of the antenna system. With a balanced antenna such as a dipole or EDZ, *decoupling* of the feed line's common-mode current path from the antenna's radiated field is also important, as explained in W7EL's article. Since common-mode chokes are difficult to create for ladder line, I oriented that portion of the feed line at close to right angles from the antenna to preserve antenna balance. Adding a choke balun at the junction of coax and ladder line was the next step. (If the coax is parallel to the antenna, add a coiled-coax choke or two along the coax to minimize common-mode current all along the feed line.)

The choke balun can take many forms, as explained in the *ARRL Handbook* and *ARRL Antenna Book*.[2] I decided against the W2DU-style balun of many ferrite beads on the coax because of the expense, and against the coiled-coax balun because it is somewhat heavy and unwieldy when suspended by the feed line (particularly if form-wound). It is also hard to create a scramble-wound choke that works well over the range of 40 to 10 meters (the EDZ is tunable on the WARC bands and 40) so I selected a compromise between all three designs.

Balun in a Jiffy

My choke balun was easily wound on a ferrite toroid core, using a bifilar winding that is really just a very closely spaced parallel-conductor feed line. By using the right mix of ferrite, the choke will create enough impedance across the HF range.

Following the guidance of Jim Brown's, K9YC, tutorials on ferrites and chokes, I chose a 2.4 inch diameter #31 mix with a winding of 10 turns.[3], [4] The ferrite tutorial estimates that the balun's choking impe-dance at 7, 14, and 28 MHz is 3000, 3500, and 2000 Ω, respectively, as shown in Figure 1. For the bifilar winding, I used two-conductor PVC-insulated #16 zip cord which is fine for 100 W power levels. (If you plan on running high power, use #12 or larger wire.) For this core you need about 3 inches of wire per turn plus the connections at either end for a total of about 36 inches of wire, including the input and output connections.[5]

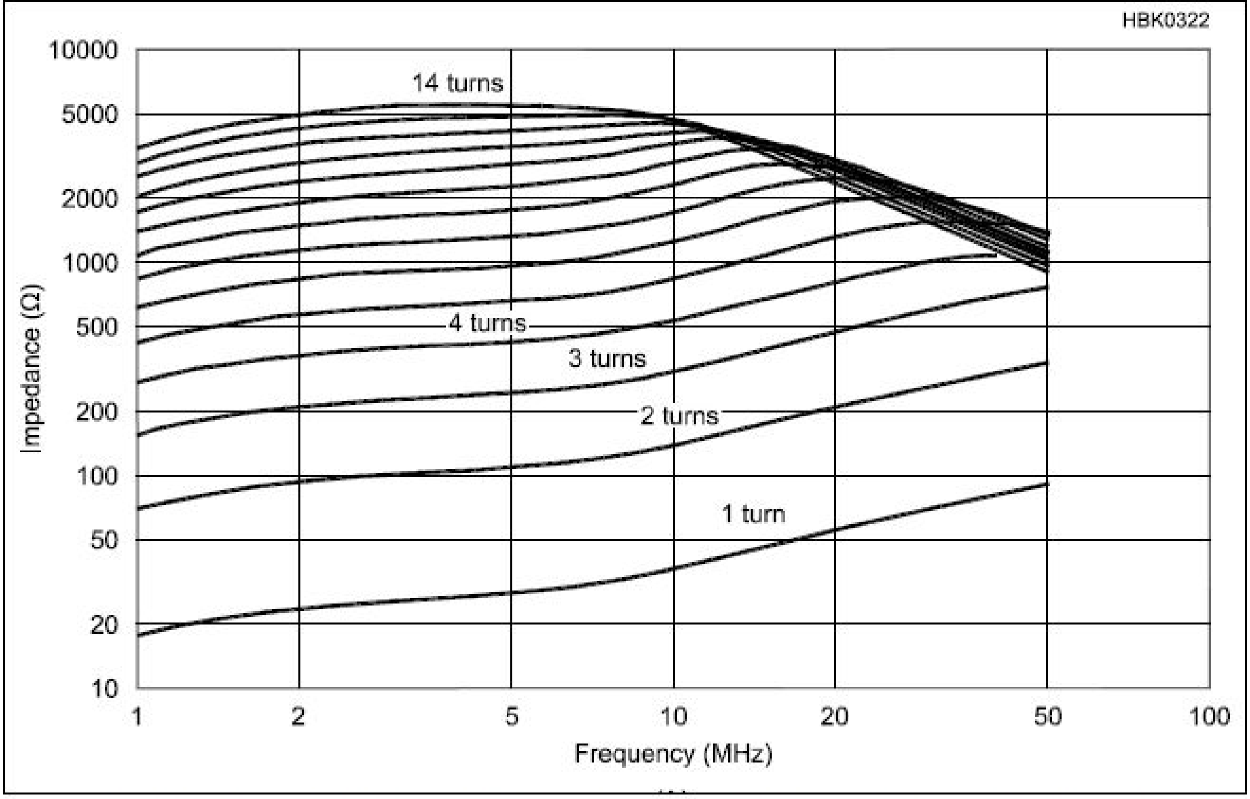

Figure 1 — Data from K9YC shows the broad response of #31 mix with graphs of impedance vs number of turns on a 2.4 inch OD ferrite toroid.

The balun would be installed outside, suspended in mid-air, with up to 30 feet of coax hanging from the balun. Therefore, I needed a lightweight, non-conductive enclosure that could accommodate an SO-239 connector and the ladder line. New enclosures all seemed to be rectangular, heavy, and expensive. PVC pipe and caps would be *really* heavy. While sorting through a bag of empty food containers that I use to hold parts, I found my balun enclosure in the form of a peanut butter jar.

The 16 oz size turned out to be perfect for a 100 W balun and 28 oz jars are large enough for high-power models (Figures 2-4 show balun assembly). The clear jar is tough and a 2.4 inch toroid fits inside after winding, although you have to squeeze the jar a bit to get it through the threaded part of the jar. To get the ladder line through the lid, punch some holes with an awl or small drill bit. Drill or cut a hole in the bottom of the jar that is a little bit bigger than the shell of a PL-259 connector. Drill three or four small holes around the bottom of the jar for drainage. For UV protection, spray paint the jar and lid with outdoor enamel.

Figure 2 — Use a hole saw, chassis punch, or hobby knife to cut a hole in the bottom of the jar large enough for a PL-259 to go through. Use an awl or small drill to make two holes in the lid for the ladder line conductors. Drill small holes around the bottom of the jar for drainage.

Figure 3 — A close-up of the balun showing the SO-239 connector with winding soldered to the sheet metal screw.

Figure 4 — A version of the balun with the crossover winding. Notice how the input and output connections are in line with the feed lines.

Begin winding by securing the first turn with high-quality electrical tape such as Scotch 33+ or a wire tie. Then wind 10 turns on the core, making sure each turn is snug on the core, securing the final turn. I have tried both a single end-to-end winding and the crossover style of winding introduced by W1JR in which after half the turns are wound, the winding crosses through and over to the opposite side of the core, then continues to the point opposite the first turn. The crossover winding has little effect at HF but it conveniently places the input and output connections on opposite sides of the core. This makes the balun easier to assemble and holds it straight between the top and bottom of the jar. Both styles work fine in this use. The input and output leads should be short enough (about 1 inch for the low-power version) that they are not bent against the jar with the lid on.

Figure 2 — Use a hole saw, chassis punch, or hobby knife to cut a hole in the bottom of the jar large enough for a PL-259 to go through. Use an awl or small drill to make two holes in the lid for the ladder line conductors. Drill small holes around the bottom of the jar for drainage.

To attach the winding to the SO-239, tin the hollow tip of a #4 self-tapping sheet metal screw. Then place a small amount of anti-oxidation compound such as Penetrox on the screw threads and turn it into one of the SO-239 flange holes. (A #6 screw also works but you'll probably have to drill out the SO-239 hole a little bit, depending on the manufacturer.) Then solder one winding wire to the screw and the other to the SO-239 center conductor. Attaching the SO-239 to the jar with more sheet metal screws during installation is optional.

To test the balun before attaching the ladder line, solder a 47 or 51 Ω resistor across the output winding and use an antenna analyzer to measure the balun's input impedance. It should be close to 50 Ω with an SWR of 1:1. Move your hand along the coax and make sure the SWR doesn't change, a symptom of common-mode current on the coax. Polarity of the input and output windings is not important unless you are making a set of baluns in which case you should be consistent in how the windings are attached to the SO-239 and ladder line.

Poke the ladder line conductors through the lid, and then use needlenose pliers to curl the wire into a circle or U for soldering. Solder the output leads to the ladder line wires. If you want, coat those connections with liquid electrical tape or aquarium RTV sealant. Leave the lid off for now.

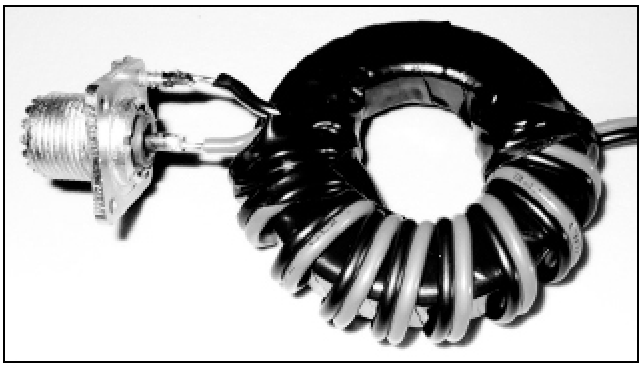

Figure 3 — A close-up of the balun showing the SO-239 connector with winding soldered to the sheet metal screw.

To install the balun, insert the coaxial feed line's PL-259 through the hole in the jar and screw it on to the SO-239. Use the coax to pull the core into the jar until the lid is against the threads. Screw the jar back into its lid. The PL-259/SO-239 should turn freely and not bind in the hole. Waterproof the coax connectors and you are done!

Parts List

2.4" #31 mix ferrite toroid core (Fair Rite 2631803802, Mouser 623-2631803802)

36" of two-conductor, #16 PVC-insulated zip cord (Radi

oShack 55057440)

16 oz peanut butter jar

SO-239

#4 self-tapping screw

1

See the discussion of mutual induction and Lenz's law in Experiments #117 and #118. All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

The ARRL Handbook and *ARRL Antenna Book* are available from your ARRL dealer, or from the ARRL Store. Telephone toll-free in the US 888-277-5289, or 860-594-0355; fax 860-594-0303; **www.arrl.org/shop/**; **pubsales@arrl.org**.

3

audiosystemsgroup.com/RFI-Ham.pdf.

4

audiosystemsgroup.com/CoaxChokesPPT.pdf.

5

The 36" bifilar winding starts to have an appreciable electrical length at 10 meters and has an additional transforming effect on the impedance that increases with frequency.

Experiment #137 — Choosing a Feed Line

An antenna system begins at a radio's output connector and ends in the space around the antenna, including the ground and everything conductive that is nearby (in terms of wavelengths). One never uses just an antenna — rather, an antenna system! What often seems a simple task — measuring, cutting, and hanging up a skyhook — really involves many more questions than you might expect.

One of the questions you must answer is "What feed line do I use?" and, as with many technical endeavors, the answer begins, "It depends…" While the choice also involves cost and the mechanical aspects of installation, the focus of this column is on the electrical aspects. Assuming the feed line is adequately rated for the power levels you'll be using, the choice boils down to loss in the line.

Evaluating Loss with *TLW*

Recently, Dean Straw, N6BV, former editor of *The ARRL Antenna Book*, updated his incredibly useful *TLW* program (*Transmission Line for Windows*).

[1]

The result is a more accurate assessment of systems using open-wire feed line, aka ladder line or window line. Often used as the feed line for multi-band "doublets" (wire antennas fed in the middle but not necessarily resonant on any band used), open-wire line (OWL) is often assumed to be less lossy than an equivalent length of coaxial cable. With the new version of *TLW* available, it seemed like a good idea to check this out. (This discussion doesn't include the loss of effects of baluns anywhere in the system.)

Starting with the center-fed, half-wave dipole used on a single band (and maybe on the third harmonic, too), the use of coax is a good choice for moderate feed line length. The feed line SWR and associated losses will be reasonably low, it's easy to connect at both ends, and it lends itself well to switching arrangements between several antennas.

Let's have a look at an example of this type of antenna system: a center-fed, 66-foot long inverted V, 50 feet above average ground, with 100 feet of feed line. According to *EZNEC* (

www.eznec.com

), at 7.1 MHz where the antenna is very nearly ½-wavelength long, the antenna's feed point impedance is 65 – *j* 41 Ω or 77 ⍰-32°.

[2]

This load creates an SWR of 2.1:1 for 50 Ω cable. One hundred feet of RG-213 coax has a total line loss of 0.7 dB and the SWR at the transmitter end of the line is 1.9:1. (The online *VSWR-RL-Reflection Coefficient* calculator by Giangrandi is another excellent free tool.

[3]

)

If instead we fed the inverted V with the same length of 450 Ω OWL, the new version of *TLW* calculates SWR at the antenna to be 6.3:1 and the line loss as 0.43 dB. While it looks like OWL is slightly less lossy, that figure doesn't include any losses in the impedance matching unit required at the line's input where SWR is still 5.7:1. ARRL Lab measurements of full-legal limit antenna tuners shows an average loss of about 11% for an SWR of 4:1 on 40 meters, which is 0.7 dB.[3] The total system loss when using OWL in this antenna system is about 1.1 dB. The advantage swings slightly more to coax for shorter feed line lengths and to OWL as feed line length increases.

Tuned Feeders

You can also use an old trick — *tuned feeders* — to reduce the need for impedance matching. If the OWL is extended to the closest length at which it is an integer number of ½-wavelengths long, the input and output impedances of the line will be very nearly the same. One wavelength of OWL at 7.1 MHz is 126 feet. At this length, a 50 Ω transmitter output sees an impedance of 73 – *j* 41 Ω (SWR of 2.1:1) and system loss is reduced to 0.5 dB, slightly better than for the coaxial cable. Adjusting the line length to 128.1 feet brings the system almost exactly to resonance with an input impedance of 73 Ω, an SWR of 1.5:1, and total system losses of 0.56 dB. Remember that this will *only* work on the one band and at the design frequency — you may still wind up needing impedance matchng at the band edges but the SWR will be lower than with the 100-foot feed line.

Feed Lines for Multi-band Doublets

The fun really begins when the antenna is to be used on several bands, including those for which the antenna is not an odd number of ½-wavelengths long, occasionally creating extreme feed point impedances. George Cutsogeorge, W2VJN, created an *EZNEC* model for *The ARRL Antenna Book* that shows center-fed feed point impedances on the HF ham bands for a 100-foot long doublet, installed at 50 feet over average ground.

[4]

I used the new version of *TLW* to calculate losses and SWR at the antenna feed point for 100 feet of both RG-213 coax and 450 Ω OWL. All of the data is shown in Table 1.

Table 1
Feed Line Losses for Center-Fed 100' Flat-top Dipole, 50' High, Over Average Ground

Frequency MHz	Antenna Feed Point Impedance, Ω	Loss for 100' RG-213, dB	SWR at Antenna	Loss for 100' 450 Ω Line, dB	SWR at Antenna
1.83	$4.5 - j1673$	26.2	1738	11.9	496.4
3.8	$39 - j362$	5.7	62	0.9	18
7.1	$481 + j964$	5.8	49	0.4	6.7
10.1	$2584 - j3292$	10.4	134	1.2	16.8
14.1	$85 - j123$	1.9	5.6	0.55	5.2
18.1	$2097 + j1552$	9.0	65	0.88	8.1
21.1	$345 - j1073$	9.7	72	1.2	10.1
24.9	$202 + j367$	5.1	18	0.5	3.9
28.4	$2493 - j1375$	10.0	65	1.1	8.1

In this common situation it's not even close — feed line losses for OWL are uniformly lower than for coax, sometimes by a lot. That's not the entire story because you'll still have to provide some kind of impedance matching device at the transmitter most of the time. As we have seen, that adds additional losses, regardless of whether the feed line is OWL or coax. In general and except in extraordinary cases, the old advice is still good: OWL is best for feeding multi-band, non-resonant antennas from the perspective of system losses.

Feed Lines for Matched Antennas

Most of our ham-band antennas, whether Yagis, dipoles, or ground-plane verticals, are designed with a feed point impedance that matches 50 Ω coaxial cables. For these antennas, OWL would not be a good choice due to mechanical and impedance matching considerations. So, how does one choose the right coax for these antennas?

You could start with the manufacturer's specified loss per standard length (usually 100 feet) and pick the lowest-loss line you can afford. However, antenna system designers generally approach the problem from the standpoint of *allowable loss*. In other words, what is the maximum acceptable amount of feed line loss? They select feed line that for the required length will have less loss than the maximum amount. This is a somewhat harder problem to solve because of the calculations involved.

You're in luck, however, because Frank Donovan, W3LPL, has performed those calculations already. The results are in Table 2 (related tables are also available in *The ARRL Antenna Book, 22nd Edition*), which shows the length of cable that results in 1 dB of *matched loss* — the loss for a feed line terminated in its characteristic impedance.

To use these figures, begin by dividing the length of feed line you'll need to use by the maximum acceptable loss in dB. This determines the lower limit for the number of feet per dB of loss in the cable. Find the column showing the frequency at which you are working. Locate the entry in that column with the lowest value of feet/dB *greater* than the calculated lower limit. That cable is the lossiest you can use and still satisfy your total loss requirement. For example, if you can tolerate 5 dB of feed line loss in a 300-foot run at 440 MHz, your lower loss limit is 300 / 5 = 60 feet / dB. In the 440 MHz column, LDF4-50A hardline is the lossiest cable you can use — start shopping! Note that the values of feet/dB don't increase uniformly from bottom to top due to variations in cable performance over the wide frequency range shown in the table.

What if the cable is *not* matched? Additional loss will result as the power is reflected back and forth in the feed line. Figure 1 shows just how much more loss. This chart, developed by Joe Reisert, W1JR, combines matched loss and additional SWR-caused loss into a single set of curves.

[5]

To find the total loss of your feed line, start with the load SWR (at the antenna) on the horizontal axis. Travel vertically to the curve (or an interpolation between curves) corresponding to the matched loss of the feed line you've selected. Travel horizontally to the vertical axis to find the total attenuation of the feed line. If this total is still below your maximum allowed loss, you can use that cable. If not, find the point at which the maximum allowed loss and load SWR intersect. That will show the maximum matched loss you can accept. You can then work backwards using Table 2 to find the cable you need.

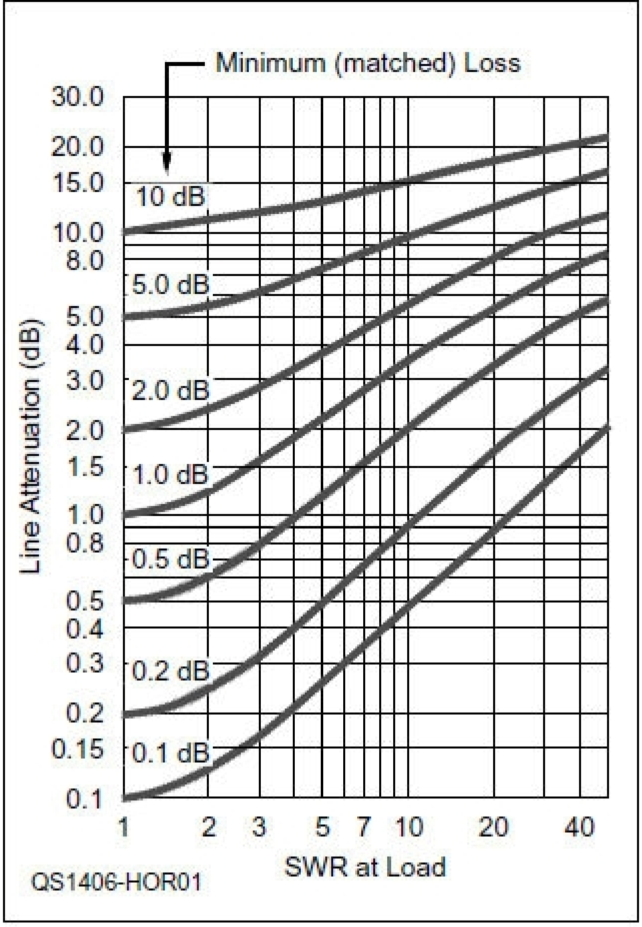

Figure 1 — Total insertion loss in a transmission line terminated in a mismatch. [Courtesy of Joe Reisert, W1JR]

Table 2
Cable Attenuation (feet per dB)

MHz	1.8	3.6	7.1	14.2	21.2	28.4	50.1	144	440	1296
LDF7-50A	3333	2500	1666	1250	1000	833	625	370	200	110
FHJ-7	2775	2080	1390	1040	833	667	520	310	165	92
LDF5-50A	2108	1490	1064	750	611	526	393	227	125	69
FXA78-50J	1666	1250	769	588	435	370	256	130	71	36
¾" CATV	1666	1250	769	588	435	385	275	161	59	33
LDF4-50A	1145	809	579	409	333	287	215	125	70	39
RG-17	1000	769	556	370	294	250	200	77	40	20
LMR-600	973	688	492	347	283	244	182	106	59	33
SLA12-50J	909	667	500	355	285	235	175	100	53	34
FXA12-50J	834	625	455	300	250	210	150	83	48	25
FXA38-50J	625	435	320	220	190	155	115	67	37	20
9913	625	435	320	220	190	155	110	62	37	20
LMR-400	613	436	310	219	179	154	115	67	38	21
RG-213	397	279	197	137	111	95	69	38	19	9
RG-8X	257	181	128	90	74	63	47	27	14	8
RG-58	179	122	83	59	50	42	30	18	9	5
RG-174	91	67	48	32	26	23	17	10	5	3

Feed line selection is not always so complicated. Most of the time you can just hook up the antenna and you'll work stations. To be sure you're getting the most out of your precious watts, though, these sorts of tools can help you make your decision.

1

Transmission Line for Windows, version 3.21, by Dean Straw, N6BV. See www.arrl.org/qst-in-depth for software availability.

[2] Joel Hallas, W1ZR, *The ARRL Guide to Antenna Tuners*, Chapter 14, ARRL, 2010.

[3] “Conversions between VSWR — Return Loss — Reflection coefficient,” Iacopo Giangrandi, **www.giangrandi.ch/electronics/anttool/swr.html**.

[4] *The ARRL Antenna Book, 22nd Edition*, Chapter 24, ARRL, 2012.

[5] Hands-On Radio Experiment #119 “The Q3Q Balun Redux.” All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

Experiment #157 — Matching — Network Design and Build

There comes a time in every ham's experiences when a single-band antenna needs to have its feed point impedance matched to 50 Ω. If practical, matching the antenna to 50 Ω at the feed point is the way to go and the subject of this month's column. You can apply the general process to your antenna farm, too.

I recently built two 60-foot towers to be used as base-fed monopole verticals on 160 and 80 meters. Both had significant top-loading from beams and mast extensions that lowered the frequency of quarter-wave resonance below the intended bands, so I decided to whip up an L network for each.
[1]

It's possible to just find a variable inductor and capacitor, hook them up, and start cranking in hopes of finding a match. But wouldn't it show a little more radio *savoir faire* to do a proper design?

Building by Design

Any design of a matching network begins with measuring the impedance you want to match, which sounds obvious but is often overlooked. This was a perfect opportunity to make use of my new SARK 110 Antenna Analyzer (www.sark110.com) to characterize what was happening at the base of my towers. Table 1 shows the impedances at selected frequencies across 160 and 80 meters.

The 160 meter tower consists of 60 feet of Rohn 25 topped by a C-31XR triband Yagi and 20 additional feet of 2-inch aluminum mast. Forty copper radials surround the bottom. The resonant frequency is 1.65 MHz (Z = 18.5 + *j* 0.03 Ω) and the resistance is stable between 15 and 17 Ω across the band with the positive inductive reactance climbing to 42.3 Ω at the top of the band. The average impedance across the band is 15.8 + *j* 32.1 Ω.

The 80 meter tower is the same height, topped with an EF-230 two-element 30 meter beam and an XR5-T multiband HF Yagi on 6 feet of mast above the tower. There were no resonances observed between 2 and 6 MHz although reactance does dip to –65 Ω just above 3 MHz. Across the 80/75 meter band, impedance varies quite a bit as shown in the table, averaging 116.9 – *j* 151.8 Ω.

Striking a Match

Once you have a good set of measurements (hold that thought) you can start coming up with a buildable network. I entered the average values of impedance into the online matching network calculator by John Wetherell, which gives component values for 16 types of networks.
[2]

For both bands, there were two unworkable versions of the L network: either the math "blew up," resulting in NaN (Not a Number), or negative values were generated for one or more of the component values.

Note that the Q value for an L network is fixed at the ratio of the source and load impedances. Q determines the matching bandwidth, and higher values of Q also means higher values of circulating current and peak voltage in the components. If the input and output impedances are very different, investigate networks that transform impedance in smaller steps.

Of the two remaining networks, I selected the circuit in Figure 1 for both networks, because I had a pair of heavy-duty 2000 pF vacuum-variable capacitors thanks to K9SD. The inductors I would wind myself, so I logged on to the K7MEM Single-Layer Air-Core Inductor Design website.
[3]

This calculator produces a fairly accurate single-layer, air-core design with lots of options for adjusting mechanical dimensions.

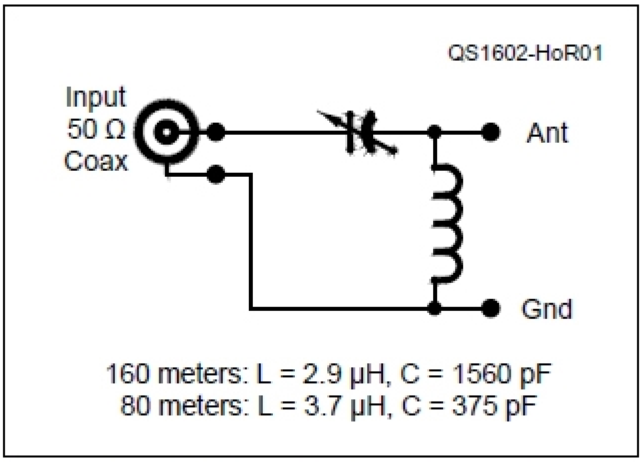

Figure 1 — The shunt-L input L network style (topology) works for both my 160 and 80 meter matching networks. Other values of load impedance may require a different L network configuration.

Figure 2 shows the completed 80 meter matching network. Wanting a low-loss, mechanically sturdy coil, I wound it on a 23⁄4-inch O.D. PVC coupling from 20 feet of 1⁄4-inch O.D. soft-copper refrigeration tubing. (This tubing is available inexpensively from hardware stores in a number of diameters.) The 20-foot length was enough tubing for a couple of extra turns. I soldered quick-disconnect male terminals on the final three turns to allow for adjustment.

Table 1 — Feed Point Impedances

160 Meter Tower Frequency (MHz)	R (Ω)	X (Ω)
1.806	15.7	21.5
1.821	15.7	23.3
1.837	15.6	25.0
1.860	15.7	27.6
1.875	15.9	29.2
1.899	15.5	31.9
1.922	16.4	34.5
1.945	16.0	36.9
1.961	15.4	39.1
1.984	16.1	41.8
1.999	16.1	42.3
80 Meter Tower Frequency (MHz)	**R (Ω)**	**X (Ω)**
3.503	120.4	–78.4
3.550	141.1	–81.8
3.596	157.9	–98.4
3.643	165.5	–122.3
3.705	160.2	–158.3
3.751	142.9	–180.1
3.798	119.5	–192.2
3.844	97.9	–196.3
3.906	73.1	–192.1
3.953	59.2	–186.2
3.999	48.3	–178.0

Winding a coil with tubing this large and soft can be difficult to do well. Kinks and bends are easy to make and impossible to remove. I enlisted the help of a friend to carefully feed the tubing out of the coil in which it is supplied. I formed the tubing around the PVC form by hand as he fed me the tubing, rotating the form and pulling the tubing into a close-wound coil. I then stretched the coil to the desired length, separating the turns by about 1⁄8 inch, as you can see in the photo.

The network is built on 1⁄2-inch plywood and uses copper strap and heavy wire to connect the components. Input and output connectors are mounted on a piece of heavy plastic sheet. The coil is strong enough to support itself, but I left it on the PVC form held in place by two sheet-metal screws. Before I installed the network in an enclosure on the tower, I used a DVM with a capacitance function to set the capacitor to 1560 pF. Installed, at 3.55 MHz the SWR at the input was less than 1.1:1! I made some small adjustments of C, both up and down, to verify the network was behaving as it should, then set it back to the original value and left it alone.

Now I have a confession to make. When I first measured the 160 meter tower's base impedance, I confused impedance values between the two frequency markers on the SARK 110 screen. I saw "55 + j 88 Ω" and did not realize this was measured outside the band! My desire for a simple match (a series capacitor of 980 pF would do the job) overpowered caution. Assuming a series capacitor would suffice, I did not have any heavy wire or tubing on hand to wind a second inductor.

As shown in Figure 3, I mounted the second 2000 pF vacuum-variable on a small plastic cutting board made of high-density polyethylene (HDPE). These inexpensive (a set of three boards for $10) and tough kitchen items make great resources for the thrifty RF builder. While the 80 meter capacitor came with its custom mounting bracket, I had to make my own for this second unit. Luckily, two of the body rings fit 2-inch U-bolts very nicely, only requiring that the saddles be supported with an extra set of nuts. The input SO-239 was attached directly to a screw on the capacitor's body ring. A large hose clamp was used to make a connection to the second body ring.

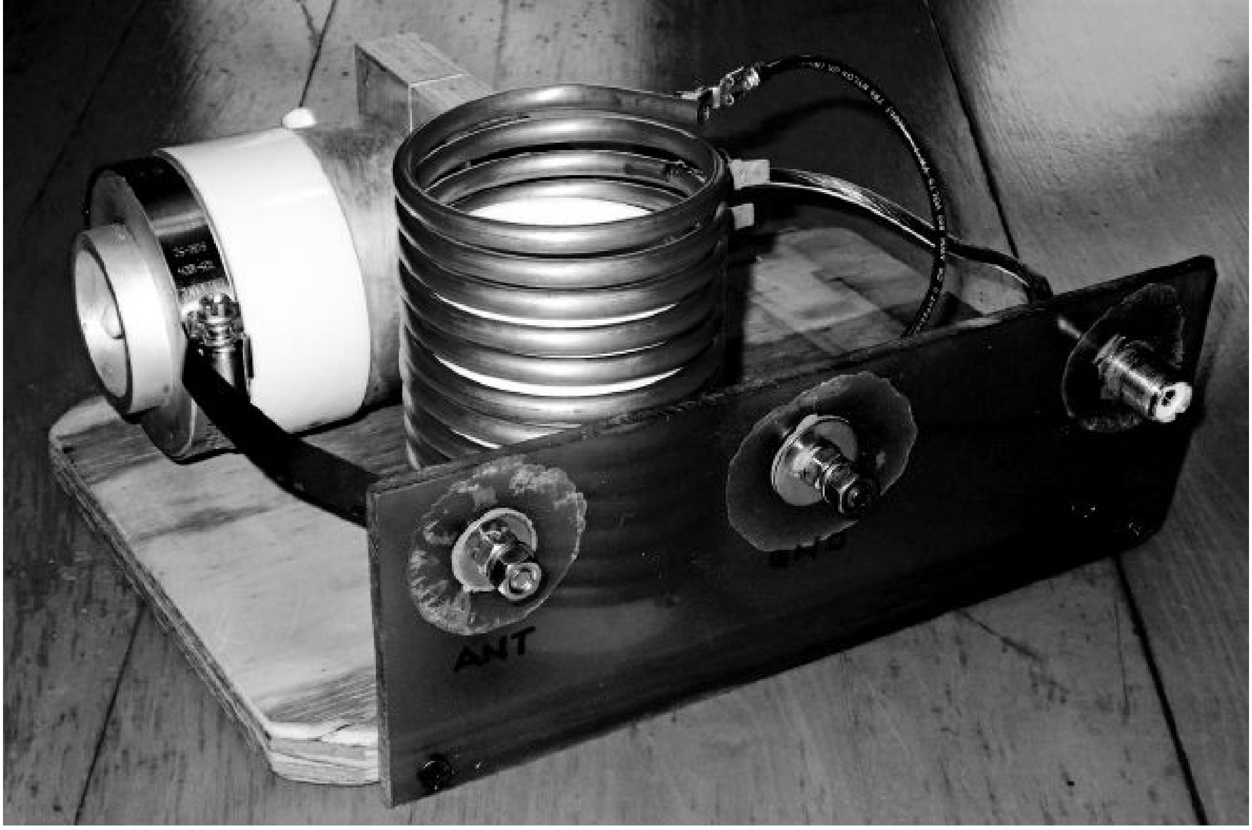

Figure 2 — The completed 80 meter matching network. The components are sturdy for mechanical stability and will easily handle full legal power.

Figure 3 — Mounted on a common plastic kitchen cutting board, the heavy-duty vacuum variable capacitor is connected in series with the load. An improvised 3 μH inductor is attached to convenient points on the capacitor and to the circuit's common ground point.

Needless to say, designed from the wrong value of input impedance, my tuner…didn't! Chastened, I took a second set of data shown in Table 1. Using average impedance values with the online calculators, I determined that I needed a shunt coil of about 3 μH.

Away from my usual supplies, I was faced with another improvisation (remember Experiment #152?) but some #14 AWG stranded THHN wire and an empty soft-drink bottle saved the day. I wound a couple of extra turns on the bottle, attached it between the capacitor and SO-239, and headed back out to the tower. After removing a turn from the coil (I'd wound two extra), adjusting the capacitor gave me an SWR of 1.5:1 at 1.840 MHz! The SWR increases to 1.8:1 at 1800 and stays below 2:1 all the way to 1.915 MHz.

Your Assignment

You might be thinking, “Well, I can do that!” and of course, you can. Do you have a single-band vertical or wire antenna that has an elevated SWR? You can measure its feed point impedance with a snazzy analyzer like an SARK 110 or by making measurements with an MFJ or similar analog instrument. #12 or #14 AWG solid copper wire salvaged from unused ac wiring makes a fine coil, too. If you don’t have an antenna to match, model a design or two and begin with that data. With a little design and workbench time, you’ll soon be singing a happy tune.

Notes

1

The L network was covered in “Hands-On Radio” experiment #21. All previous “Hands-On Radio” columns are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

The calculator is available at **home.sandiego.edu/~ekim/e194rfs01/jwmatcher/matcher2.html** and on other websites.

3

There are a number of design calculators at **www.k7mem.com/Electronic_Notebook/inductors/coildsgn.html**.

Experiment #159 — More L Network Design

My recent column on designing matching networks generated a fair amount of interest from readers interested in solving a particular matching problem or learning more about impedance matching.[1] I thought it would be a good idea to have a practice session and explore a little bit of the Smith chart at the same time.

A Moving Tale

Let's start by having a look at Figure 1, which shows the various possibilities of using L networks to match a load (Z_{LOAD}) to some nominal reference impedance (Z_0).[2, 3] These are simplified Smith charts, with the unshaded half representing the set of impedance values that are matched using the circuit immediately below. Each of the four network configurations shown can be designed to *transform* (match) any impedance in the unshaded region to the impedance at the center of the chart.

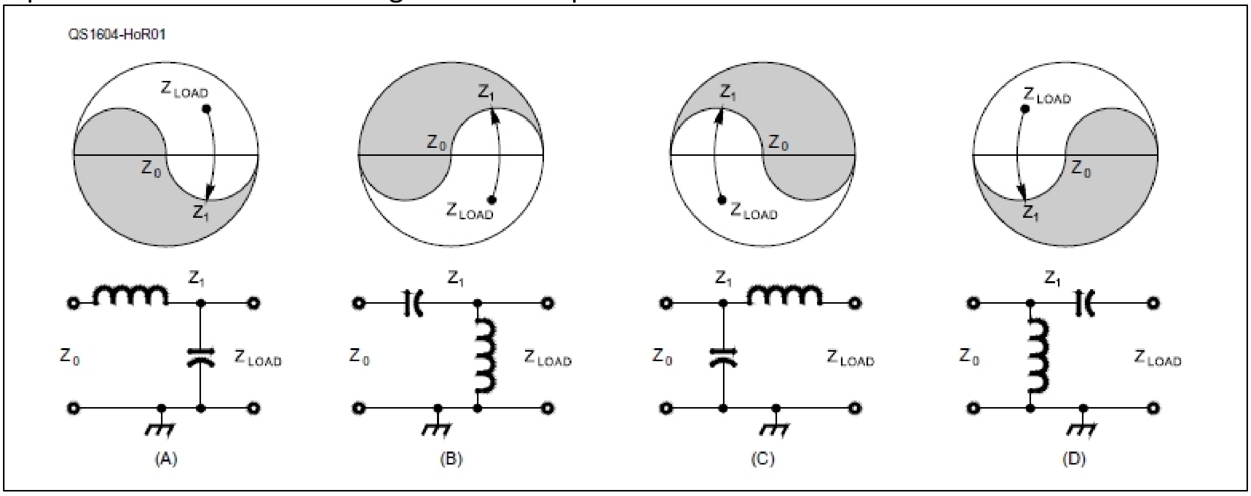

Figure 1 — The four types of L-C L networks and the regions on the Smith chart that represent impedances they can match.

Each particular starting point — the load, Z_{LOAD} — can be transformed into Z_0 in two steps. The first step takes you to a point labeled Z_1 on the border of the matchable region. The second step takes you from Z_1 to the chart's reference impedance, Z_0, which is usually 50 Ω for Amateur Radio systems. Each move is following some kind of curved path. What are those curves?

Figure 2 has two parts. The first part is a standard Smith chart, showing impedance coordinates on the black outer circle and center line, plus a set of red arcs and circles. The second part shows the coordinates of *admittances*, which are the reciprocals of impedances. Admittance coordinates share the outer circle and center line with impedance coordinates but use a mirror-imaged set of blue arcs and circles.

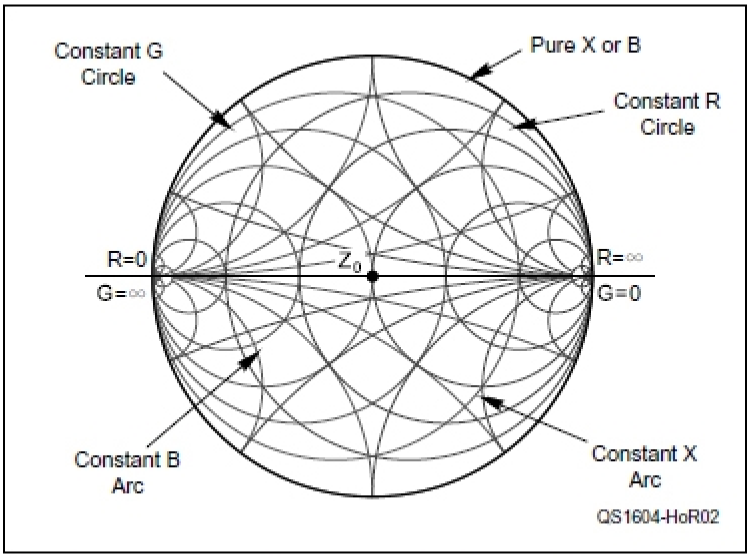

Figure 2 — A combined impedance-admittance Smith chart. Red circles represent constant R and red arcs constant X. Blue circles represent constant G and blue arcs constant B. Chart graphics and printed charts are available with both sets of circles and arcs in finer detail.

Any point on the chart's center line represents a pure resistance or conductance (R + *j*0 or G + *j*0) and any point on the outer circle represents a pure reactance or susceptance (0 + *j*X or 0 + *j*B). Points on one of the red circles all have the same resistance and points on one of the blue circles all have the same conductance. Points on a red arc all have the same reactance and points on a blue arc all have the same susceptance.

[4]

Think of this set of intersecting circles, arcs, and lines as our impedance-matching chessboard.

How does this chessboard "work?" Pieces (impedances or admittances) are moved (transformed) by adding reactances connected in series or parallel (called a *shunt* connection). As shown in the circuits of Figure 1, the moves can only be along the constant resistance (or conductance) circles, since they don't add or subtract resistance (or conductance). For example, if I add some positive inductive reactance in series with a load, the added reactance will "move" the load clockwise along a constant-resistance circle. Similarly, if I add some positive capacitive susceptance in parallel with a load, the point will move clockwise along a constant-conductance circle. Adding reactance or susceptance in series or parallel with load can only "move the load" around the chart on these circles.

If I want to transform Z_{LOAD} to the reference impedance Z_0, then I have to move part of the way along one type of circle until it encounters the complementary circle that intersects the chart's center. Then I move along that new circle to Z_0. Just like a knight in chess or a Manhattan cab. I have to move first in one direction, then switch to another direction.

Figure 3 shows how there are two solutions to every such problem. Start with the point labeled Z_{LOAD}. It has an impedance of approximately 0.5 – *j* 0.1 Ω. (Remember, everything is normalized so that the center point, presumably 50 + *j* 0 Ω, is represented as 1.0 + *j* 0 Ω.) By adding series X_L or X_C, I can move clockwise (X_L) or counterclockwise (X_C) along the red R = 0.5 Ω circle. I need to add enough reactance to reach the blue circle for G = 1.0 S (S stands for siemens, the unit of conductance). Once there, I can move to Z_0 by adding either shunt L or shunt C as noted. One set of "moves" represents what happens in the circuit of Figure 1C and the other in Figure 1D. Depending on the value of Z_{LOAD}, there are always two possible circuits that will get you to Z_0.

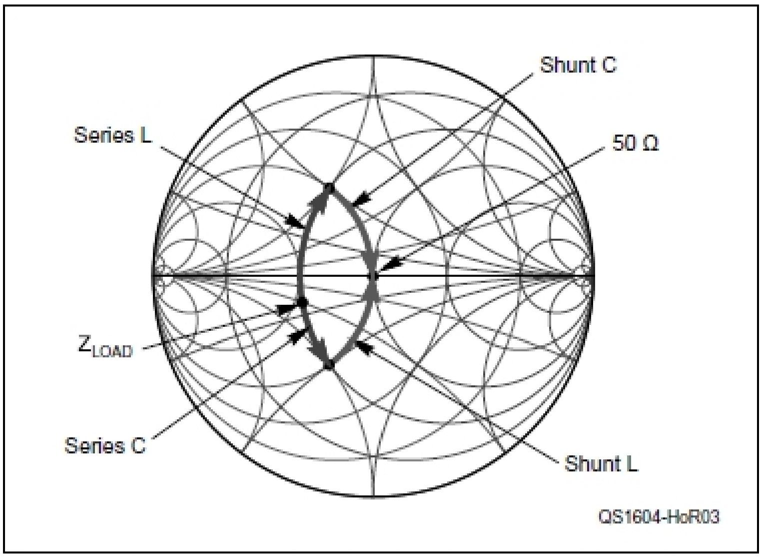

Figure 3 — Any point on the Smith chart can be transformed to Z_0 by using one of two complementary L networks. In this example, series L or C transforms the point to a location on the G = 1.0 circle where shunt (parallel) C or L results in an impedance of Z_0.

Why would I choose one circuit over the other since both choices wind up at the same point on the chart? Unlike the chess knight that must always move in an *L* of 2 squares by 1 square, the *L* network's pair of reactances can take a wide range of values. One of those pairs will have more practical or convenient values (or be cheaper to purchase) than the other.

Working On Your Knight Moves

Okay, your turn! You should now understand why Wetherell's L network calculator can solve an impedance matching problem for only two out of the four possible L-C L networks.

[5]

Those are the two network configurations with Z_{LOAD} in matchable regions of the Smith chart.

Another thing you've just learned: If an L network can't match an impedance — turn it around! If you look at the matchable regions of the two networks that are mirror images of each other, such as in Figure 1A and 1C, you'll see they are complementary. So by turning the network around, the previously unmatchable impedance is guaranteed to be in the matchable region for the new circuit's configuration.

Let's try some example calculations to get you into the game and moving your pieces around. Find the two circuits in Figure 1 that can be used for each example and the component values required. (Answers are at the end of the Notes list.)

Example 1: Z_{LOAD} is the feed point impedance of a 10.1 MHz full-wave loop: 110 + *j* 50 Ω.

Example 2: Z_{LOAD} is the feed point impedance of a Yagi's driven element at 144.2 MHz: 22 – *j* 10 Ω.

Example 3: Z_{LOAD} is the feed point impedance of an off-center-fed dipole at 24.9 MHz: 250 – *j* 150 Ω.

Once you get the hang of it, you can have a lot of fun trying different versions of the matching networks, pushing the impedances to extreme values (high, low, highly reactive), and evaluating which circuit you'd choose to build.

Notes

1

See Experiments #21, "The L Network" and #157 "Matching Network — Design and Build." All previous "Hands-On Radio" columns are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

A complete set of all reactive networks and their matchable regions of the Smith chart are available in the RF Techniques chapter of *The ARRL Handbook*.

3

The ARRL Handbook is available from the ARRL at **www.arrl.org/shop** and from dealers who carry ARRL publications.

4

"Hands-On Radio" Experiments #59 – 61 explain the Smith chart, and there is additional material on the CD-ROM accompanying recent editions of *The ARRL Handbook*.

5

Wetherell's calculator is available at **home.sandiego.edu/~ekim/e194rfs01/jwmatcher/matcher2.html** and on other websites.

Answers to example problems:

Example 1: A and B; A – 1013 nH & 207 pF; B – 2515 nH & 245 pF

Example 2: D and C; D – 48.9 nH and 74.5 pF; C – 38.4 nH and 24.9 pF

Example 3: A and C; A – 770 nH & 34 pF; C – 722 nH & 53 pF

Experiment #166 — Optimizing Placement of Stubs

The idea of making harm

onic filters using stubs — resonant sections of transmission line — is attractive because stubs are frugal, use widely available material, and give pretty good performance. Or do they? I've heard more than one confused stub-builder asking why they only saw a few dB drop in the unwanted signal. They want to know what they're doing wrong. Their stub could be just fine — the problem might stem from where the stub is attached!

Stub Review

Lets back up for a minute and review.

[1]

Stubs used as harmonic filters are usually 1⁄4- or 1⁄2-wavelength long and terminated in a short- or open-circuit. They are connected in parallel with the main feed line by using a T adaptor, assuming coaxial cable is being used. Figure 1 shows the basic connection technique.

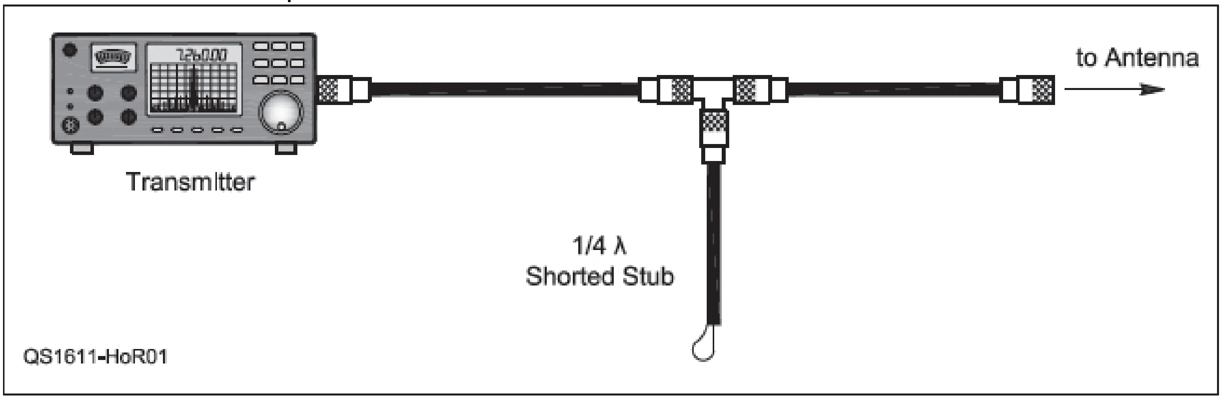

Figure 1 — A harmonic-canceling stub, such as the shorted, ¼-wave stub shown in this figure, will leave a fundamental-frequency signal unaffected while canceling the unwanted 2nd harmonic.

A wave encountering the junction splits, with some of the wave traveling along the stub to the short or open circuit where it is completely reflected back toward the junction. The stub length and the termination are selected so that at the frequency of the undesired harmonic, the wave returning to the junction will be out of phase with and cancel harmonic waves in the main feed line, but leave waves of the fundamental frequency unaffected.

[2]

That simplified view neglects one important thing — *wave impedance* in the feed line. We're not talking about the feed line's characteristic impedance, which is generally 50 or 75 Ω. Wave impedance is the ratio of voltage to current in the feed line. If the feed line is not terminated in its characteristic impedance (i.e. SWR > 1:1), the reflected wave interferes with the forward wave to create standing waves along the feed line. At some points, separated by 1⁄2 wavelength, the voltage to current ratio (wave impedance) will be a maximum. Just 1⁄4 wavelength away, the wave impedance will be a minimum.

You can see this change by using a Smith chart that shows the wave impedance in a feed line. From the point representing the termination impedance (Z_{LOAD} in Figure 2), as you move along the line away from the termination ("Toward the Generator" on the chart) each new point on the chart shows the different wave impedance at that location in the feed line. It is important to remember that the electrical distance in wavelengths will be different for the fundamental and for each harmonic!

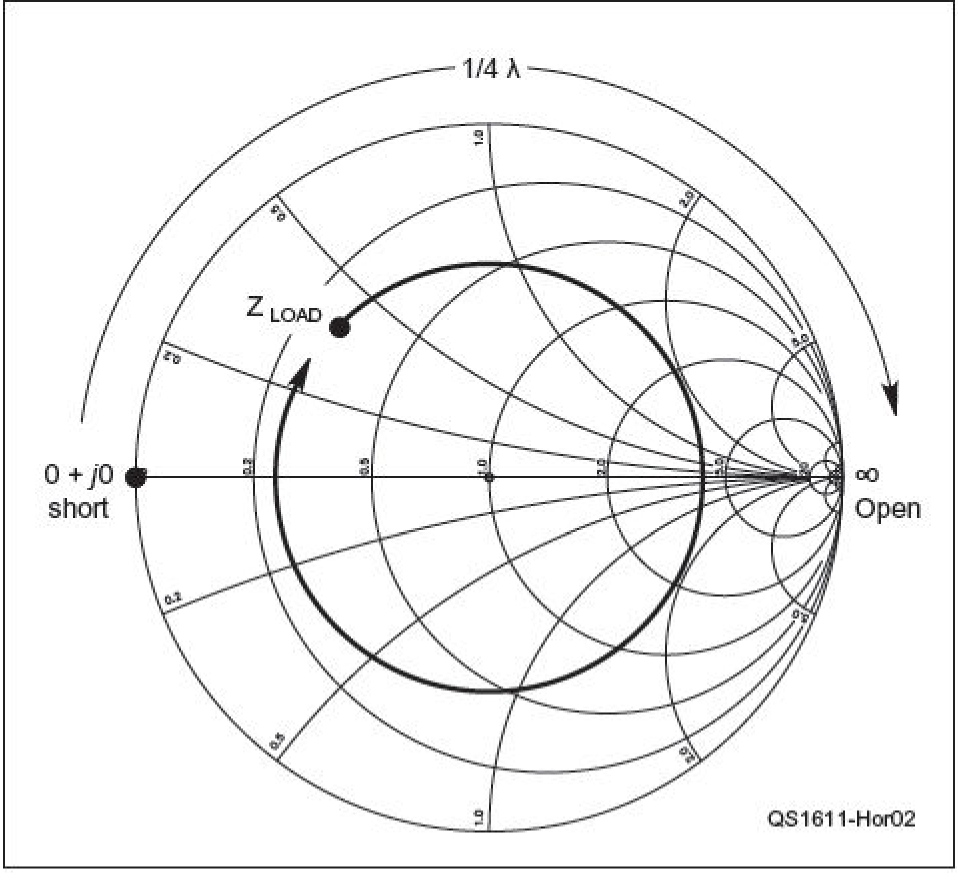

Figure 2 — The Smith chart shows wave impedance at different points in a feed line. Starting from a termination such as Z_{LOAD}, moving away from the termination also moves the point representing wave impedance clockwise around the chart. Each ¼-wavelength of feed line moves halfway around the chart.

If a feed line is terminated by a short circuit (0 + *j*0 Ω at the left side of chart), moving toward the junction increases the wave impedance along the way. After 1⁄4 wavelength (halfway around the chart) the wave impedance reaches infinity. If the stub is long enough, wave impedance will return to zero 1⁄2 wavelength from the termination.

Note that if the shorted stub is 1⁄4-wavelength long at the fundamental — presenting a high impedance at the junction — it will be 1⁄2-wavelength long at the harmonic and present a low impedance at the junction. The result is to short out the harmonic while leaving the fundamental unaffected. Neat trick, huh? Table 1 shows a variety of 1⁄4- and 1⁄2-wavelength stubs that can be used for a number of useful harmonic-canceling jobs. (See W2VJN's book for a lot more!)

Table 1
Common Harmonic-Canceling Stubs

Stub Type	Passes	Cancels
λ/4 160 meters shorted	160	80, 40, 20, 15, 10
λ/4 80 meters shorted	80	40, 20, 15, 10
λ/4 80 meters open	40, 20	80
λ/4 40 meters shorted	40, 15	20, 10
λ/4 40 meters open	20, 10	40, 15
λ/4 20 meters shorted	20	10
λ/4 20 meters open	10	20

Why Impedance Matters

There is one more thing to consider. The wave impedance in the main feed line determines how much effect the stub's low impedance will have at their junction. If the main feed line's wave impedance at the junction is high, the stub's low impedance will effectively "short out" the harmonic, reducing its amplitude quite a bit. If the main feed line's wave impedance at the junction is low, however, the stub's low impedance placed across it will have comparatively little effect.

For a monoband antenna like a single-band Yagi or ground plane matched to 50 Ω, the wave impedance along a 50 Ω feed line will be relatively constant *at the fundamental*. However, as anyone knows who has tried to transmit on 10 meters into a 20 meter monoband antenna, there is a sizable mismatch at the harmonic. This means the wave impedance for the harmonic will vary dramatically at different locations along the line and so will the effectiveness of a harmonic-canceling stub.

How big is the effect of stub placement? Both W2VJN and K9YC have modeled the harmonic attenuation of a stub at

different locations in a feed line.
3
,
4

For the typical example of trying to cancel the 10 meter harmonic of a 20 meter signal, W2VJN's simulation found that harmonic suppression could range from a worst-case low of 6.2 dB to a best-case maximum of 51.1 dB! (George's model also simulated the effects of a transmitter's output circuit, which is usually mistuned at the harmonic frequency.) So it's quite realistic to find the performance of a filter stub very hit-or-miss just because of the varying wave impedance at different points along the feed line.

Finding the Right Place

How *do* you find the right place to locate a stub — or avoid the wrong place? The best way to start is by measuring the impedance of the antenna connected to the feed line — at the frequency of the harmonic you are trying suppress. Then find impedance along the feed line (again, at the harmonic) by using a modeling program like *TLW* (which comes with the *ARRL Antenna Book*), or *TLDetails* and *ZPlots* from AC6LA (**ac6la.com**), or *SimSmith* from AE6TY (**www.ae6ty.com/smith_charts.html**).

What if you can't measure or model the antenna system? Or you can't install the stub where a model says you should? The most common location for a harmonic filter stub is at the output of an amplifier or transmitter. The impedance at this point is strongly influenced by the output impedance of the transmitter so you can make a good guess about whether this is a high- or low-impedance point based on the output circuit.

If the final component (closest to the output connector, not including any RF chokes) of the transmitter's filter or tuning circuit is a series inductor (as for a Pi-L network), the impedance at the harmonic will be high. (See **k9yc.com/Coax-Stubs.pdf** for a table listing the output circuits of many common amplifiers as well as more applications of stubs.) Attach the stub at the amplifier output or some multiple of 1⁄2-wavelengths away (at the harmonic). If the output component is a shunt capacitor (as for a Pi network), the impedance will be low so there should be 1⁄4-wavelength (or some odd multiple) of feed line between the transmitter output and the stub to place the junction at a high impedance point.

Another option is to force a high-impedance point by using a *double stub* as in Figure 3. Place the initial stub wherever you can. Then add 1⁄4 wavelength of cable (again, at the harmonic) to the main feed line and attach a second harmonic-canceling stub. The first stub creates a low-impedance point in the system and a high-impedance point 1⁄4-wavelength away. The second stub can take full advantage of the high-impedance point and provide an additional 30 dB or more of suppression.

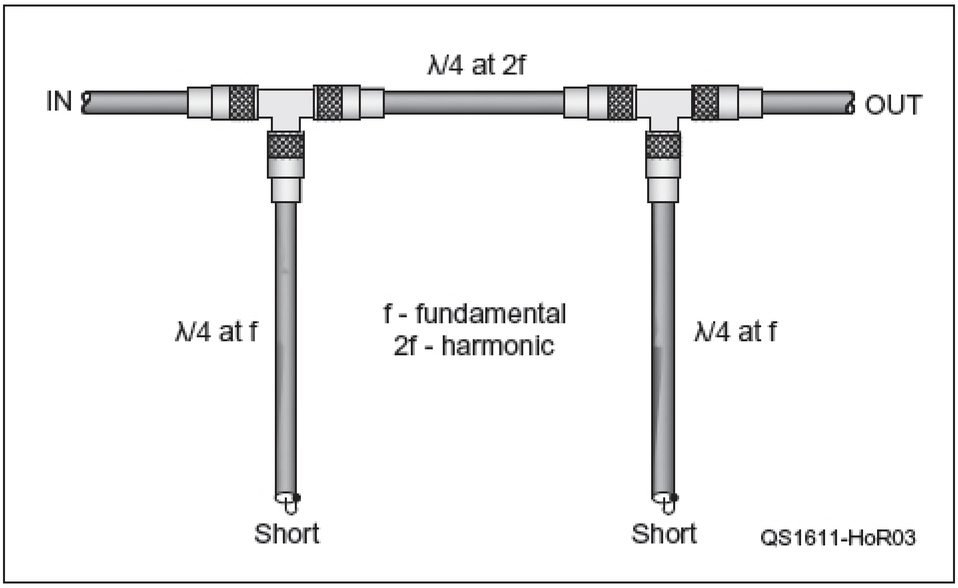

Figure 3 — Even if a stub is not at the optimum, high-impedance location in the feed line, it will still create a low-impedance point in the line with a high-impedance point ¼-wavelength away. A second stub at this new high-impedance point will have its full harmonic-canceling effect.

The key to getting the most out of your stubs is to measure their effectiveness. Don't expect one stub placed randomly in your antenna system to solve your harmonic problems completely. Study how the stubs work, learn to make the necessary measurements, and use modeling tools like those listed here to design and install them. You will be glad you did!

Notes

[1] How transmission line stubs work and how to design the basic types were covered in "Hands-On Radio" experiments 22, 57, and 58. Experiment 81, on synchronous transformers, covers related ideas. All previous "Hands-On Radio" experiments are available to ARRL members at **arrl.org/hands-on-radio**.

[2] There are many variations on this basic scheme. See the *ARRL Antenna Book* chapter on Transmission Lines (**www.arrl.org/antenna-book**) for more information. Even more detail is available in W2VJN's *Managing Interstation Interference*, available from International Radio (**www.inrad.net**).

[3] J. Brown, K9YC, "Optimizing the Placement of Stubs for Harmonic Suppression," *National Contest Journal*, July/August 2015, pp 8 – 11.

[4] G. Cutsogeorge, W2VJN, "Optimizing the Performance of Harmonic Attenuation Stubs," *National Contest Journal*, Jan/Feb 2015, pp 3 – 4.

Experiment #177 — Feed Line Comparison

"Feed Lines, Decibels, and Dollars" was the title of Steve Ford's, WB8IMY, article in the August 2017 issue of *QST* on choosing a feed line — but you have to have the data! His column reminded me of some important graphs and tables in *The ARRL Handbook* and *Antenna Book* that can help you make those decisions.
[1]

I don't know when you last purchased some new coax, but it's sold by the foot and can cost an arm and a leg! We have limited resources to build our stations, so how do we balance feed line dollars against radio, antenna, computer, and gadget dollars? It's not enough to say, "Just buy the best you can afford," and hope you didn't overspend.

Matched Loss and Non-Matched Loss

Let's start by revisiting the term *matched loss*. This is the loss when the feed line is terminated in its characteristic impedance and standing-wave ratio (SWR) is 1:1, such as 50 Ω cable attached to a 50 Ω load. Matched loss is specified in dB/100 feet by most North American manufacturers, and you can get it from their websites or from tables in the ARRL reference books.

In the real world of antennas, SWR is almost always greater than 1:1. That mismatch causes some of the power to be reflected back and forth in the line until it is either radiated as a signal or turned into heat. Figuring out this extra *mismatch loss* can be a chore, but it has been simplified by Joseph Reisert, W1JR, in Figure 1.
[2]

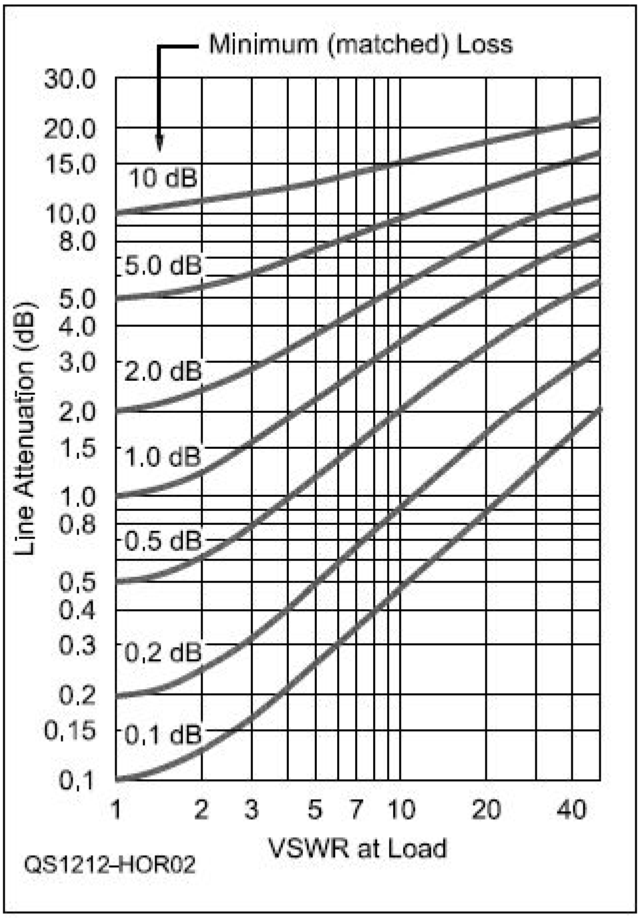

Figure 1 — Total loss in a feed line with a mismatched load. Start on the horizontal axis with the SWR at the load (not at the transmitter). Go upward until you come to the curve representing the line's matched loss. Then proceed horizontally to the left to find the total loss. [Graphic courtesy of Joe Reisert, W1JR]

Start with the matched loss for your length of feed line. Let's say the feed line has 2.0 dB of matched loss and the SWR at the antenna is 3:1. From 3:1 on the *x*-axis, go vertically to the 2.0 dB curve. Then turn left and intersect the *y*-axis at 2.8 dB. The 3:1 SWR "costs" another 2.8 – 2.0 = 0.8 dB of loss. This scenario isn't very dramatic, but you can see from the graph that higher SWR and matched loss can eat up a lot of your transmitter's output pretty quickly.

How Much You Can Tolerate

Let's use the graph in a different way to answer the question, "What is the feed line with the highest matched loss I can accept?" This is a very common question when you are designing an antenna system. Perhaps you set a goal of not having more than 6 dB (one S-unit) of feed line loss. If you know the maximum SWR of your antenna, you have enough information to find the answer. Let's say you have a non-resonant doublet that has one difficult band for which the SWR is 5:1.

Start on the *y*-axis of Figure 1 at 6 dB and follow that horizontal line across the chart. Now go to the *x*-axis and the vertical line for an SWR of 5:1. Where those two lines cross, interpolate between the curves to get a loss of about 4 dB (the curves are spaced logarithmically, so you have to take that into account). That's how much matched loss you can tolerate.

Note that for a given maximum amount of allowable loss, as SWR increases the maximum matched loss goes *down*. That's a little counter-intuitive until you realize that for higher SWR, you need cable that is less lossy to keep total loss below your maximum loss.

There is one more missing piece of the puzzle, and that is the length of the line between your transmitter and the antenna. Let's say your feed line is 400 feet long — not too unusual for a Field Day or expedition-type setup. You need cable with a matched loss of 4 dB (or less) at the frequency of your "problem" band — let's say that's 15 meters, or 21.2 MHz.

You can start browsing through websites, but you'll find that the loss data is typically only specified at 1, 10, 100, and 1,000 MHz. This makes the process harder than it needs to be, so Table 1 was developed by W3LPL using a feed line loss calculator created by VK1OD.
[3]

It shows the "feet per dB" for a wide variety of coaxial cables. The lossier the cable (toward the bottom of the table), the fewer feet required to incur 1 dB of loss.

Table 1
Cable Attenuation (feet per dB)

MHz	1.8	3.6	7.1	14.2	21.2	28.4	50.1	144	440	1,296
LDF7-50A	3,333	2,500	1,666	1,250	1,000	833	625	370	200	110
FHJ-7	2,775	2,080	1,390	1,040	833	667	520	310	165	92
LDF5-50A	2,108	1,490	1,064	750	611	526	393	227	125	69
FXA78-50J	1,666	1,250	769	588	435	370	256	130	71	36
¾" CATV	1,666	1,250	769	588	435	385	275	161	59	33
LDF4-50A	1,145	809	579	409	333	287	215	125	70	39
RG-17	1,000	769	556	370	294	250	200	77	40	20
LMR-600	973	688	492	347	283	244	182	106	59	33
SLA12-50J	909	667	500	355	285	235	175	100	53	34
FXA12-50J	834	625	455	300	250	210	150	83	48	25
FXA38-50J	625	435	320	220	190	155	115	67	37	20
9913	625	435	320	220	190	155	110	62	37	20
LMR-400	613	436	310	219	179	154	115	67	38	21
RG-213	397	279	197	137	111	95	69	38	19	9
RG-8X	257	181	128	90	74	63	47	27	14	8
RG-58	179	122	83	59	50	42	30	18	9	5
RG-174	91	67	48	32	26	23	17	10	5	3

In the case of our 15-meter problem, find the column labeled "21.2." Our matched loss budget is 4 dB and the feed line length is 400 feet, so we can accept any cable with more than 400 / 4 = 100 feet/dB of loss. If we begin at the bottom, we see that RG-174, RG-58, and RG-8X are too lossy — it doesn't take enough cable before 1 dB of loss at 21.2 MHz has been created. RG-213 shows a loss of 111 feet per dB, so a 400-foot length will result in 400 / 111 = 3.6 dB. You could spend the extra money on a cable with less loss, but you don't have to!

How much RG-213 could you use and still make that 4.0 dB loss budget? You could use 111 feet/dB × 4 dB = 444 feet. If you wanted to make things a little simpler and just use 500-foot spools without having to divide the cable, select a type for which it takes *more* than 500 / 4 = 125 feet to create 1 dB of loss. Continuing our journey up the list, LMR-400 is the first cable to make the cut, so to speak.

This is a very handy table. If you have a copy of the *Antenna Book* (23rd edition), take a look at Chapter 23's section on "Choosing and Installing Feed Lines." There you'll find two other tables that show how long a cable would have to be before you would gain 1 dB by replacing it with 7⁄8-inch or 1⁄2-inch Heliax hardline.

Does It Really Matter?

Maybe not so much at 160 meters, but as Steve Ford's article points out, it certainly can make a difference, particularly above HF. VHF/UHF and microwave operations are incredibly sensitive to feed line loss, because the cable losses increase quickly with frequency. At 1,296 MHz, a popular EME frequency, it doesn't take much of even the best flexible cable before replacing it with hardline begins to pay off. And signals converted to heat can *never* be recovered, not even with the best antennas or preamps. Sometimes, that expensive cable is the best bargain around.

Coaxial Cable Types

Hams use the familiar
RG- numbers for coaxial cable like RG-8, RG-58, and RG-213. However, these designators no longer mean that the cable meets any particular specification. It is simply a matter of convenience that vendors still use them and there are wide variations in loss, velocity factor, and quality between cables with the same RG- designator. For example, the Mouser Electronics website (
www.mouser.com
) lists 72 different cables as "RG-58." Pay less attention to the RG-type designator and more to the manufacturer's part number, such as Belden 8267 or Alpha 9213, which are cables meeting the usual specifications for what we call "RG-213" coax. Look for the manufacturer and part number printed on the cable. If it's not there, you can't be sure what

you’re getting.

Notes

1 The *ARRL Handbook* (2017 edition, Item 0628) and *Antenna Book* (23rd edition, Item 0444) are available from the ARRL Bookstore at **www.arrl.org/shop**.

2 J. Reisert, W1JR, “VHF/UHF World,” *Ham Radio*, Oct. 1987, pp. 27 – 38.

3 *Antenna Book*, 23rd edition, Table 23.4.

Electronic Fundamentals

Experiment #127 — Phasors, Part 2

No, not the kind of phaser you set on stun, silly! If you passed your General exam, you learned about phase and a little bit about angular frequency. Amateur Extra class licensees (and those studying for the Extra) have even used phasor notation, although it was called by another name.

In this two-part column, we'll start by developing basic concepts to show what a phasor is and how it relates to things you already understand. Then we'll progress to examples of using phasors to describe electrical and radio phenomena, such as modulation. As ham radio begins to use more advanced types of modulation, understanding phasors will provide an important bridge between the familiar AM/SSB modulation and the future.

The Sinusoid

Like many meals in radio, this dish begins with a sine wave and is seasoned with complex numbers. (You can find these subjects in the math tutorials for the General and Extra exams listed on the ARRL website.
[1]
) The sine wave (or *sinusoid*) looks like a regularly increasing and decreasing wave but as Figure 1 shows, it is really related to rotation.

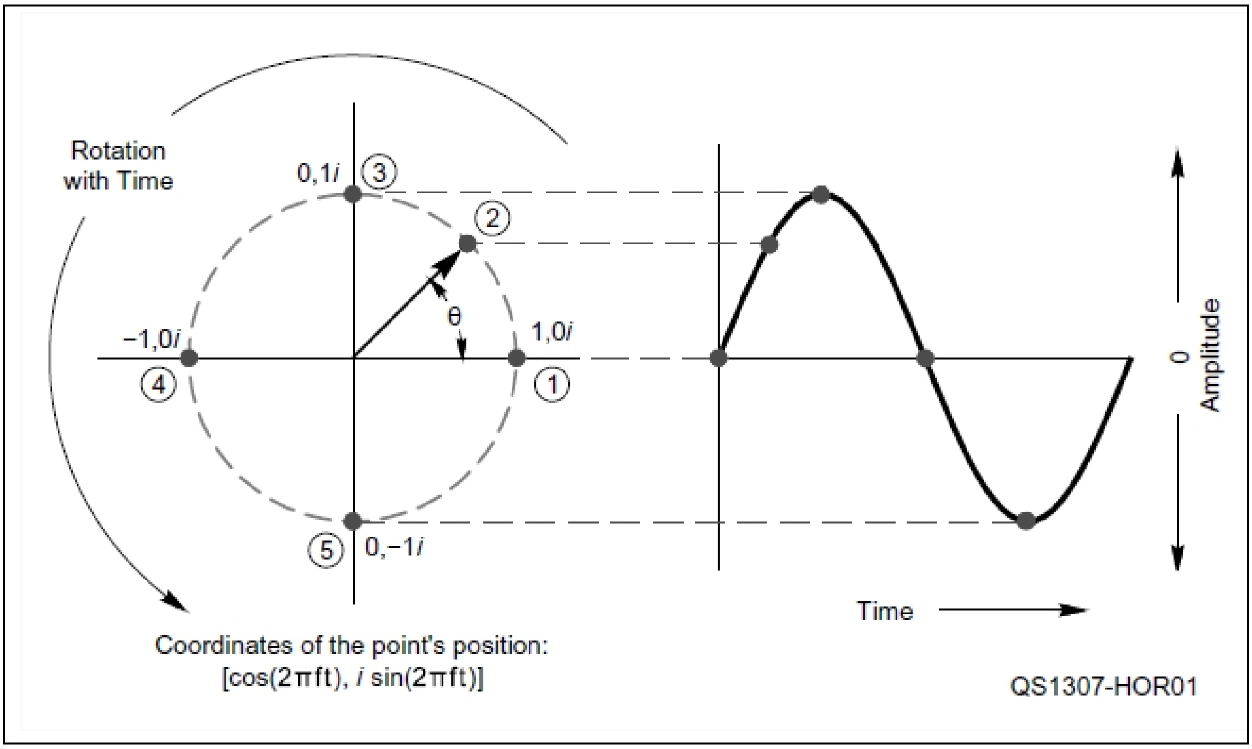

Figure 1 — The sine wave at right shows the y-axis coordinates of a point revolving around the origin, shown at left.

Imagine a point revolving counter-clockwise in a circle around the origin of the complex plane as at the left of Figure 1. If the point is one unit from the origin the coordinates for each location visited by the point are [cosθ, *i* sinθ], where θ (the Greek letter theta) is the angle from the positive x axis to the line from origin to the point. This circle, not surprisingly, is called the *unit circle* because the value of its radius is 1 or unity.

As the point revolves around the origin, θ steadily increases from 0 to 360° and begins again at 0° with each cycle. (Counterclock-wise is considered the positive direction.) If the point always moves at the same speed, the frequency of the cycles around the origin, f, does not change and the point moves 360 × f degrees every second. That means the number of degrees a point has moved in t seconds, $\theta = 360 \times f \times t$.

There are 2π radians (another unit of angular measurement) in a circle so $\theta = 2\pi \times f \times t$. The quantity 2πf is referred to as angular frequency, ω, and you see it used in the formulas for reactance and many other electrical calculations that depend on frequency.

Tying it all back together, the coordinates for the position of the point for every point in time as it moves around the circle are [cos(2πft), *i* sin(2πft)] and the graph at the right of Figure 1 plots the point's y (or imaginary) coordinate versus time, creating a sine wave. If we plotted the point's x (or real) coordinate versus time, it would create a cosine wave.

Making the leap from a point moving around the circle to something more electrical, the magnitude of the point's imaginary coordinate, sin(2πft) = sin(wt), can also be a voltage or current or field strength. In fact, the familiar sine wave of ac power comes from the rotational motion of a generator's field coil. As the coil passes through the magnetic field in the generator, the angle between the coil and the field changes in the same way as our point moves around the origin. This changing relationship between the coil and the field creates a sinusoidal voltage in the coil.

While both sine and cosine waves are generally referred to as sinusoids, they differ from each other in an important way. Starting from t = 0, the cosine wave starts at a value of 1 + 0*i* and the sine wave at 0 + *i*. Other than starting at different values, the waves are identical. The cosine wave describes the real coordinate and the sine wave describes the imaginary coordinate.

You can also think of that difference in starting value as a difference in angle, in which the sine wave is $\pi/2$ radians (90°) ahead of the cosine wave. This difference never changes because both waves are describing the same thing — constant rotation. The position of a particular point on the wave is its phase and the amount of the difference is called the *phase angle*, 90° in this case.

This is where the following trigonometric identities come from: $\sin\theta = \cos(\theta - 90°) = \cos(\theta - \pi/2)$ and $\cos\theta = \sin(\theta + 90°) = \sin(\theta + \pi/2)$. Many, many more such relationships between sine and cosine waves become obvious (or at least more understandable) when viewed from the standpoint of rotation and the unit circle.

Polar Notation

So far we have used the *rectangular* form for the coordinates of our moving point: x + *i* y. In most engineering technical literature, the letter *j* is used instead of *i* to avoid confusion with current and from here on in the column, we'll do so, as well.

Next is the form that you may have already learned (or will learn!) for your Extra class exam and that is the polar form in which the coordinates take the form of a radius and an angle: $r \angle\theta$. Polar form is read "r at (an angle of) theta." Using polar form coordinates for a point on the unit circle is easy because they're always $1\angle\theta$. If you are describing the point's position as it whirls around the circle, you can use the equation for angular frequency we figured out earlier and the coordinates become $1\angle(2\pi ft) = 1\angle(\omega t)$. So this particular method is a good shorthand way of describing what the moving point is doing.

Introducing the Phasor

When dealing with RF signals and circuitry, it's often true that the frequency of the signals doesn't change. Think of an RC low-pass filter, for example: the input signal $V_{IN} \sin(\omega t + 0)$ and output signal $V_{OUT} \sin(\omega t + \varphi)$ have the same frequency, even though their amplitudes are different by the ratio of V_{OUT}/V_{IN} and they are offset in phase by φ.

Assuming the same frequency for both signals, our polar form can now be simplified to $V\angle\varphi$ where φ is just the phase angle between a signal and some reference signal or phase. The input signal to a circuit is usually the reference for measuring phase differences.

Hey, guess what? $V\angle\varphi$ is a phasor! A phasor is just a complex number that represents the amplitude and phase of a sinusoid and the $V\angle\theta$ polar notation is just convenient mathematical shorthand. Phasors are a type of vector – quantities that have both a magnitude and a direction. In the case of $V\angle\varphi$, the magnitude is $|V|$ and the direction is the phase angle, φ, so the more cumbersome name "phase vector" was shortened to "phasor." (As a vector, a phasor is often shortened even further and written as a single bold letter, such as **V** or **I**.)

Imagining a Wheel

If you are having trouble visualizing how the vigorous circular motion of a point translates into the smooth undulation of a sine wave, Figure 2 illustrates a demonstration that may help: Take the front wheel off of a bicycle and tape a battery-powered light to it so you can see the light from the side of the wheel. In a large room, have a friend hold the wheel, turn on the light, and give the wheel a gentle spin — say about one revolution per second. Turn off the room lights and have the friend walk slowly and steadily across the room with the wheel edge-on to you about 10 or 20 feet away. The up-and-down motion of the light stretched out across the room traces out a sinusoid! (Note to license exam instructors — this is also a good way to demonstrate the relationship between frequency and wavelength by changing the wheel's spin rate as you walk across the room to represent a signal traveling at the speed of light.)

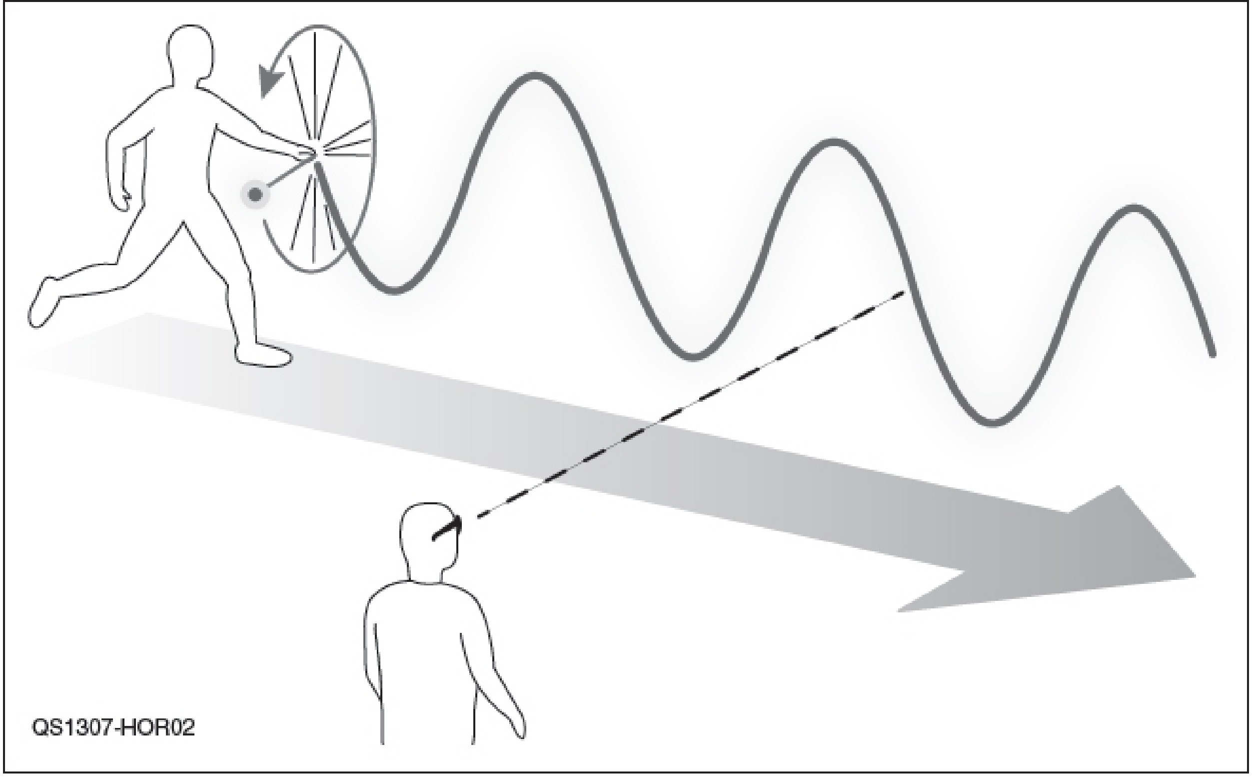

Figure 2 — Mounting a light on a spinning wheel and carrying it across a darkened room edge-on to the viewers is an effective way of demonstrating the link between rotation and sinusoids.

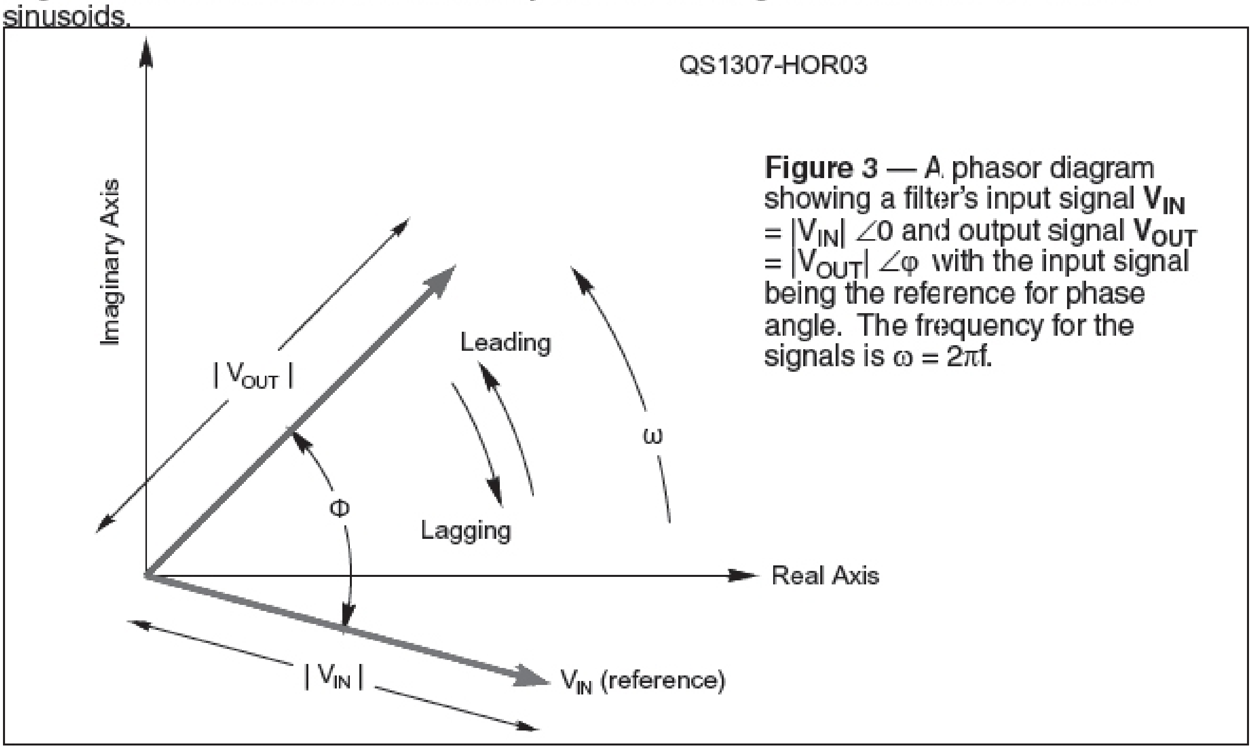

Figure 3 — A phasor diagram showing a filter's input signal $\mathbf{V_{IN}} = |V_{IN}| \angle 0$ and output signal $\mathbf{V_{OUT}} = |V_{OUT}| \angle \varphi$ with the input signal being the reference for phase angle. The frequency for the signals is $\omega = 2\pi f$.

If our original sine wave is the reference signal, the phasor describing the sine wave is $\mathbf{V}\angle 0$ and the cosine wave is $\mathbf{V}\angle -90°$ or $\mathbf{V}\angle -\pi/2$. Remember, the frequency is assumed to be the same for both signals, whether 60 Hz from the power grid or 14.200 MHz on 20 meters. Figure 3 shows a *phasor diagram* for the signals at the input and output of a filter.

There is a final way to describe the signal — the *exponential form* in which it is represented as $V\ e^{j\theta}$. This form comes from the mathematics behind Euler's equation [2] in which the coordinates of our point are miraculously shown to be equivalent to $e^{j\theta} = \cos\theta + j\sin\theta$. The serious and beautiful math [3] behind this equation lies at the heart of much of electrical engineering and leads to the jaw-dropping Euler's identity: $e^{j\pi} = -1$ which unites the two most widely used transcendental numbers (e and π), imaginary numbers (*j*), negation and unity. Not bad for a point moving in a simple circle, huh?

Notes

[1] www.arrl.org/studying-for-the-general-license, click on "Math Tutorials," then "Tutorials on Math for License Exams"

[2] Nahin, Paul J., *Dr. Euler's Fabulous Formula: Cures Many Mathematical Ills* (Princeton University Press, 2006).

[3] In his *Lectures on Physics*, physicist Richard Feynman characterized the equation as "our jewel" and "one of the most remarkable, almost astounding, formulas in all of mathematics."

Experiment #126 — Phasors, Part 1

Last month, we introduced the phasor — a way of representing a sinusoid in terms of its amplitude and some value of phase compared to a reference. Phasor notation looks like $V\angle\phi$ where V is the amplitude and ϕ is the phase. Let's learn a few more things about phasors.

Basic Phasor Math

One of the nice things about phasors is that multiplying them is pretty easy. Multiplying phasor A by phasor B requires you to multiply the magnitudes and add the angles:

$$V_A\angle\phi_A \times V_B\angle\phi_B = V_AV_B\angle(\phi_A + \phi_B)$$

Similarly, to divide phasors, divide the magnitudes and subtract one angle from the other.

$$V_A\angle\phi_A \div V_B\angle\phi_B = (V_A / V_B)\angle(\phi_A - \phi_B)$$

Remember that to use phasor notation this way requires both signals to have exactly the same frequency so that ϕ_A and ϕ_B are constant. If that isn't true, the math gets a lot fancier.

How about adding phasors? Not quite as easy. Because we are operating in polar notation, you must break down each phasor into its X axis and Y axis components, add those components together and then change them back to phasors:

$$V_C\angle\phi_C = V_A\angle\phi_A + V_B\angle\phi_B$$

X axis component = $[V_A \cos \phi_A + V_B \cos \phi_B]$

Y axis component = $[V_A \sin \phi_A + V_B \sin \phi_B]$

$V_C\angle\phi_C = X + j\,Y = \sqrt{(X^2 + Y^2)} \angle (\tan^{-1} Y/X)$

Bleh!

Fortunately, scientific calculators and software usually have routines to do this math automatically — look in the manual or Help file under *polar notation*. (Remember that online tutorials for this kind of math are listed on the ARRL website.[1])

Graphically Adding and Subtracting

We have been drawing all of the phasors with their head at the origin and their tail (where the arrowhead is) at the point representing the magnitude and angle. Phasors can be drawn anywhere on the X-Y plane, though, as long as they have the same magnitude and angle! This makes adding them together graphically very simple, as shown in Figure 1A, by arranging the phasors "head to tail." The resulting phasor is drawn from the head of the first phasor to the tail of the last phasor.

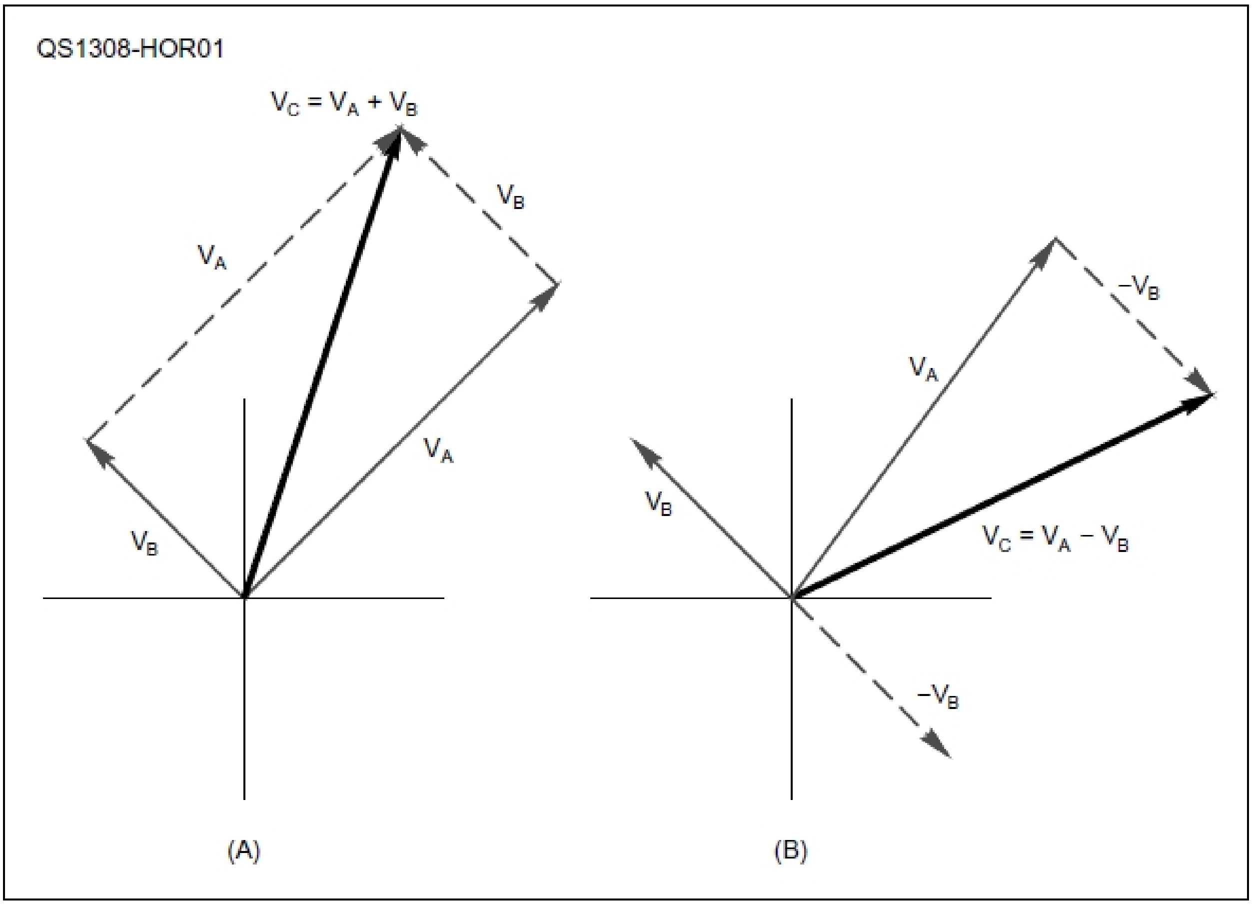

Figure 1 — Adding and subtracting phasors.

Just like ordinary numbers, you can add phasors together in any order. What about subtraction? Turn the phasor to be subtracted 180° and add as in Figure 1B — just like subtracting an ordinary number by multiplying it by -1 and adding instead. Now you know how to add, subtract, multiply, and divide phasors all having a common frequency.

Let's learn another neat trick — if the phasors represent voltages, how do you find the difference in voltage between two phasors? When you measure voltage at a point in a circuit, you measure voltage "from" ground "to" the point. In effect you are measuring the voltage at the point and then subtracting the voltage at your ground reference, which is zero. If our phasor ground reference is at the origin as in Figure 2, the tail of the phasor (with the arrowhead) shows the voltage measurement with respect to ground.

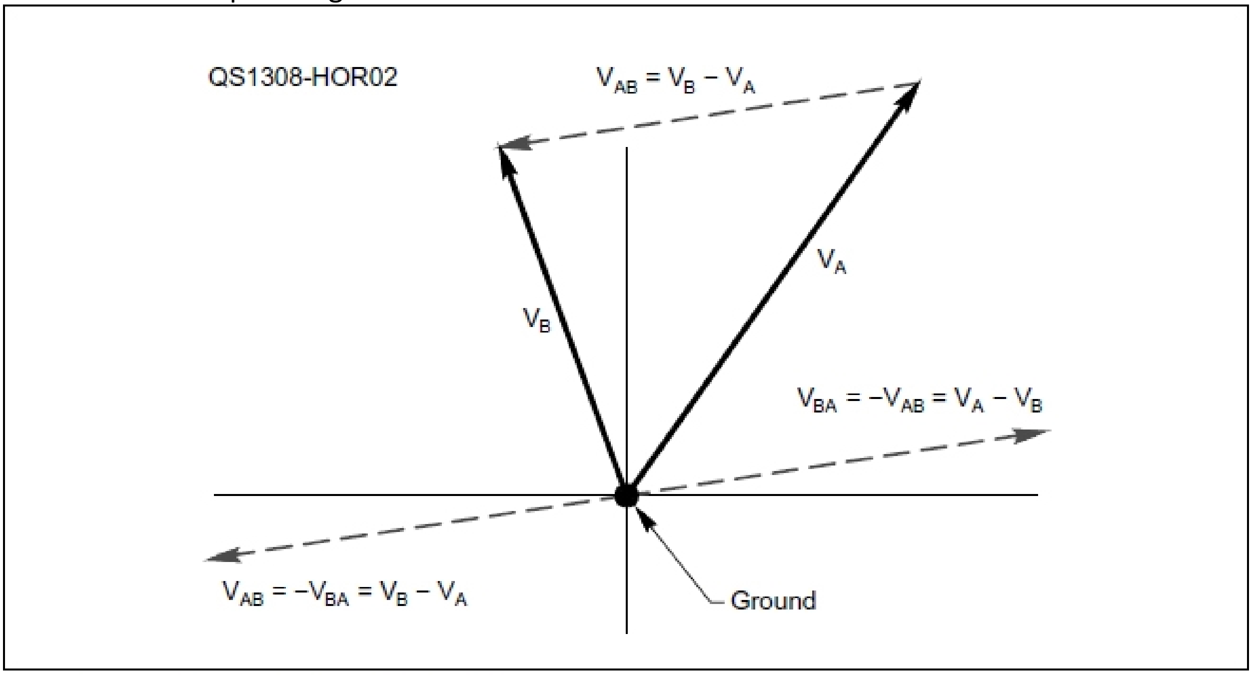

Figure 2 — Phasor-to-phasor voltages.

When you measure voltage between two ungrounded points in a circuit, your meter's negative probe is the reference and you measure voltage "from" the reference point "to" the point where the positive probe is. Figure 2 shows how this works if the two voltages are phasors and our reference "ground" point is at the origin. The voltage "from" phasor A "to" phasor B is itself a phasor, written V_{AB} and calculated as $V_B - V_A$. We could also measure the voltage from phasor B to phasor A as $V_{BA} = V_A - V_B$. You can see that V_{BA} has exactly the same magnitude but the opposite angle to V_{AB}. Take a minute and sketch out the subtraction of the phasors to make sure you see how I came up with V_{AB} and V_{BA}.

Phasor-to-Phasor Voltages

This is all fine and dandy, but does it have any practical value? Would you ever encounter phasor-to-phasor voltages? Yes and closer to home than you imagined. Residential ac power electrical service supplies two phases to the main breaker box, each 120 V. The power comes from a transformer at the utility pole with a single primary winding and two secondary windings. Figure 3 shows the secondary windings each supplying one phase of your electrical service and connected together at one end as the neutral. The polarity of the windings is opposite so that the phasors representing their voltages point in opposite directions as shown in the phasor diagram. This is called *split-phase* power.

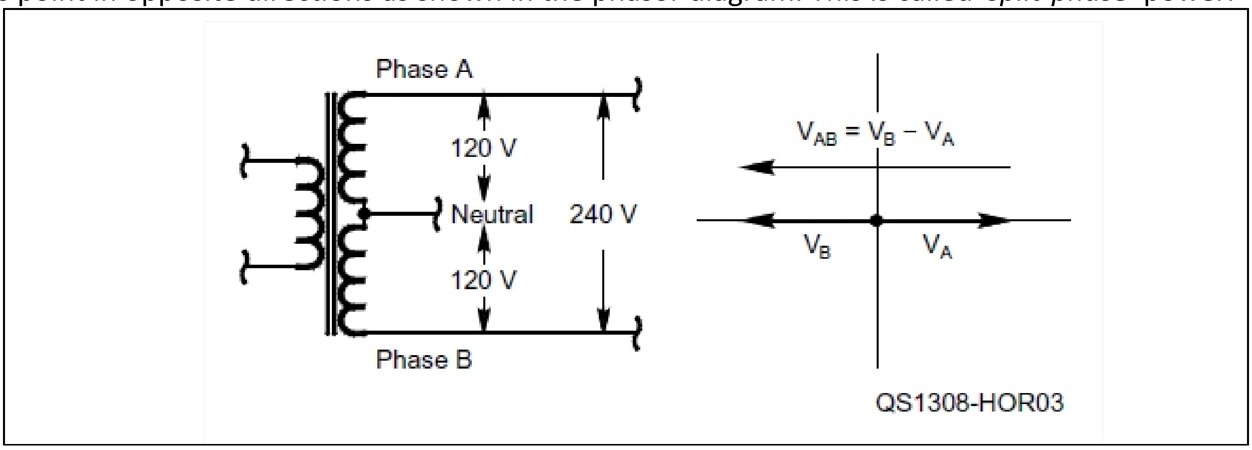

Figure 3 — Residential split-phase ac power.

If you have two equal-and-opposite phasors, what is the magnitude of the voltage between them? (Answer: The sum of the phasor magnitudes.) If each phasor has a magnitude of 120 V, the magnitude of the voltage between them is 120 + 120 = 240 V. If you connect one hot wire to each phase and one to neutral, that's where the ac for your amplifier (or your clothes dryer) comes from!

Three-phase Power

Now let's take this one step further — three-phase power.

The ac coming from generating facilities like dams and power plants has three phases. That's why there are three wires (or pairs of wires) making up the high-voltage lines (not counting any protective ground wires). Large power consumers would unbalance the power grid if they used power from just one of the phases, so they are wired to use some power from each of the phases and the electricians are in charge of configuring things so each phase is loaded by about the same amount. That is why buildings and businesses of any size have ac service with three phases, not just two.

The phasors representing each of the three phases — A, B, and C — are shown in Figure 4. They are all spaced equally around the circle, 1⁄3 of the circumference or 120° apart. Let's say your apartment in a big building is supplied with two phases of power, just like residential split-phase ac power, and each phasor has an amplitude of 120 V. What happens when you try to run the drier by connecting it to the two phases (let's say phase A and B)? Why don't you get 240 V?

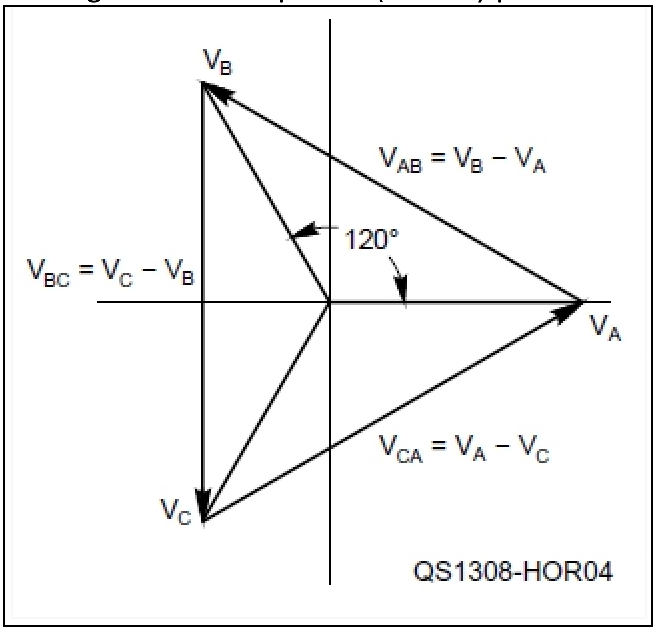

Figure 4 — Phasor diagram for three-phase power.

Look at Figure 4 and the phasor V_{BA}. The two phasors representing the phases of electrical service, V_A and V_B, are not pointing in opposite directions — they are only 120° apart — so their magnitudes don't add as in the split-phase situation. In fact, if you look up the trigonometry, the magnitude of phasor $V_{BA} = \sqrt{3}\, V_A = 1.732\, V_A$, not $2\, V_A$. If each phase is supplying 120 V, what voltage will your dryer see if it is connected across two phases? (120 × 1.732 = 208 V)

This dependence on how your electrical service is derived from the utility grid makes a big difference when running a heavy load — such as an amplifier. If your amplifier is designed to run from 240 V power and you connect it to 224 V instead, that is about 7% low. It's common for amplifiers not to supply their full rated output power when run at slightly lower input voltage. The opposite case — higher than expected input voltage — can stress high-voltage components, too.

If your equipment does not have a "universal" power rating of something like 90 to 260 V ac, determine how to configure it for the voltage you have available. Many appliances and amplifiers have selectable input voltage "taps" or connections on the primary winding of a power transformer that can accommodate 240, 220 or 208 V power. (Where does 208 come from? Two hundred and eight is approximately 1.732 × 120 V, the usual voltage for home ac service.)

I said we'd have a two-part article, but it will take one more to get to some real radio meat-and-potatoes: AM and PM modulation from the perspective of phasors. That, in turn will usher you to the gates of modern data communication: I-Q modulation.

Notes

1

www.arrl.org/studying-for-the-general-license

, click on "Math Tutorials," then "Tutorials on Math for License Exams."

2

A thorough treatment of three-phase power in both Y and delta connections is available online at

www.ece.msstate.edu/~donohoe/ece3414three_phase_power.pdf

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Experiment #128 — Phasors, Part 3

I confess to using the wrong value for √3 in last month's column. It's 1.732. The implications of that mistake are addressed on the "Hands-On Radio" web page.
[1]

Now, on to modulation!

What if, as seems to happen on a regular basis, one tuner-upper attracts a competing tuner-upper with another unmodulated carrier identical in frequency and amplitude except for having a slightly different phase — say 45° — ahead of the original carrier? The new signal's phasor is given as A∠45°, just ahead of the first signal by 45°. Even though both phasors are rotating around the origin, that relationship never changes.

Since both of the signal phasors have the same frequency, why not do away with the rotating and look only at the differences? What would happen if you take a seat on the first carrier's phasor, looking out toward the arrow's head from the origin, and spin around with it? From your new perspective, the phasor doesn't move or change at all because you're rotating with it at the same rate (frequency) and its length (amplitude) is constant. The second carrier with the 45° phase difference is pointed off to the left, halfway between straight ahead and to your left. It, too, doesn't move or change, but the phase difference means it points in a different direction.

Let's say that the competing tuner-upper starts to drift down a little bit in frequency. As the frequency of the second signal drops, the rate at which its phasor rotates gets a little slower, too. That means it will start to fall behind the original phasor and from your perspective, the second phasor appears to rotate clockwise or backwards according to our counterclockwise-equals-positive convention. The lower the second signal's frequency, the faster it rotates backwards. Let's say the second signal stabilizes at a frequency 1 Hz lower. To you, it appears to rotate backwards, passing backwards across your phasor once per second. Similarly, if the frequency of the second signal increases, it will appear to rotate counterclockwise.

Another possibility is that the phase of the second signal (with respect to the original signal) jumps around. In this case, what you would see is the phasor for the second signal shifting its position relative to the first signal — sometimes ahead, sometimes behind.

AM from the Phasor Point of View

AM produces three signals when a carrier is multiplied by a modulating signal. The first signal is the carrier with frequency, f_c. If the modulating signal is a single tone with frequency, f_m, two sidebands are created with frequencies, f_c+f_m (the upper sideband) and f_c-f_m (the lower sideband). See the Modulation chapter of the *ARRL Handbook*.
[2]

Each of these signals can be treated as a phasor and the trio can be added together as we discussed in the previous column.

The amplitude of the three phasors doesn't change but their relative directions do. Figure 1A shows what the three phasors look like from your perspective, sitting comfortably on the carrier phasor rotating at the carrier frequency, f_c. Since the upper sideband (USB) phasor has a higher frequency than the carrier, you see it rotating counterclockwise at the modulating frequency, f_m. Similarly, you see the LSB phasor rotating clockwise at f_m. (Viewed all by themselves, the USB and LSB phasors are actually rotating at $f_c \pm f_m$.)

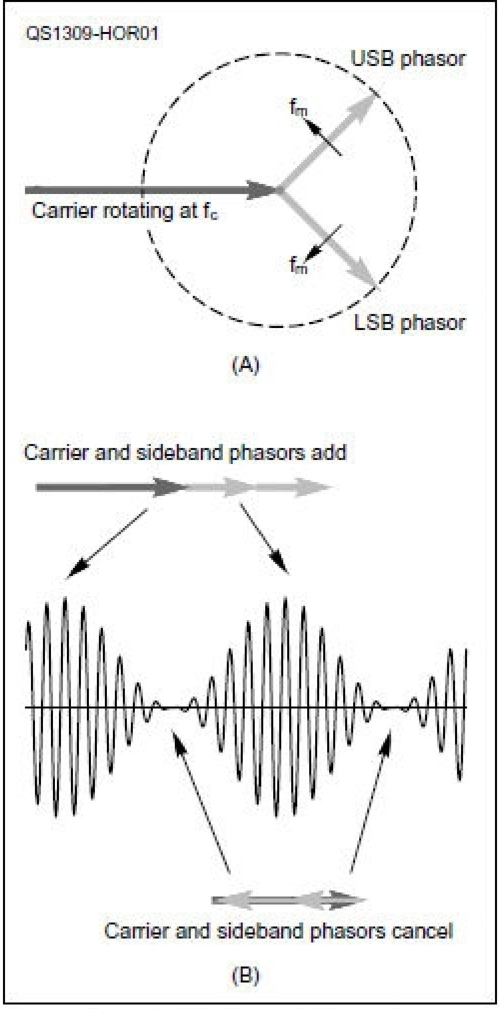

Figure 1 — Amplitude modulation shown as a combination of three phasors. At A, the sideband phasors are shown rotating in opposite directions at the frequency of the modulating signal, f_m, with respect to the carrier which is rotating at f_c. 100% modulation is shown at B in which each sideband has half the carrier amplitude so that the sum of all three phasors varies from zero to twice the unmodulated carrier amplitude.

Note that these counter-rotating sideband phasors have the same amplitude and are always ahead of or behind the carrier phasor. Think about what this means for the sum of the three phasors. Using the tip-to-tail method of adding phasors, the resulting AM signal's phasor will always be aligned with the carrier phasor because of the symmetry of the sideband phasors. However, the amplitude of the AM phasor will grow and shrink as the two sidebands add to, then oppose, the carrier phasor.

What happens if each sideband has exactly half the amplitude of the carrier? When the sideband phasors are both "pointing out" the resulting AM phasor's amplitude equals the sum of the carrier plus the two sidebands: twice the original carrier's amplitude. When the sideband phasors are "pointing in" their sum cancels with the carrier and there is no signal. Thus, the AM phasor's amplitude varies from zero to twice that of the original carrier — just as you see in Figure 1B, which represents 100% modulation.

FM and PM from the Phasor Point of View

From the standpoint of the unmodulated carrier, the phasor of an FM or PM signal moves ahead and behind that of the carrier as the amplitude of the modulating signal changes. (For the rest of this column, FM will be used to mean both FM and PM.)

Just as for AM, a pair of counter-rotating sideband phasors with frequencies of $f_c \pm f_m$ add and cancel just as for AM. Unlike AM, however, they are oriented so that they are creating a separate modulating phasor at right angles to the carrier phasor as in Figure 2A. The resulting FM phasor created by the sum of the carrier and the modulating phasor shifts ahead of and behind its unmodulated position as in Figure 2B.

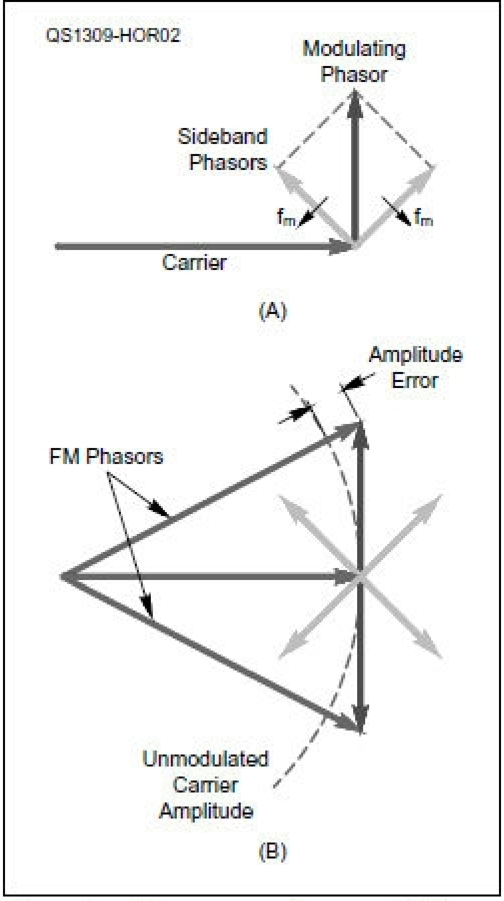

Figure 2 — Frequency or phase modulation shown as a combination of three phasors. A shows sideband phasors similar to AM but oriented so they create a modulation phasor at right angles to the carrier. B shows the FM phasor moving ahead of and behind the carrier according to the alignment of the sideband phasors. Using only one set of sidebands reduces bandwidth but creates a small amplitude error in the resulting signal.

It's not that simple, however, because FM and PM signals have constant amplitudes — only the frequency (or phase) may shift with modulation. That means the final sum of the phasors must have a constant amplitude, that of the original unmodulated carrier, shown as the arc in Figure 2B. The figure shows the small amplitude error created by including just the one set of modulation sidebands. When the modulation level is low, the error is small enough that one pair of modulation sidebands is acceptable and this is called "narrowband FM."

As the modulation level increases ("wideband" FM) and the resulting FM phasor moves farther and farther from the unmodulated carrier, the resulting amplitude error would become larger. To keep the FM phasor close enough to the required amplitude, additional sets of sideband phasors are required. Each successive set operates at right angles to the previous set. This is the complex set of sidebands.

IQ Modulation with Phasors

As Figure 3 shows, the phasor of a modulated carrier moves around in an area defined by whether the modulation is AM or FM. If AM, the movement is horizontal, changing the phasor's amplitude. If FM, the movement is along an arc, changing the relative phase. There's no reason a signal can't have both AM and FM components with the resulting phasor located anywhere within the indicated area.

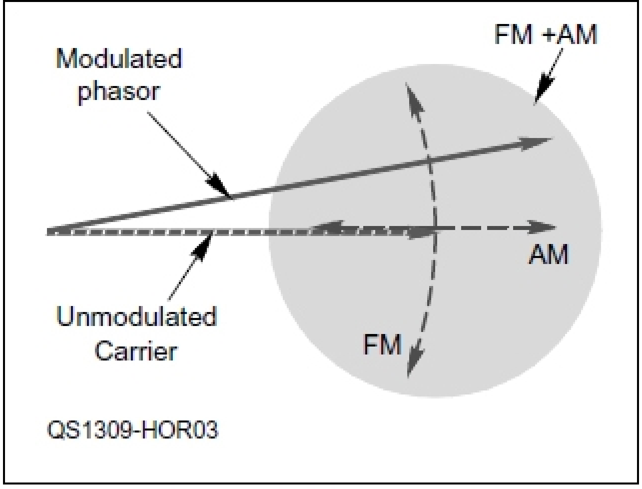

Figure 3 — The effects of AM and FM modulation on an unmodulated carrier. Combining AM and FM results in a two-dimensional region for the resulting signal phasor.

Oversimplifying to a degree, this is what IQ modulation is in which two different modulated signals are combined: the I signal (for *in phase*) and the Q signal (for *quadrature*). Both the I and Q signals are regular carrier signals, but the Q signal is 90° ahead of the I signal as shown in Figure 4. Modulating the I and Q signals independently and combining them can cause the resulting phasor to move around in the pattern of any of the AM or FM phasors discussed previously.

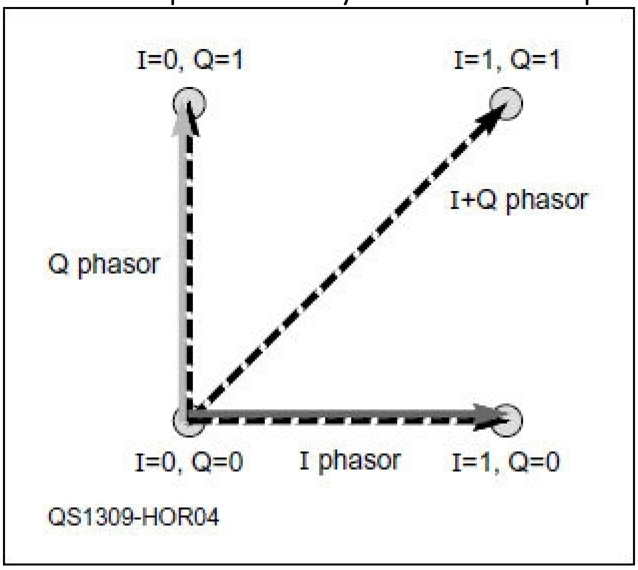

Figure 4 — IQ modulation uses two independent carriers, I (in-phase) and Q (quadrature) shifted 90 degrees. By turning the I and Q carriers on and off, the resulting signal phasor takes any of four positions in the resulting constellation diagram, representing four different data symbols.

Digital data can be transmitted by turning the I and Q signals on (1) and off (0) independently (also called *amplitude shift keying*), creating four possible combinations (00, 01, 10, and 11). By adding the on-or-off I and Q phasors together, the result is four different phasors shown in Figure 4. This is called *quadrature amplitude modulation* or *QAM* and each position of the phasor is called a *symbol*. If there are four possible symbols, it is called 4-QAM. A receiver demodulates the I and Q signals separately and decodes the phasors into the same on/off combinations, reproducing the same stream of digital data.

From your perspective, sitting on the I signal's phasor, the end points of the four phasors form a square called the modulation's *constellation diagram*. Complex schemes with hundreds of points in the constellation have been devised — for example, digital cable TV signals use 64 or 256 points, called 64-QAM or 256-QAM, respectively.

All this from simple rotation! The interested reader may want to tackle additional information found online. You can learn more about IQ Modulation at

www.home.agilent.com/upload/cmc_upload/All/IQ_Modulation.htm?cmpid=zzfindnw_iqmod

, Amplitude and Frequency/Phase Modulation at

www.zhinst.com/blogs/michele/files/downloads/2012/12/AMFM.pdf

and Digital Modulation at

http://ee.eng.usm.my/eeacad/mandeep/EEE436/CHAPTER2.pdf
.
Nevertheless, even if you stop here, you'll have traveled from a basic definition of phasors to how they can be used to visualize the modulation processes we use every day.

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at
www.arrl.org/hands-on-radio
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Available from your ARRL dealer, or from the ARRL Store, ARRL order no. 6948. Telephone toll-free in the US 888-277-5289, or 860-594-0355, fax 860-594-0303;
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pubsales@arrl.org
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Experiment #134 — Effects of Junction Temperature

The equations describing how a bipolar junction transistor circuit works are usually simplified to ratios of resistances or currents. Many assumptions are made so that the calculations are straightforward — and most of the time this works just fine. The designer and technician should understand that underlying the simplifications and assumptions is a fundamental relationship that can have a major effect on how a circuit behaves as temperature changes. This month, you'll observe the effect directly.

PN Junction Volt-Amp Characteristic

You've probably seen the basic graphs of Figure 1, perhaps combined in a single graph with different current scales for forward and reverse current. These graphs show the relationship between current and voltage — the *I-V characteristic* — for any semiconductor PN junction. The equation that generates the graphs is called the *Fundamental Diode Equation*. With positive for both voltage and current defined as from the P-type to the N-type material:

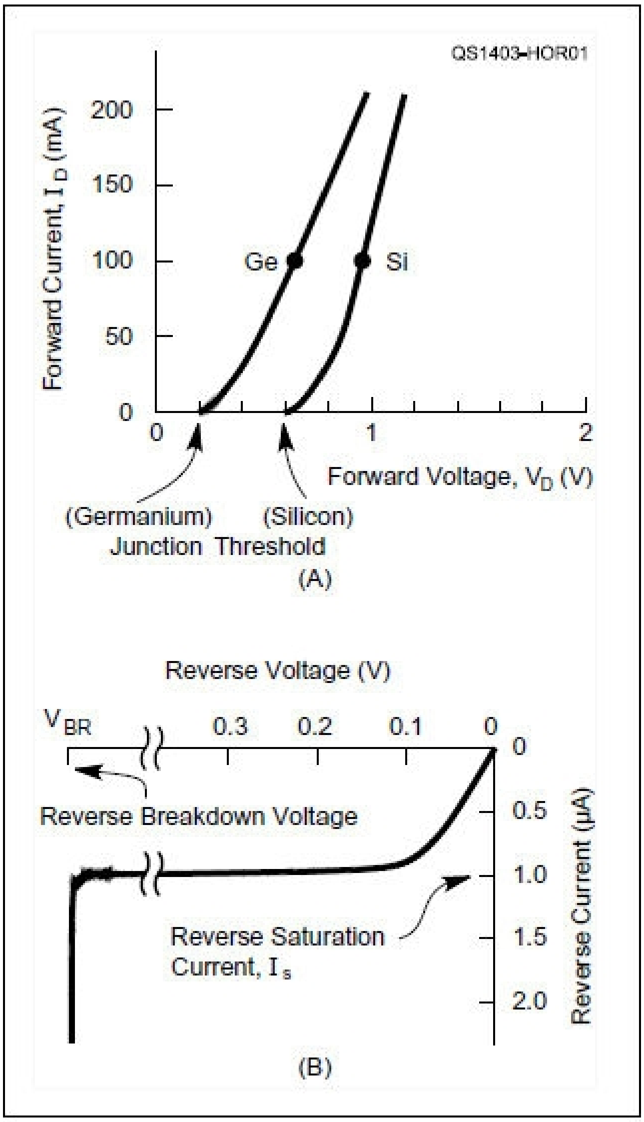

Figure 1 — The basic I-V characteristic of a semiconductor junction diode shown for forward bias in (A) and for reverse bias in (B).

$$I_D = I_S\left(e^{\frac{V_D}{\eta V_T}} - 1\right)$$

where I_D is the *forward current* through the diode, V_D is the *forward voltage* across the diode, and I_S is the *reverse-bias saturation current* that flows when V_D is negative. (I_S is measured at the lowest voltage that will produce a stable current level — typically a few tens of mV of reverse bias and is much smaller than the typical values for *reverse leakage current* specified in datasheets.) η is the *ideality factor* (also called the *quality factor* or *emission coefficient*) that depends on how the carriers of current recombine at the diode junction and varies from 1 to 2. For normal currents, η = 2 works well for most diodes.

The most interesting bit of this equation is the *thermal voltage,* $V_T = kT/q$ where k is the Boltzmann constant that relates the energy of a particle (the electron) and its temperature, T is the absolute temperature in degrees Kelvin, and q is the charge of an electron. At room temperature (300 K is commonly used in simulation software), V_T = 28.85 mV. A common simplification is that $V_T \approx T / 11{,}600$.

What happens as temperature increases? Since V_T goes up along with T, if everything else on the right side of the equation stays constant, I_D "should" go down. How about an experiment? Figure 2 shows a simple test circuit for evaluating a diode's I-V characteristic. V and A represent a voltmeter and ammeter, respectively. Assuming you are using multimeters, set them to read 0 – 1 V and an initial current of 0 – 200 or 0 – 300 μA.

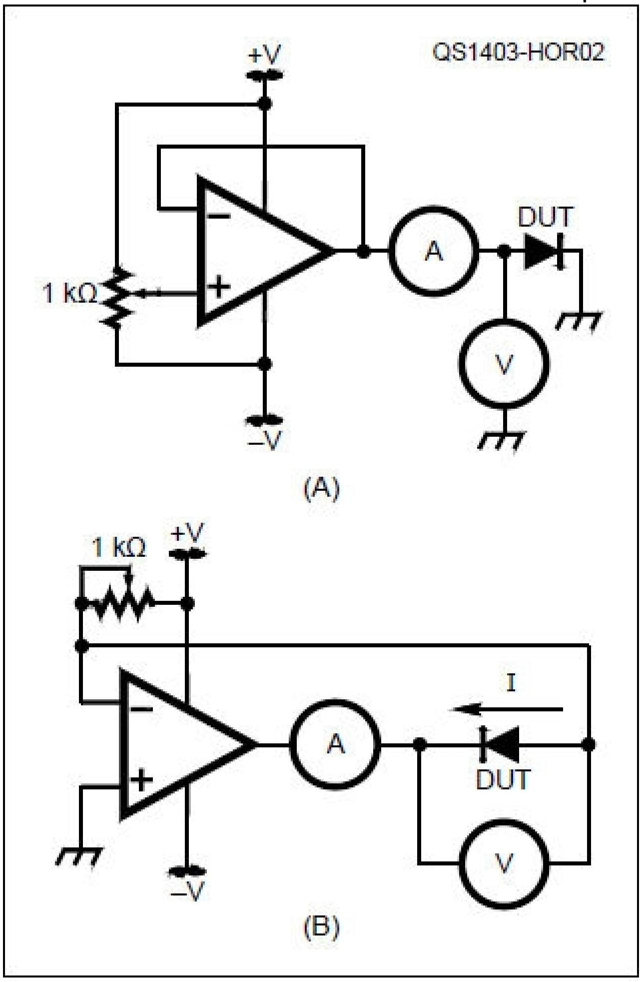

Figure 2 — A test circuit for measuring a diode's I-V characteristic. DUT stands for Diode Under Test. At A, the op-amp is configured as a voltage source. At B, the op-amp is configured as a current source. Note that the diode is reversed from A to B.

The op-amp is connected as an adjustable voltage source with its output set by the 10 kΩ pot. Any garden variety op-amp will suffice, such as the venerable LM741. Your power supply needs to have an output of at least ±3 V and up to ± 15 V will do if it is within the op-amp's maximum voltage ratings.

Select an ordinary silicon diode such as a 1N914, 1N4148, or 1N4000-series part. Before connecting the diode, set the op-amp output voltage to 0.3 V on the voltmeter. Then connect the diode and slowly increase the voltage to 0.7 or 0.75 V in 0.05 V steps, recording both voltage and current. (A spreadsheet to generate a graph is provided on the Hands-On Radio web page.

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) The measurements start at 0.3 V because current is too low to be measured with ordinary test equipment below that level. Most multimeters are not very accurate at the low end of their ranges so expect your measured values to diverge quite a bit from the calculated values.

Now cool the diode by at least 30 °C and measure the currents again. One way to get a relatively consistent temperature during the measurements is to put a metal object that is many times larger than the diode, such as a large nut, in your home freezer for an hour or so. Orient the diode so you can sit the object directly on it. Measure current through the diode and when the reading stabilizes begin taking data — you'll have to work fast so the temperature remains about the same.

Why does I_D have different values when the diode is cooled? Because the value of I_S is strongly dependent on temperature, as is V_T. In fact, for silicon diodes, the value of I_S changes by about 7 % / °C, a positive shift with temperature. That means I_S will double (or halve) with every 10 °C increase (or decrease) in temperature!

The two temperature dependencies for V_T and I_S work against each other. As temperature increases, V_T goes up which works to lower I_D because it is in the exponent's denominator. On the other hand, increasing I_S causes I_D to increase. Thus, it is a balancing act with the change in I_S having the larger effect. While the exact change requires some detailed calculations, around room temperature the net result is that V_D changes about –2.2 mV / °C if current through the diode is held constant.

Next, go the other way: heat the diode by about the same amount (putting the metal object in hot water will do the job) and take the same set of measurements. Figure 3 shows a set of data I took for a 1N4148 diode. As you can see, below about 100 µA, my measurements started to show some signs of being inaccurate. (This is typical of low-level home-lab test setups, so don't be too concerned if the data doesn't make a nice straight line.) The diode you choose and the temperatures you obtain will probably give significantly different values of current — all you are trying to do is observe the effect of temperature.

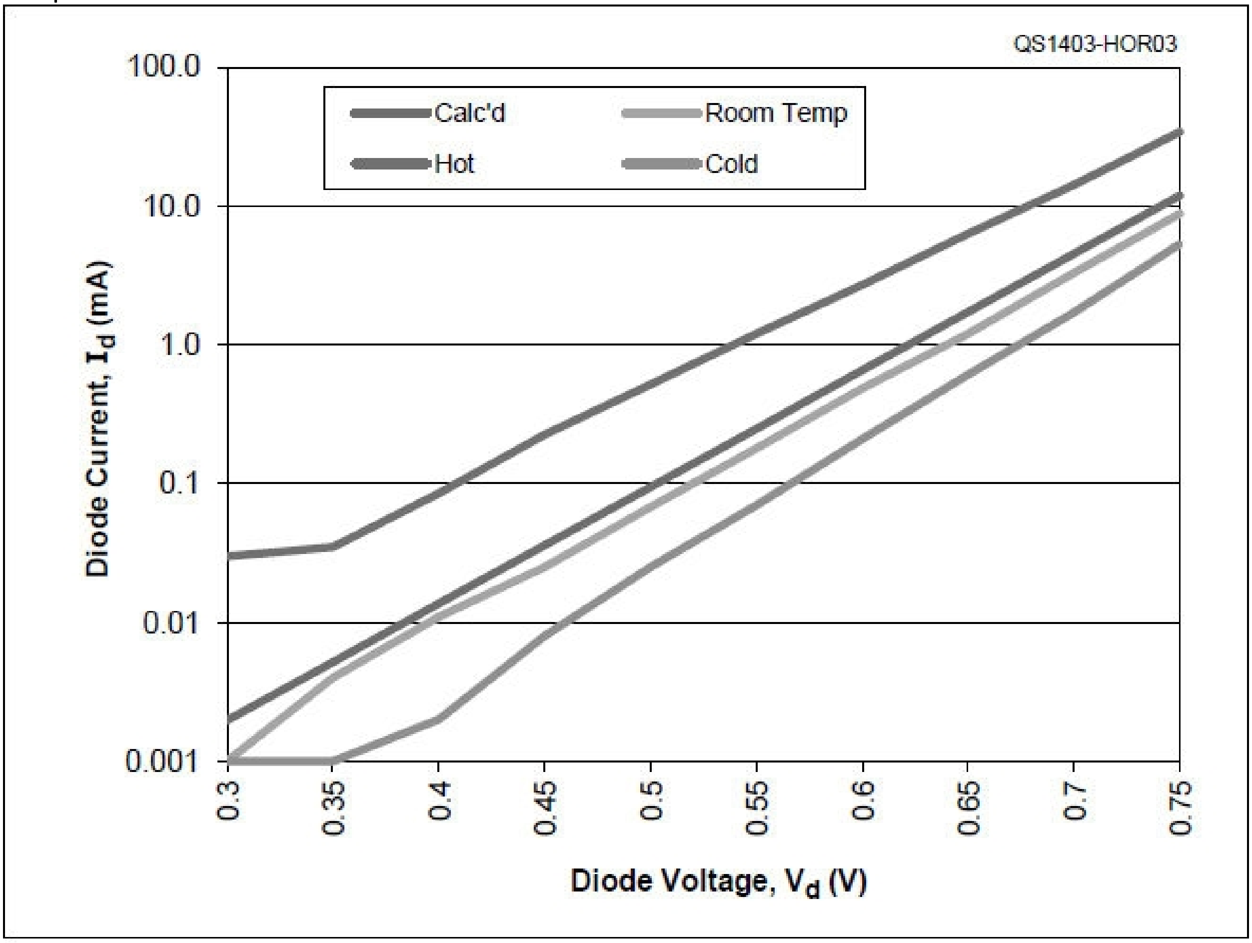

Figure 3 — A diode's I-V characteristic of forward current, I_D, versus forward voltage, V_D, depends heavily on temperature. Increasing temperature causes higher current for a given voltage. The values plotted at different temperatures are for a 1N4148 measured by the author.

If you'd like some extra credit, try taking the measurements by controlling I_D and measuring V_D. Figure 2B shows how to turn the op-amp into a current source. (See Experiment #3 — Op-Amps for an explanation of this circuit.)[1] Start with a short circuit instead of the diode and confirm that with the pot set to maximum resistance, approximately +V / 10 kΩ of current is flowing. Reinstall the diode and adjust the current to get about the same values of V_D as in the previous set of measurements. You should see fairly similar results.

Double your extra credit by trying a germanium diode, such as the common 1N34A. Germanium (Ge) diodes have a much higher value of I_S than do silicon (Si) diodes by three to four orders of magnitude! Thus, the current values you measure will be much larger for a given value of V_D. That can work to our advantage, however, because for a given amount of I_D, a germanium diode will have a lower V_D than a silicon diode. That may be an advantage in a sensitive circuit, such as a diode detector.

While you are at it, remember that a bipolar junction transistor is constructed from a pair of back-to-back PN junctions. Substitute the collector-base or emitter-base junction of an inexpensive PNP (2N3906 or 2N4404) or NPN (2N3904 or 2N4401) transistor for the diode. The resulting measurements should be fairly similar. Remember that temperature dependency when designing your next transistor amplifier!

Temperature Sensing

Since the effect of temperature on diode current and voltage are so predictable, it's quite possible to make a temperature sensor out of a diode. Rearranging the diode equation gives

$$V_D = \frac{\eta T}{11600} \ln\left(\frac{I_D}{I_S}\right)$$

So you can see that measuring V_D while holding I_D constant gives a pretty good idea of temperature. Using a microprocessor to "do the math" or applying a comparator circuit to detect when V_D crosses a threshold is an excellent method of temperature control at low cost.

Parts list

All parts can be substituted by any equivalent.

Op-amp — LM741

Silicon diode — 1N4148 or 1N4000-series

Germanium diode — LM34A

Transistor — 2N4401/4403 or 2N3904/3906

Potentiometer — 10 kΩ

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All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**

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Experiment #138 — E versus V

One of the first bits of technical lore every ham, indeed, every electronic hobbyist learns is the venerable Ohm's Law. This simple formula explains the relationship between current through a material with a known resistance, and the voltage across the material. Students dutifully memorize it in all its forms: $I = E / R$, $E = I \times R$, and $R = E / I$ but often ask the question, "Where do E and I come from?" They understand the use of *R* to represent resistance. The use of *E* for voltage and *I* for current is a little confusing. Why not *V* and *C* or *V* and *A*? That's a very good question and its answer goes back to the beginning of the electric era.

Instructors correctly respond that "*E* stands for *electromotive force* or *emf*." Isn't that just a fancy way of saying "voltage?" Not really. Let's go back to the early 1800s, when experimenters like Faraday and Henry and Ørsted were discovering electric and magnetic fields, the relationships between them, and their ability to influence the movement of a mysterious substance known as electric charge. (Some of these early discoveries are related in Hands-On Radio Experiments #117 and #118.[1])

Before the Electron

The early experimenters were completely in the dark about what was moving around in their wires and reacting to the presence or motion of electric or magnetic fields. Even Faraday's concept of a field that explained how energy could be distributed in space was radical and new. So all they knew was electric charge and that there was a force that could make the charge move — the electromotive force.

Remember that no one at the time had the slightest idea what electric charge was. Dalton's description of the atomic role in chemistry had only been introduced between 1802 and 1805, and Lord Kelvin's discovery and description of the electron as an individual particle didn't come until 1897. Bohr's model of the hydrogen atom with electrons arranged in shells around a nucleus and which could leave an atom and move about, was not developed until 1913. Even today, we don't really understand what electric charge actually is, why it exists, or even what the electron consists of. That those early experimenters were able to ascertain fundamental laws which still hold true today is nothing short of remarkable!

From E to V

Yes, yes, but what about *E* and *V*? If you browse through an introductory electrical engineering textbook that introduces students to the world of electric fields and circuits, you will likely see that the book begins with these same basic relationships between fields and current. Students learn a general description of what makes electrons move in response to electric fields and magnetic fields and the initial equations usually use *E* or *e* to represent the electromotive force. (The symbol ε is also used.)

In this general environment, it is appropriate for the equations to use electromotive force. The resulting electron motion is analyzed through a volume or across a surface that has some *conductivity* (σ) which describes the ease with which electrons move through it. Figure 1 shows a diagram of charge moving through a cross-section area, A. *Resistivity* (ρ) is the reciprocal of conductivity, so that $\rho = 1 / \sigma$. Resistivity is measured as ohm-meters or ohm-m.

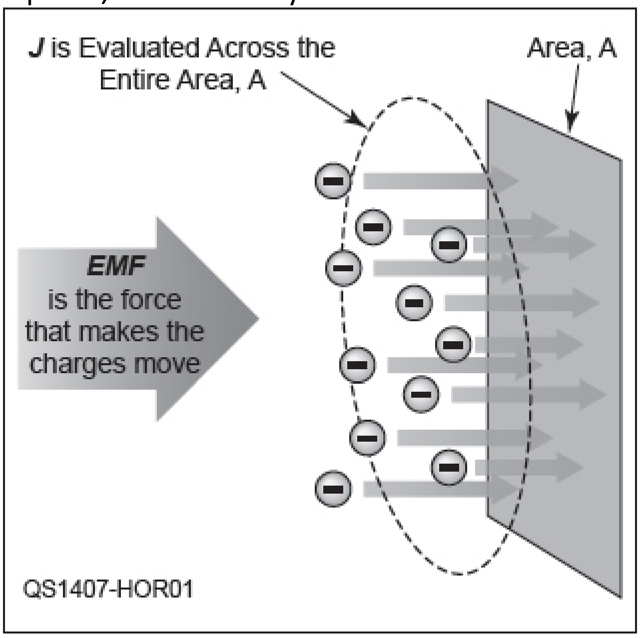

Figure 1 — In the most general sense, electromotive force, or emf, causes electrical charge to move through some area.

So far, we haven't seen a hint of a circuit and all the equations are using *E* and the general parameter *current density* (*J*) which is the amount of current per unit of area. Boring! When do we get to the good stuff? Patience is eventually rewarded as the conducting material is formed into a thin loop — a circuit — and subjected to electric and magnetic

fields. Now the electrons are moving around in this circuit and *J* can be replaced by *current*, *I*, the flow of all electric charge through the entire cross-sectional area of this thin conducting volume.
2

Similarly, it now makes sense to talk about the loop's *resistance*, represented by *R*, since the current is now flowing in this highly constrained path. R accounts for all of the material's resistance and for a cylindrical wire as in Figure 2, $R = \rho\ l / A$, where *l* is the wire's length, *A* is its cross-sectional area, and ρ is the material's resistivity. (We are only referring to dc, so the *skin effect* of ac current does not apply.)

We still have no batteries or power supplies, just an externally generated field that is a source of emf, represented by *E* and so the relationship between the three parameters becomes the familiar $I = E / R$. Here *E* represents the emf developed all the way around the loop by the external field. If you did the experiments of Experiment #118, you observed this current as the magnet moved near the coil of wire which made a long loop.

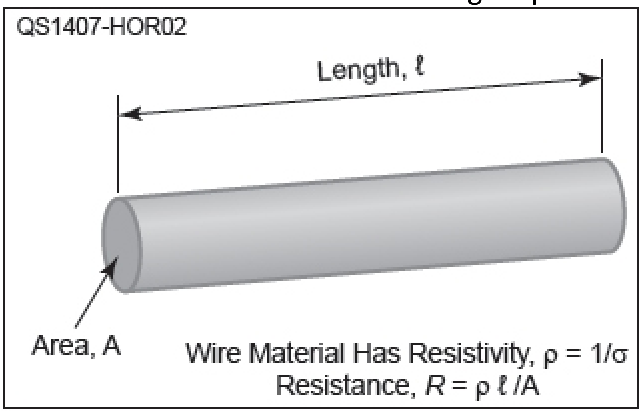

Figure 2 — The resistance, *R*, of an entire wire can be calculated from its area, length, and the resistivity or conductivity of the material the wire is made of.

Figure 3 shows how to see this emf by constructing your own loop and placing it near a power transformer or motor which is the source of a changing magnetic field. (For a field to generate emf in a loop that makes a current flow, the field must be changing or the circuit must be moving through the field.)

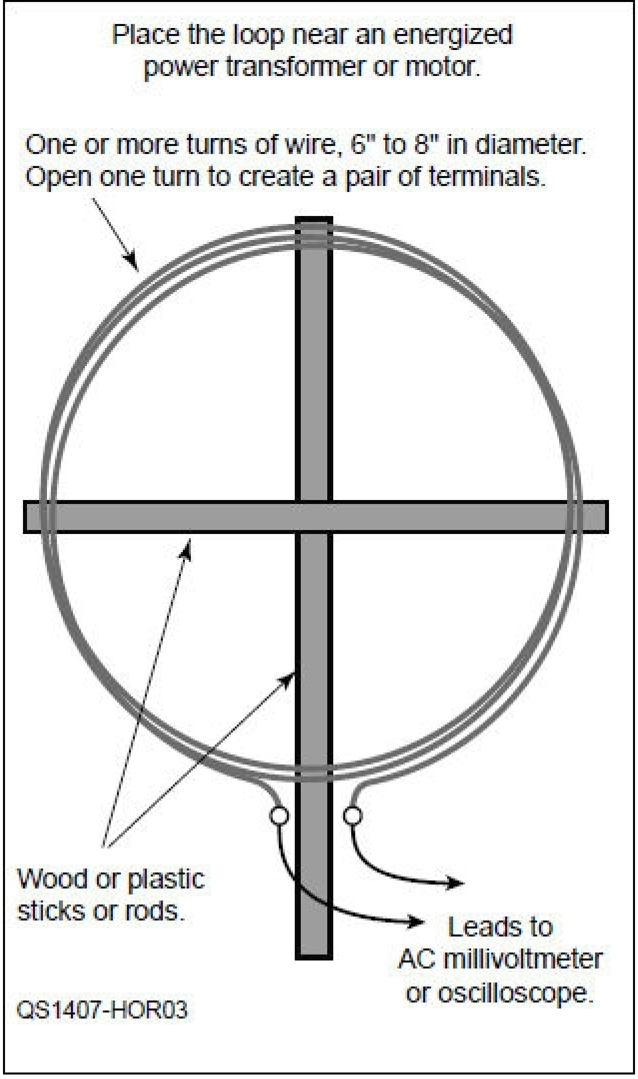

Figure 3— Emf developed by a field can be measured by open-circuiting a loop of one or more turns and measuring the voltage across it with a sensitive voltmeter or oscilloscope.

To measure the emf the loop is cut as in Figure 3, creating a pair of terminals. Current flow stops, of course, but an *electric potential* now exists between those terminals.[3]

By measuring it, we can tell how much energy (in joules) an amount of electric charge (in coulombs) will gain from the effects of the field on the loop. Electric potential is thus measured in joules/coulomb and the value of 1 joule/coulomb defined as 1 volt in honor of Alexander Volta, inventor of the battery. Electric potential was given the simpler name of *voltage* as well. The symbol for volts is the familiar V and voltage is represented in an equation by V or v.

Can V be substituted for E without risk of confusion? If the discussion is of ordinary electronic circuits, the answer is almost always yes. It is safe to say that substituting $I = V / R$ for $I = E / R$ will not get you into trouble around ham radio equipment! You should use E, however, when the discussion is about fields — for motors, generators, antennas, transformers, etc. Another place you'll see emf is the term *kickback emf* that refers to the reverse voltages generated by inductances when current through them is suddenly interrupted or turned off, causing the magnetic field created by current through the inductance. Similarly, a motor develops a *back emf* that opposes the applied voltage, limited current through the motor to only that required to drive the load and account for internal losses.

Don't confuse units of measurement, such as volts (V), amperes (A), and ohms (Ω), with their corresponding physical phenomena, such as electric potential (also V), current, (I), or resistance (R). For example, Ohm's Law is *never* written as A = V / Ω.

We have answered almost all of the original question but what about the mysterious I? Returning back to the early days, experimenters had learned to discriminate between a quantity of charge and the flow of charge (current). The French experimenter Ampere (for whom the unit of current was later named) gave flow of charge the name "Intensité de Courant" (Intensity of Current) and assigned it the symbol I or i in equations. (This is one reason why electrical engineers use j instead of the mathematician's i to represent the imaginary square root of negative one in their equat ions.)

The ampere is a flow of one coulomb (C) per second. How many electrons make up 1 C? 6.24×10^{18} electrons! *Basic Radio* from 1942 gives us an idea of how many that is: "If 3 million people were to count for 8 hours per day at the rate of

200 per minute, they would have to count from the time of the Trojan Wars in 500 BC down to the present" in order to count the number of electrons passing by at a rate of 1 ampere.

Another source of frequent confusion is the overuse of the symbol *C*, which is used at various points to represent coulombs, capacitance (measured in farads with the symbol F), battery capacity and battery charge or discharge rates, not to mention the speed of light, *c* — see?

In closing, I would like to acknowledge the comments of Dana Brown, AD5VC, and Sam Neal, N5AF, on the "ham_instructor" e-mail reflector.[4]

Sharing expertise is one of the hallmarks of Amateur Radio and it is particularly important for instructors so that our newest licensees have a common (and correct) understanding of radio's physical environment.

Notes

1 All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2 As an example — if a current, *I*, of 1 ampere is flowing in a 16 AWG wire 0.051 inches in diameter, the wire has a cross-sectional area of $\pi d^2/4$ = 0.00204 in^2 and the corresponding current density, J = 1 / 0.00204 = 490 A / in^2. If the wire is made smaller, *I* will not change but *J* will increase.

3 The electric potential across the open-circuited terminals of a source of emf is also known as the *open-circuit voltage*.

4 **https://groups.yahoo.com/neo/groups/ham_instructor/info**

Experiment #139 — Digital Code Basics

While tuning the bands, I've become more and more fascinated by the sheer variety of "digital" signals I encounter. Ever since the FCC allowed amateurs to develop their own protocols and codes in the early 1990s, we have seen an explosion of amateur digital innovation. The *Fldigi* software includes support for more than 30 digital modes, along with numerous variations on the protocols and modulations. *WSJT*, originally developed for VHF meteor scatter, is now a full suite of protocols. *PACTOR* and *WINMOR* are pushing the boundaries of digital data transfer over the very difficult HF channel. For voice communication, *CODEC2* is now a fully-capable digital voice protocol. More are on the way as amateurs put the processing power of the modern PC to work, even copying that most venerable of all digital modes, CW!

Regardless of whether you are a phone, CW, or digital fan (or all three!) it's important to understand the basics of modern digital technology. There are certain bits of vocabulary that describe the data stream that underlies wireless data. This column will grapple with a few.

Bits of Data

Let's start with that most elementary of digital concepts — the *bit*. This is the 1-or-0 information element from which all other data is constructed. When using CW, RTTY, or PSK31, bits are what you hear being transmitted, either as a tone or as the presence of a carrier.

As you can see from Figure 1, the Morse code character for E is *seven* bits long — three 0 bits, a single 1 bit, and a concluding three 0 bits! Morse code is composed of two fundamental elements: the dot (labeled 1 in the figure) and the inter-element space (labeled 0 in the figure). Dashes are composed of three dot elements with no intervening space and the inter-character space is composed of three inter-element spaces.

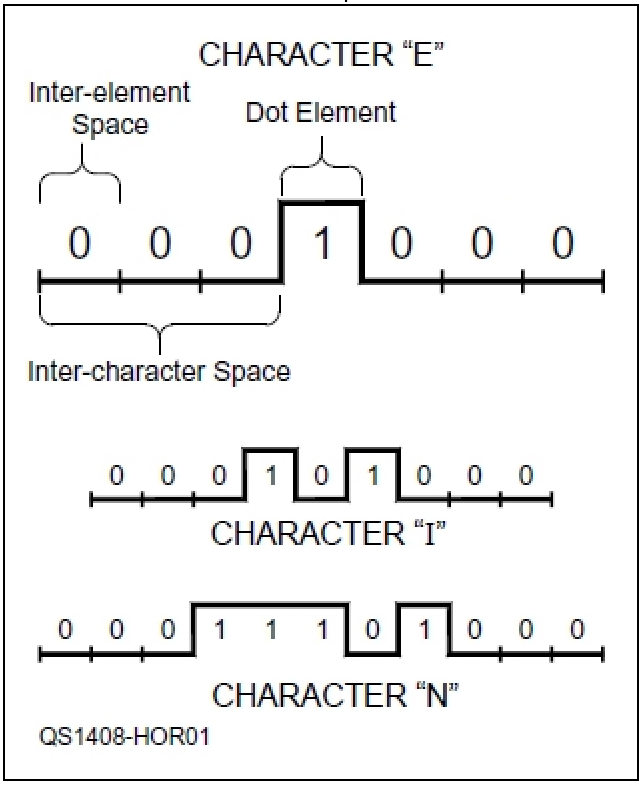

Figure 1 —The elements of Morse code are dots and spaces of equal lengths. Dashes are made up of three consecutive dots and inter-character spaces are made up of three consecutive spaces.

The bits are assembled into characters on the receiving end by the human operator in the case of CW or by software in the case of RTTY or PSK31. The decoding process also requires some information about when a character's-worth of bits begins and ends, so *framing bits* are added. In the case of Morse code (CW), the framing bits are actually the inter-element spaces: three inter-element spaces in a row means a character has just finished or is just about to start. Don't discount the value of framing bits — if you've ever tried to copy CW with spaces that are too short, you know how important framing bits are!

Know the Codes

Amateurs refer to Morse as "the code" but there are really lots of codes in the digital world. In this sense, a code is simply a method of representing characters as a pattern of bits. To *encode* a character means to turn it into its proper bit pattern and to *decode* it means to turn the pattern of bits back into the character. There may be additional codes operating on the characters such as compression (like a ZIP file) or abbreviations (like Q signals) but at the basic level all we're talking about is the rules for turning whatever represents 1s and 0s into characters and vice versa.

Morse and PSK31 are somewhat unusual in that the required number of bits to send a character varies from character to character. Not including the framing bits or prosigns, Morse code requires 1 bit for its shortest character (E) and several characters require 19 bits. These are examples of *Huffman codes,* in which the length of the individual characters are controlled to improve the rate at which information can be transmitted.

Most other codes are *fixed-length*, such as RTTY's Baudot code (named for its inventor, Emile Baudot) and the common computer character code, ASCII. Figure 2 shows how a character is constructed from *mark* and *space* elements in the Baudot code. Each Baudot character consists of five consecutive bits with no intervening elements.

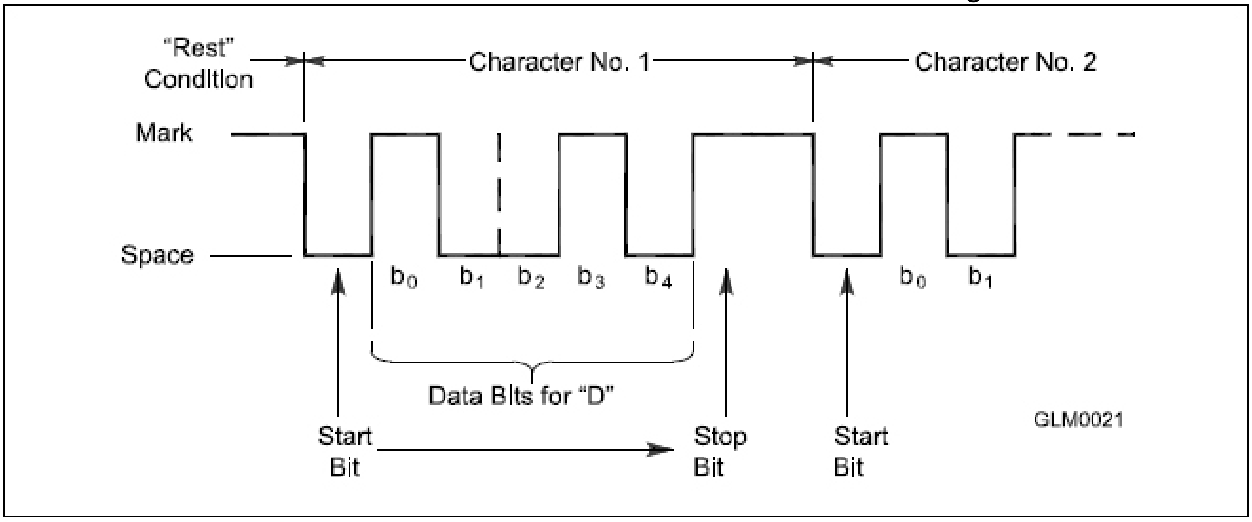

Figure 2 — The Baudot code's elements are termed mark and space. Over a radio channel, mark and space are represented by different audio tones. The bit on the right is sent and received first.

Baudot has two framing bits: a *start bit*, which consists of one bit period of the space element and a *stop bit*, which consists of at least one bit period of the mark element. Stop bits can be configured to have a minimum length of 1, 1.5, or 2 bit periods so that the decoder functions properly.

ASCII characters can have 7- or 8-character data bits plus the start and stop bits. A *parity bit* can also be added to indicate whether there are an even or odd number of bits set to 1 in the character. This is a very simple method of error detection — the number of 1 bits in the character are counted and if the count and the parity bit do not agree, there was an error in that character. ASCII (and the extended 16-bit Unicode) character sets are used in digital communication that is not conducted via RTTY or PSK31.

Symbols and Bits

One of the more important measurements or specifications of a digital communication channel is its *bit rate*. In Morse code, RTTY, and PSK31, during the period when one bit is sent you would hear the pattern or tone for a 1 or a 0. The reciprocal of that period is the bit rate, given in bit/s or bps. For example, if the period for one bit is 1 msec, the channel's bit rate is the reciprocal of 1 / 1 msec = 1000 bps. Take care not to confuse bit/s with byte/s!

Some methods of encoding data into transmittable (and receivable) signals are not restricted to sending only one bit at a time. For example, if the transmitted signal consisted of two tones, each corresponding to a 1 or a 0, then during each bit period four different combinations of two bits could be sent: 00, 01, 10, or 11. Each of these combinations is called a *symbol* and the rate at which different combinations are sent is called the *symbol rate*. In this example, if the symbol rate was 300 symbols per second, the bit rate would be 2 × 300 = 600 bps.

In honor of Baudot, symbol rate is measured in units of *baud* with 1 baud = 1 symbol / second. Thus, one does not say "baud rate," because baud is already a rate — there is no need to say "symbol rate rate." Just "baud" or "bauds" will do.

Packaging more than one bit into a symbol is common, and amateurs have been experimenting with protocols that use up to 63 different tones (MT63) to send an entire character during one symbol period. This speeds up data transmission at the expense of a wider signal bandwidth and requiring more sophisticated data coding and decoding equipment and software.

One of the more popular modulation methods to transmit symbols representing multiple bits is called *quadrature amplitude modulation* or *QAM*. In this method, two carriers are transmitted with a 90° phase shift between them. One is called the *I signal* and the other is the *Q signal*. They are turned on and off in various combinations to represent 00, 01, 10, and 11.

[1]

The four combinations are plotted on a graph shown in Figure 3, which is called a *constellation display*. (QAM and the similar QPSK are well-suited to digital signal processing techniques. Advanced modulation schemes having up to 256 different symbols have been used!)

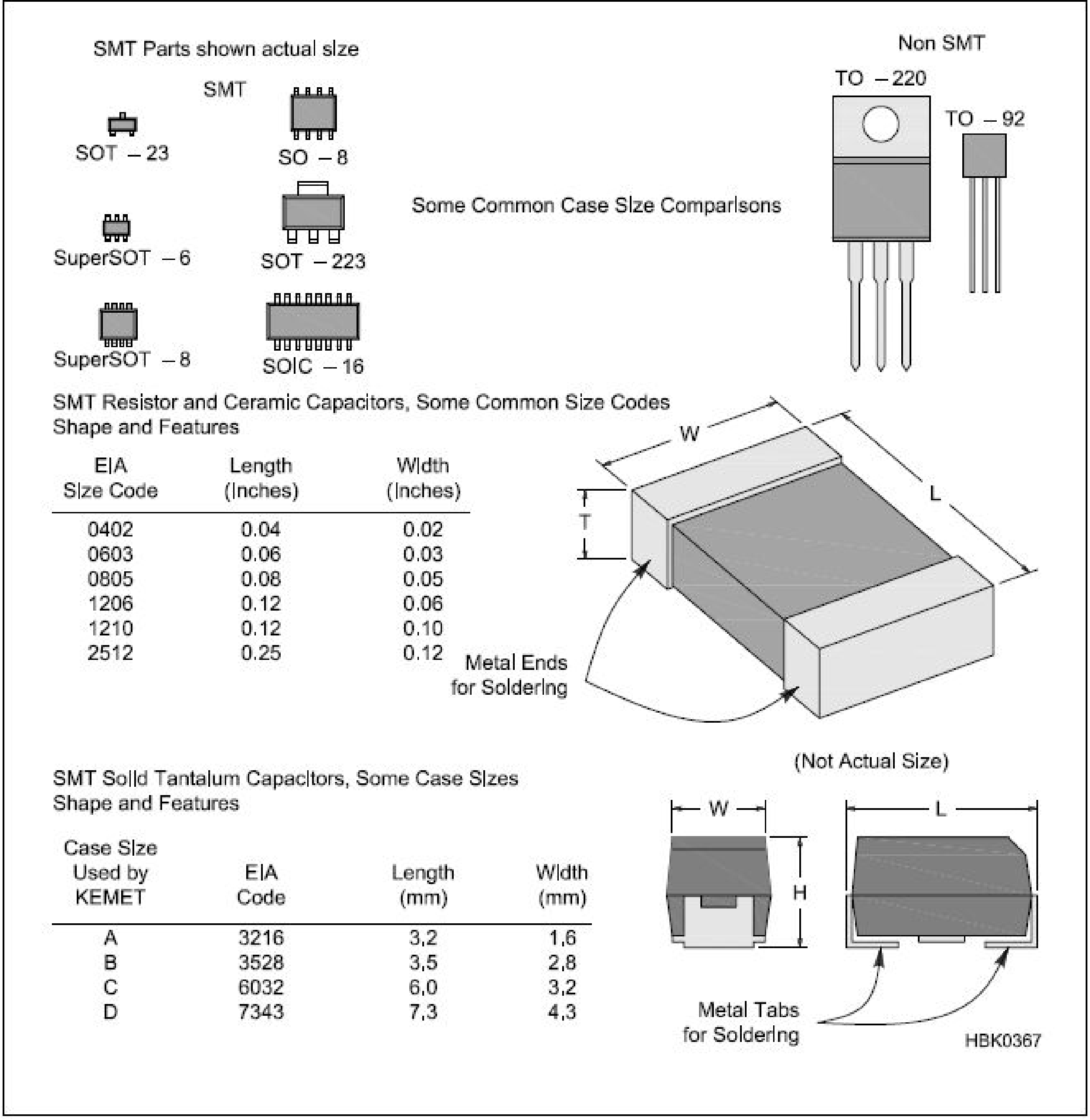

Figure 3 — The constellation display of an I-Q signal over an extended period. The points at which the lines cross in the corners represent the individual symbols representing 00, 01, 10, and 11.

Figure 3 shows how the received signal moves from point to point in the constellation. When the signal reaches any of the four points, the receiving decoder generates the combination of 1s and 0s represented by that point. Variations in the trajectories between each of the four corner points are caused by noise and non-linearities in modulation circuits. The "fuzzier" the constellation display, the harder it is for the receiving decoder to make correct decisions about which symbol was actually transmitted.

In order to help the receiver operate better under adverse conditions, further coding schemes such as *Viterbi encoding* place restrictions on which codes can be transmitted in sequence. This reduces the number of possibilities the receiving decoder must consider and so makes it easier to reject noise and distortion.

Finally we get to what really matters to a communication system user — *data transfer rate*. Measured in bytes / second, data transfer rate describes the ability of the entire system to move data from end to end and includes the slowing-down effect of the extra bits sent with each character, characters added to create packets or other structures, protocol timing delays, and so forth. It also includes the speeding-up effect of multiple bits per symbol. Generally speaking, data transfer rates in most systems are anywhere from 1⁄2 to 1⁄10 of the system's bit rate.

Fielding a Recommendation

Readers of Hands-On Radio may remember a few columns that addressed the early experiments establishing the link between electricity and magnetism.

[2]

Recently, I discovered a terrific book that takes the subject further — to the electromagnetic field. *Faraday, Maxwell, and the Electro- magnetic Field: How Two Men Revolu-tionized Physics*, by Nancy Forbes and Basil Mahon explains very well how the idea of a "field" grew out of Faraday's discoveries and suggestions to be given its mathematical description by Maxwell. Not only is it an interesting tale of technical history but using minimal math, it provides some good background on what a field is — a hard thing to understand clearly as one begins learning about radio.

Notes

1

I-Q modulation was discussed in Experiment 128. All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

Experiments 117 and 118, "Laying Down the Laws."

Experiment #154 — Power Factor and Phase Angle

Someone just learning electronics first learns Ohm's Law relating voltage, current, and resistance: $R = E / I$. The next step is usually to learn how to calculate power: $P = E \times I$, but this equation must be qualified by the term, "in a resistive circuit." Inductive and capacitive circuits are somehow different, we learn. Why would inductance and capacitance have anything to do with power?

A similar conversation takes place when the terms "reactance" and "impedance" are introduced. Understanding resistance is fairly intuitive — it's like electrical friction — but this reactance stuff can seem a little odd at first: "Okay, capacitive reactance is like a spring and inductive reactance is like a flywheel. I think I get that. But energy is energy, isn't it? Why does this affect power, which is $E \times I$? What is power factor? Phase angle?"

Understanding why inductance and capacitance and their abilities to store energy affect the relationship between power and energy can feel like deep water. In this column, we'll explore these electrical fundamentals.

Power at Work

Before jumping in, let's start from solid ground — power and energy in a resistor. When voltage is applied to a resistor, electrons begin to move (i.e. *current*) through whatever material the resistor is made of; carbon or metal film, nickel-chromium (nichrome) wire, a chunk of metal oxide, etc. The actual speed of the electrons is surprisingly low [1] but there are an incredible number of them, even in the thinnest conductor. Each electron collides with the atoms making up the conductor and transfers some of its energy to the atom, causing it to vibrate, which is heat. Because the electron moves through the material in response to the applied voltage, the source of the *electromotive force* is said to have done *work* in the physics sense. [2] In this case, the work is heating up the material.

Work (W) is measured in the same units as energy (joules) and can be thought of as "expended energy." The rate at which work is performed per unit of time (t) is power: $P = W / t$ and is measured in watts if t is measured in seconds. One watt (also abbreviated W) is equal to one joule of work done (or energy expended) per second. Higher power means more work has been done in a specific period of time or that a specific amount of work done in a shorter period of time. What is important is that the electron moves through the material at the same time the voltage is applied.

Power and Phase

If the voltage applied to a resistor is a steady value, the resulting *direct current* (*dc*) will be a steady value, too. At each moment of time, you can calculate the *instantaneous power* by multiplying voltage and current and that, too, will be a steady value. If the applied voltage periodically reverses, creating *alternating current* (*ac*), the instantaneous power dissipated by the resistor as heat will also vary.

Figure 1 shows voltage, current, and instantaneous power in an 8 Ω resistor when a 4 V sine wave is applied. [3] It is important to note that when an ac voltage is applied to a resistor, voltage and current have exactly the same phase; reaching zero, maximum, and minimum at the same time. Instantaneous power ($E \times I$) in this case is always positive even when both voltage and current are negative to indicate reversal.

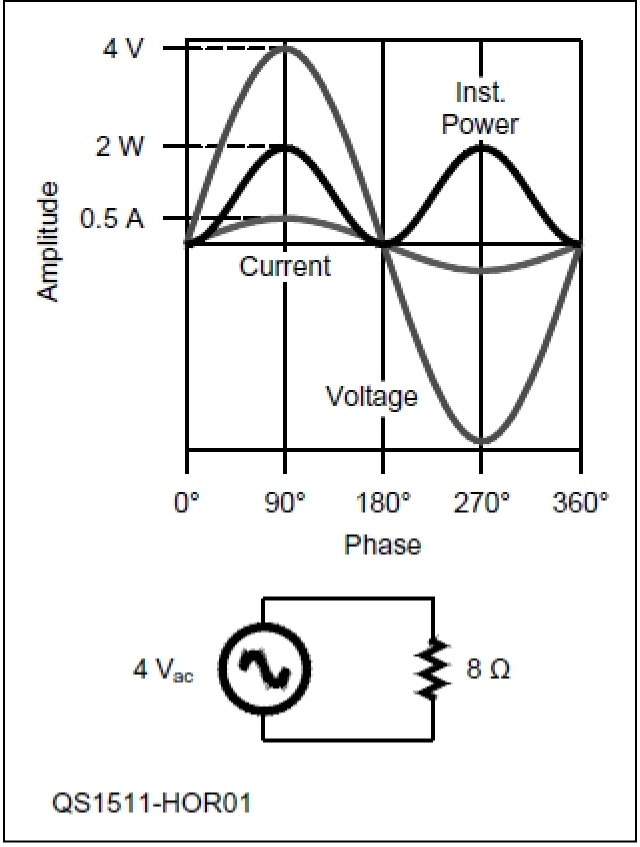

Figure 1 — Voltage (red), current (blue), and instantaneous power (black) in a circuit composed of a 4 Vac source and an 8 Ω resistor.

What happens if an ac voltage and current are not precisely in phase? First, why wouldn't they be precisely in phase? Well, if the component in the circuit stores and returns some of the energy instead of dissipating it as heat, that alters the timing between the current and voltage waveforms. For example, if an ac voltage is applied to the *capacitive circuit* of Figure 2A, the resulting current leads the applied voltage by 90°. (For an explanation, see the "Electrical Fundamentals" chapter of *The ARRL Handbook* or the section "Reactance and Impedance" in the *General Class License Manual*.[4, 5, 6])

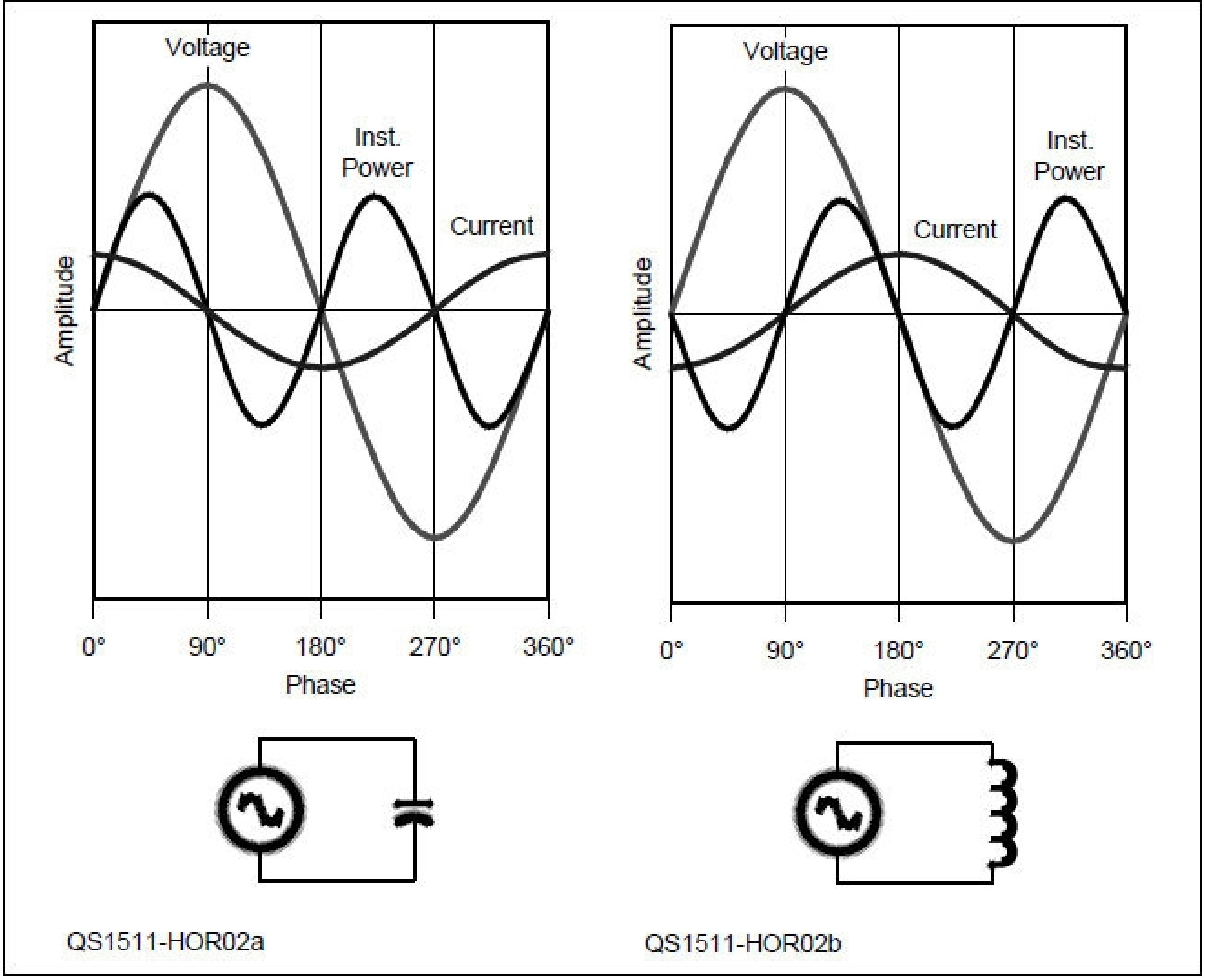

Figure 2 — Voltage (red), current (blue), and instantaneous power (black) in the series circuit of an ac voltage source and a capacitor (A) or an inductor (B).

Instead of instantaneous power always being positive as in the resistive circuit, it is positive for half the cycle (0° to 90° and 180° to 270°) and negative for the remaining half-cycle. When we add up instantaneous power throughout the cycle, the result is zero and no net work has been done at all! Figure 2B shows the complementary situation for an *inductive circuit* — the net result is the same. If total work is zero, then total power is also zero and no energy has been consumed.

Minding the Ps and Qs

"But, but, but…" I hear you exclaim, "sure, total power is zero over the whole cycle but during half the cycle, power is being consumed! What happened to that power?" An excellent question! During the positive power half-cycle, energy is not being consumed or dissipated, it is being stored in an electric field (for a capacitor) or in a magnetic field (for an inductor). During the negative-power half-cycles, energy is returned to the source.

Power for which the voltage and current are in phase is called *real power* because the power does "real work" and is not stored or returned. Real power is labeled *P* and is measured in watts. Power for which the voltage and current are 90° out of phase (see Figure 2) is called *reactive power* because of the reactance creating the phase shift. Reactive power is labeled *Q* and is measured in *volt-amperes reactive* or *VAR*.

If a circuit contains both resistance and reactance, also called a *reactive load*, the resulting instantaneous power waveform is made up of both real and reactive power. Figure 3 uses complex numbers to show the relationship between real and reactive power in such a circuit. Q is drawn parallel to +90° imaginary axis if the reactance is inductive (as in Figure 3) or at –90° if the reactance is capacitive.

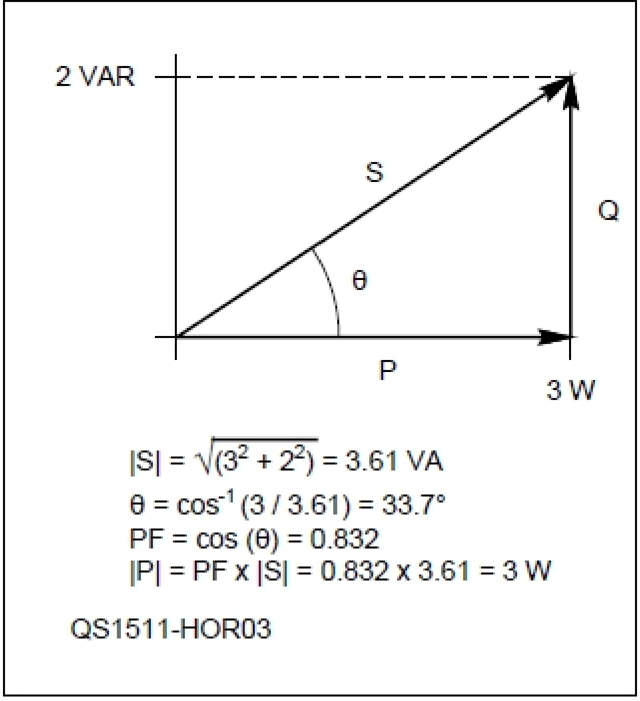

Figure 3 — Complex numbers representing 3 W of real power (P), 2 VAR of reactive power (Q), and the resulting complex power (S) for a circuit with both resistance and inductance.

The addition of the values for real (P) and reactive power (Q) results in the complex number $S = P + jQ$, representing *complex power.* The magnitude of complex power, $|S| = \sqrt{P^2 + Q^2}$ is called *apparent power*, which is measured in *volt-amperes* or *VA*. Apparent power is also equal to the magnitude of the voltage and current measured separately and multiplied together without regard for phase differences.

Power Factor and Phase Angle

We've finally arrived in the deep water! The *phase angle*, θ, from *P* to *S* is the amount of phase difference between the applied voltage and the resulting current in the circuit. *Power factor, PF,* is the cosine of the phase angle, θ, and so is always in the range of 0 to 1.

A PF of 1 means the voltage and current are exactly in phase and all power is real power. PF becomes gradually smaller as Q increases, eventually reaching 0 when all of the power is Q with no P. It doesn't matter whether Q is inductive ($\theta > 0$) or capacitive ($\theta < 0$) because the cosine of θ is the same for positive and negative angles.

Power system engineers use PF to specify how much reactive power is present in a particular circuit and they prefer PF to be close to 1. Why? If PF gets smaller for a given amount of real power, P, that means reactive power, Q, must be increasing. Oversimplifying somewhat, that means more volts and amps are dedicated to reactive power and not doing any real work. Those reactive volts and amps stress insulation and cause heating from I^2R losses so wires have to be bigger and insulation thicker. Keeping PF close to 1 minimizes the bad effects of Q while the power system delivers the required amount of P to customers.

This might be starting to sound familiar to hams. RF designers use θ when calculating impedances and designing matching circuits to make your transmitter happy with a $50 + j\,0\,\Omega$ impedance — just another way of saying PF = 1 and $\theta = 0°$! So you see, power factor and phase angle are just another way of discussing the relationship between voltage and current in an ac circuit, whether at 60 Hz or in our RF bands.

Notes

1

For typical currents in copper, the progress of a single electron through a wire (*drift velocity*) is much less than 1 mm/sec (
hyperphysics.phy-astr.gsu.edu/hbase/electric/ohmmic.html#c2
).

2

There are various definitions of work depending on the system involved. The precise definition of electrical work is discussed at
www.physicsclassroom.com/calcpad/energy
.

3

Unchanging voltage is referred to as a “dc voltage” and a regularly-reversing voltage as an “ac voltage” even if current is zero.

4

The ARRL Handbook, 93rd edition, ARRL, 2015.

5

General Class License Manual, 8th edition, ARRL, 2014.

6

All previous “Hands-On Radio” columns are available to ARRL members at **www.arrl.org/hands-on-radio**.

Experiment #170 — Noise Figure

On our MF and lower HF bands, the received signal-to-noise ratio (SNR) is dominated by galactic noise from 15 to 30 MHz at night by atmospheric noise from storms and man-made sources below 15 MHz.

The type of noise we are mostly concerned with in sensitive RF receiving systems is *thermal noise*, also called *Johnson-Nyquist noise*. There are many other types of noise — shot noise, flicker noise, and even popcorn noise — that are generated inside our electronic devices. From the external world, atmospheric noise is accompanied at VHF and above by galactic noise and sun noise.

Thermal noise is generated in all conducting materials by the random vibration of free electrons due to their thermal energy. The higher the material's temperature, the larger the vibrations of the free electrons become. This motion of charge creates a *noise voltage* in any conductor not at absolute zero.

Based on material [1] by Paul Wade, W1GHZ, in *The ARRL Handbook*, the basics of thermal noise are as follows: Every resistance (and all conductors have resistance) generates a root-mean-square (RMS) noise voltage:

$e = \sqrt{4kTRB}$

where k is Boltzmann's constant (1.38×10^{-23} joules / K), T is the absolute temperature in kelvin (K, which is equal to the temperature in Celsius plus 273), R is the resistance, and B is the bandwidth in hertz. Converting to power, $e^2 / 4R$, the noise power generated by the resistor is:

$P_n = kTB$ (watts)

This independence of frequency is why thermal noise is called *white noise* and has a power that depends only on temperature.

Note that all resistances at the same temperature generate the same noise power. Similarly, if two noise sources generate the same power in the same bandwidth, they are said to have the same *noise temperature*, T_n. This is the temperature at which a resistor at the same temperature would generate the same noise power as the source, whether it is an electronic circuit, a cable, or an antenna.

The amount of power per unit of bandwidth is called *power density* or *spectral density* and is equal to kT watts per Hz. Because power is proportional to the square of voltage, the corresponding *voltage density* or *spectral density* is measured in volts / √Hz, spoken as "volts per root hertz."

Calculating noise power density at 290 K (290 K = 16.9°C = 62.6°F) gives:

$P_n = (1.38 \times 10^{-23} \times 290)\ B = 400 \times 10^{-23}\ B$

Multiply by the bandwidth in hertz to get the available noise power at 290 K. The choice of 290 K is simply for convenient calculations because P_n is 400 B at that temperature. Converting to dBm by calculating 10 log(P_n), we get –174 dBm / Hz.

Measuring Noise with Noise Figure

Continuing to build on W1GHZ's material in *The ARRL Handbook*, all amplifiers add additional noise to the noise present at their input. The input noise per unit of bandwidth is $N_i = kT_g$, where T_g is the noise temperature at the amplifier's input. Amplified by power gain, G, the output noise power is kT_gG. The noise power added by the amplifier, kT_n, is then added to the amplified input noise to produce a total output noise, $N_o = kT_gG + kT_n = k(T_gG + T_n)$.

We can model the amplifier as ideal and noise-free and add a noise-generating resistor of temperature $T_e = T_n / G$ at the input. In this way, all sources of noise can be treated as inputs to the amplifier, as illustrated by Figure 1. Substituting for T_n in the previous equation, the output noise is then $N_o = kG\,(T_g + T_e)$.

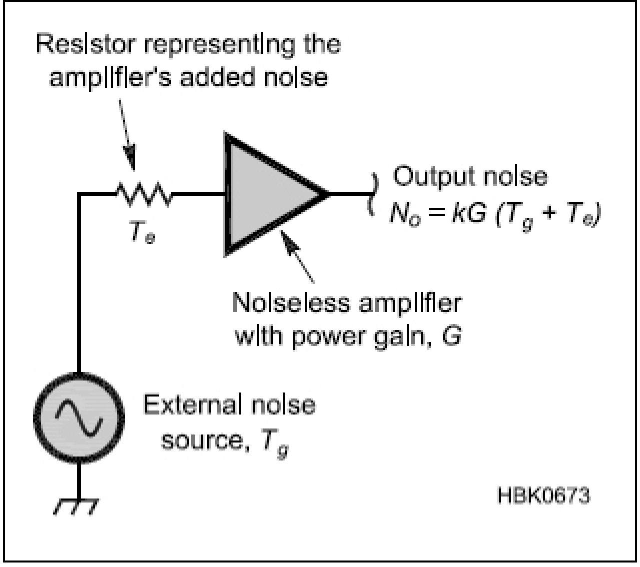

Figure 1 — The noise generated by an amplifier can be represented as an external resistor with a noise temperature of T_e connected at the input of an ideal, noise-free amplifier.

The noise added by an amplifier can then be represented as kGT_e, which is the amplifier's noise temperature amplified by the amplifier's gain. T_e is sometimes referred to as *excess temperature*.

Because amplifiers have many different values of gain, there needs to be a way to compare their noise performance without a bunch of calculations. The answer is *noise factor* and *noise figure*.

The noise factor F of an amplifier is the ratio of the total noise output of an amplifier with an input T_g of 290 K to the noise output of an equivalent noise-free amplifier. The easiest way to do this is to use noise temperatures: $F = 1 + T_e / T_g$. It is often more convenient to work with *noise figure*, *NF*, the logarithm of noise factor expressed in dB:

$NF = 10 \log F$ and $F = \log^{-1}(NF / 10)$

If the signal-to-noise ratio (SNR) in decibels is known at the input and output:

$NF = SNR_{in} - SNR_{out}$ or $SNR_{out} = SNR_{in} - NF$

Noise figure is sometimes stated as *input noise figure* to emphasize that all noise sources and noise contributions are converted to an equivalent set of noise sources at the input of a noiseless device. In this way, noise performance can be compared on equal terms across a wide variety of devices. It also makes comparing their relative contributions to output noise much easier.

Location, Location, Location

When more than one noise-generating component is connected in series or cascade, they all contribute differently to the overall output noise level and SNR. Even a simple receiving system consists of three such components — antenna, feed line, and receiver — with each adding to, or even amplifying, the noise at its input. The *Friis equation* takes into account the gain or loss of each component and gives an overall *system noise factor* for N devices connected in series.

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots + \frac{F_N - 1}{G_1 G_2 \cdots G_N}$$

where device 1 is the one at the input to the system. Clearly, if the gain of the first stage, G_1, is large, then the noise contributions of the succeeding stages become too small to be significant. In addition, the noise temperature of the first stage is the largest contributor to the overall system noise because it is amplified by all remaining stages.

The important thing to learn from the Friis equation is that noise-reduction efforts should be made at the input to the system. A good example is the question of where to put a preamp in an antenna system. Figure 2 shows the difference between placing the preamp at the input to a lossy feed line and placing it at the receiver input.

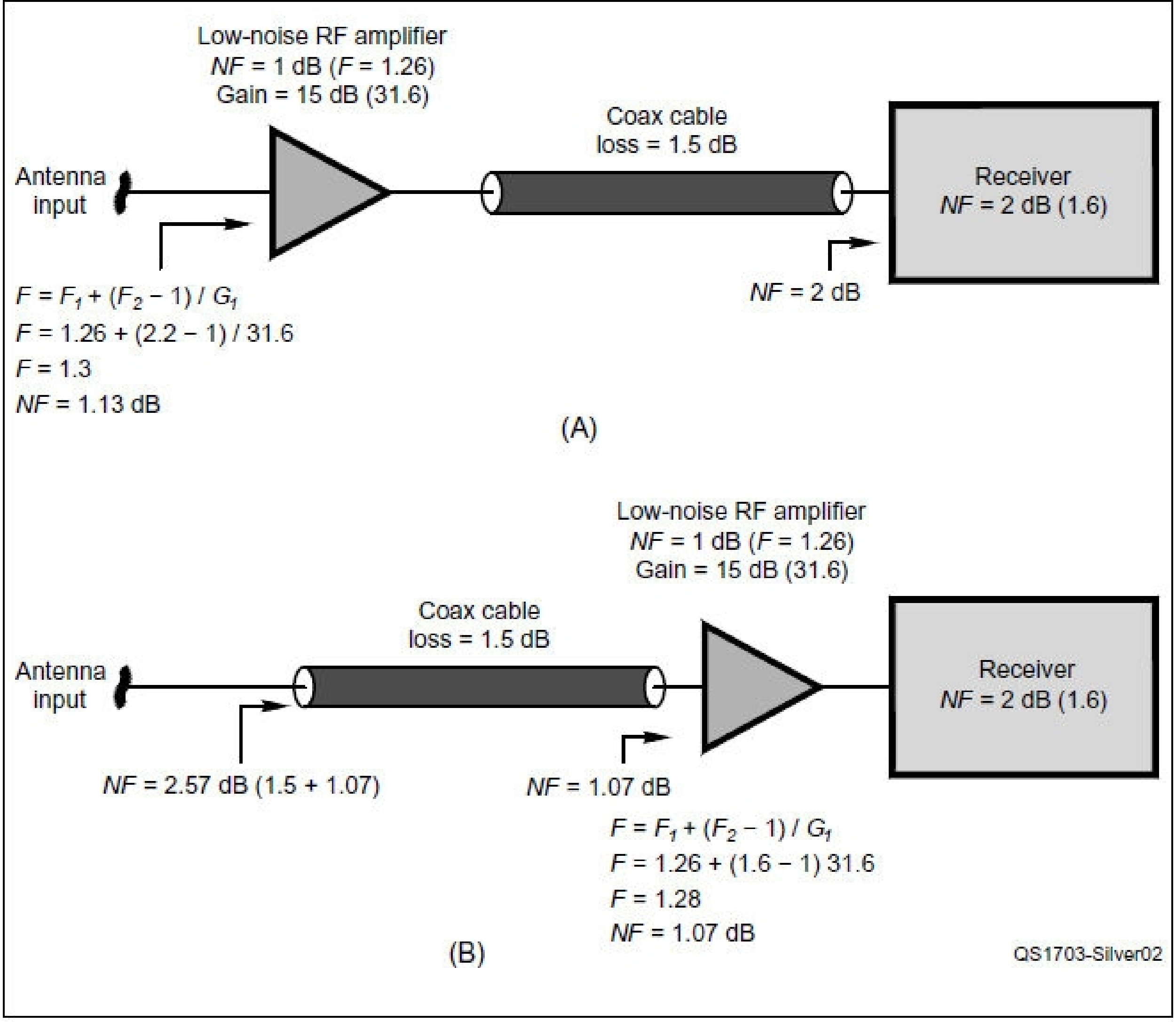

Figure 2 — The difference in the system noise figure between placing a high-gain RF preamplifier at the antenna (A) versus at the receiving input (B). In this example, moving the preamp to the feed-line input reduces *NF* at the system input from 2.57 dB to 1.13 dB, a significant improvement. The arrows and associated values indicate the point at which noise factor and noise figure are calculated.

You can hear the Friis equation in action with a simple experiment using a VHF or UHF FM transceiver, a long length of coax or an attenuator, and a preamplifier. [2], [3], [4]

Choose some coax with about 3 – 6 dB of loss — 100 feet of RG-58 or RG-8X will do nicely at VHF or UHF. Connect the coax between a small antenna and your transceiver, then tune in a weak (distant) repeater, which is received with a noisy signal at your location. (If you use an attenuator, just add attenuation until a local repeater signal becomes noisy.) Add the preamp at the antenna and observe how much the SNR improves. Then move it to the other end of the coax, connecting it directly to the transceiver input, and observe SNR again. You will find that placing the high-gain preamp closer to the input of your receiving system results in a better SNR.

Notes

1 Available from your ARRL dealer, or from the ARRL Store, ARRL Item no. 0628. Telephone toll-free in the US 888-277-5289 or 860-594-0355, fax 860-594-0303, **www.arrl.org/shop**, **pubsales@arrl.org**.

2

m2inc.com

3

amsat-uk.org/info/g0mrf-144-mhz-preamp-kit

4

www.w1ghz.org/small_proj/Simple_Cheap_MMIC_Preamps.pdf

Experiment #175 — Dissimilar Metals

When putting up my station with tower maven Don Daso, K4ZA, I got an earful of good advice about what happens when different types of metal are clamped together out in the weather. Hint — nothing good! Don gave me some good tips on how to manage the situation at reasonable cost.
[1]

Because metal-to-metal connections are so common in amateur antenna and ground systems, I thought it was a great candidate for a "Hands-On Radio" column.

Electrochemistry at Work

Atoms that have an affinity for electrons are called *oxidizing agents.* Similarly, atoms that donate electrons are called *reducing agents.* There is a whole range of strengths for oxidizing and reducing agents. The greater the difference between the material's relative affinity for electrons, the stronger the reaction between them can be, with atoms of one material acquiring electrons from the donating material in an *oxidation-reduction reaction.* The relative difference in strength is called *electropotential*.

If two materials with different electropotentials are in contact with each other, or there is a conducting path between them, electrons will move from the reducing to the oxidizing agent. This results in a change in the chemical makeup of the two materials. When the reaction occurs between parts of our antenna system, we apply a more descriptive word — *corrosion*. Corrosion is generally bad because the result of the reaction is generally a compound with less strength and conductivity than the original metal. Rust is one such corrosion product and so is that crusty stuff that builds up on your car battery terminal. Corrosion can proceed until the connection no longer conducts current or fails mechanically, i.e. the wire, antenna, or tower falls down.

Because not everything is made from the same metal, contacts between *dissimilar metals* are very common. If the metals are kept dry, corrosion proceeds slowly. But outside, the water from rain and condensation collects in the junction. This creates the conducting path, enabling electrons to flow and cause corrosion.

The Galvanic Series and Anodic Index

It is useful to know which materials can "get along" and which cannot. Chemists developed a list called the *galvanic series*, ranking materials from the strongest electron donors to the most inert material. The farther apart the two materials are in the series, the stronger the reaction between them will be. The more active material will act as the *cathode*, which donates the electrons. (Remember that chemistry is based on electronic current — the flow of electrons — and not conventional current, which is the flow of positive charge that is used in radio electronics. Confusing? Yes.)

Electropotential can be measured with a voltmeter. Measured in volts, it is what Alessandro Volta discovered and made use of to create the *voltaic pile*, known today as a battery. Because electropotential is a relative difference, it is measured as a voltage with respect to some reference material. That voltage is a metal's *anodic index* with respect to gold as the reference material. Table 1 shows the anodic series for materials common in our stations.

Because no two metals have the same anodic index value, won't there always be some corrosion when there is contact between dissimilar metals? Yes, but very small differences in anodic index result in very slow corrosion. If the environment is harsh, such as exposed to the weather or salt spray, limiting the difference in anodic index to 0.15 V or less means the corrosion will be manageable. In normal conditions, up to 0.25 V difference can be tolerated. (This is discussed in more detail at **corrosion-doctors.org/ Definitions/galvanic-series.htm**.)

Connecting to Galvanized Surfaces

Because zinc is an extremely active material (which is why it is so effective at protecting steel and iron), connecting copper wire or bronze clamps directly to a galvanized surface creates a very strong corrosion potential. What should you do?

Although there are anti-corrosion compounds, continuous exposure to weather will eventually wash them away. Instead, an inexpensive method of protecting the connection is to place a thin piece of 300-series stainless steel between the galvanized surface and the clamp or wire. 24-gauge (0.025-inch) shim stock is sufficient and acts to slow down the corrosion process while maintaining good electrical connectivity for lightning protection and RF connections.
[2]

Figure 1 shows a photo of a ground clamp on a galvanized tower leg with a shim between the clamp and the leg. Thin shim can be cut with heavy scissors and bent by hand. Another option is to use a tower leg clamp designed specifically for this application, such as the Rohn R-CPC1/1.25 (available from several vendors).

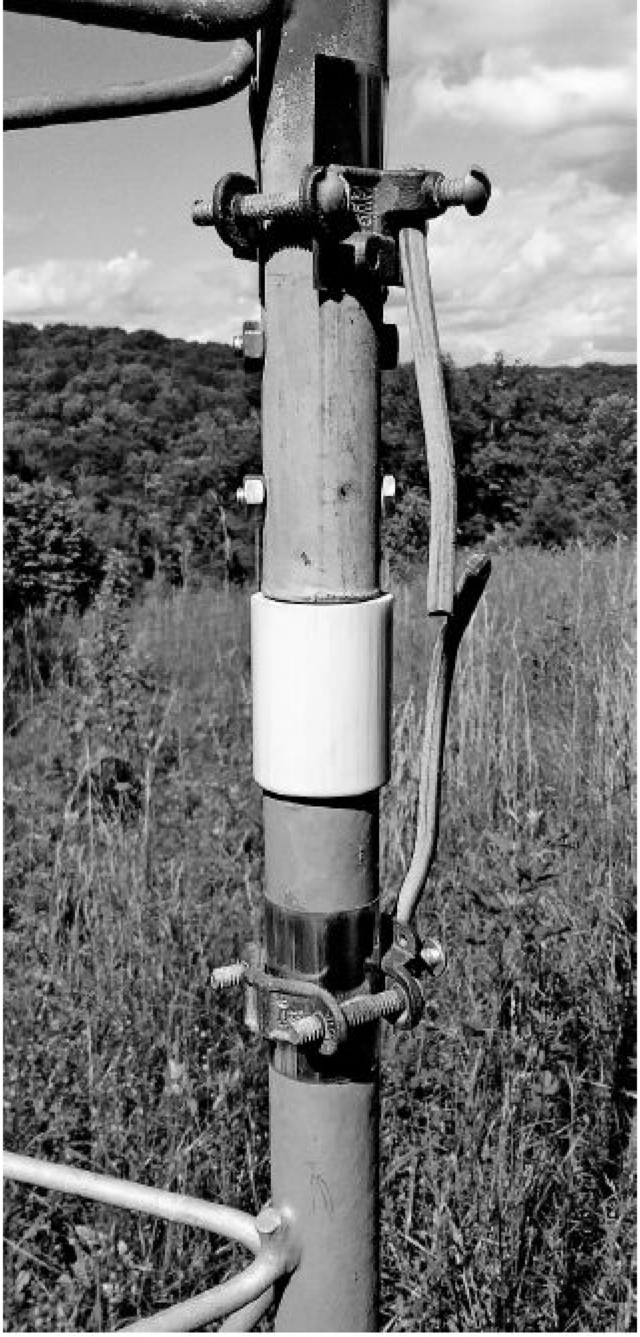

Figure 1 — A 300-series stainless-steel shim between a bronze ground clamp and galvanized tower leg protects the zinc layer and the clamp.

Some areas have corrosive soils and other environmental characteristics that make corrosion an ongoing issue with towers. Tony Fisher's, K1KP, article in the October 2010 issue of *QST* — "Is Your Tower Still Safe?" — discusses setting up a *sacrificial anode* system that will protect your tower from corrosion. You can also check with local contractors to see if corrosion is a special problem in your area.

Making Your Own Galvanic Cell

This is all very academic until you see it for yourself. The easy and non-destructive way is to make your own *galvanic cell* and take some measurements.

Here's what you need: a plastic or glass jar (12 – 16 oz. capacity), a tablespoon of kitchen salt, a multimeter, two clip leads, and some test metals. I used common materials: copper, aluminum, stainless steel, galvanized steel, and 63/37 lead-tin solder. Any metal you can carry in the palm of your hand is safe to test, including coins and household metals. Remove any oxidation or grease from the surface, so that it's shiny and clean.

Dissolve the salt in enough water to fill about 3⁄4 of the jar. Set the multimeter to measure voltage (the 2 V scale will work best) and attach clip leads to the probes. (This keeps your probes from getting salt water on them.) Set the jar on an absorbent towel (not a good one!) and put the copper wire in the water with one end held out of the water by the clip lead or jar lip. (Keep the clip lead out of the water to keep its surface clean, too.) Now attach the other clip lead to one of the remaining pieces of metal placed in the water, but not touching the copper wire. You can see the experimental setup in Figure 2. The voltage shown on the meter will be approximately the difference in anodic potential between the two metals in Table 1. (The exact voltage will probably vary from Table 1 values because the metal may be an alloy or there may be surface oxidation that alters the chemistry a bit.)

Table 1
Anodic Index

Metal	Index (V)
Gold, solid and plated	0
Silver, solid or plated	–0.15
Nickel, solid or plated	–0.3
Copper, solid or plated; silver solder	–0.35
Brass and bronzes	–0.4
18% chromium-type stainless steels (Type 304)	–0.5
Tin-plate; tin-lead solder	–0.65
Iron, wrought, gray or malleable, plain carbon and low alloy steels	–0.85
Aluminum, wrought alloys (other than 2000 series)	–0.9
Hot-dip-zinc plate; galvanized steel	–1.2
Zinc, wrought; zinc-base die-casting alloys; zinc plated	–1.25

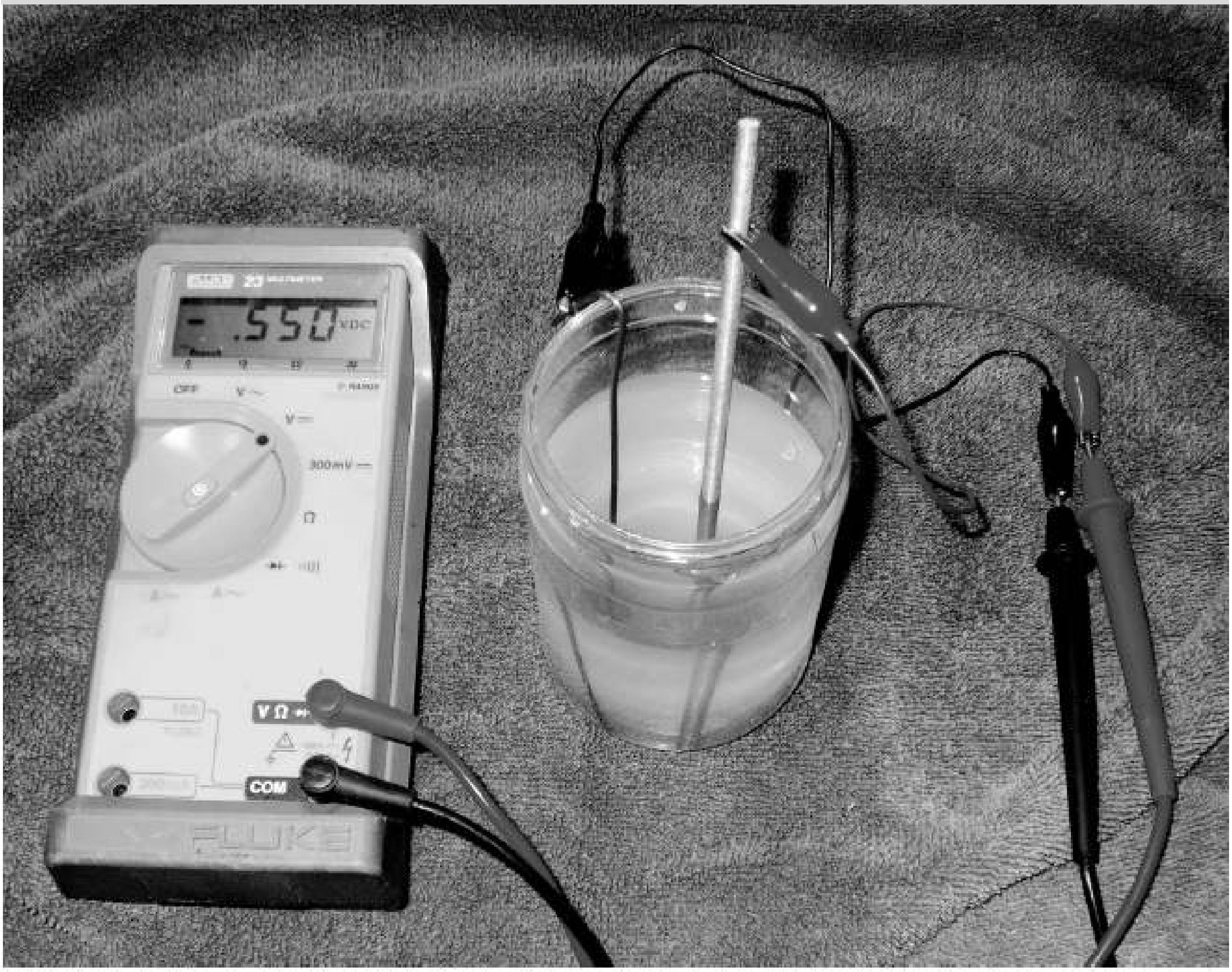

Figure 2 — A salt-water galvanic cell showing a 0.550 V difference in electropotential between a copper (Cu) wire (left) and aluminum (Al) rod. For Cu-Stainless steel, the voltage was 0.140 V. From Cu-Zn (a galvanized nail), the voltage was 0.805 V. From Cu-Pb/Sn 63-37 solder, the voltage was 0.234 V.

Notes

1

D. Daso, K4ZA, *Antenna Towers for Radio Amateurs*, Appendix B, ARRL, 2010.

2

https://www.galvanizeit.org/images/uploads/drGalv/Stainless_Steel_in_Contact_with_Galvanized_Steel.pdf

Experiment #178 — Maxwell's Equations — Grad, Div, Curl

Given their fundamental nature, it is natural to think of Maxwell's equations as describing laws of nature. They do, but it was not Maxwell who discovered them. As we learned in Experiments #117 and 118, those insights came from Faraday, Gauss, and Ampere.
[1]

What Maxwell contributed was to see the relationships of electric and magnetic fields, voltage and current, as different components of a single natural phenomenon — *electromagnetism*. He simplified the relationships down to the level of "first principles" that aren't derived from any other more fundamental ideas — these are bedrock descriptions of the universe.

In the process, he came to understand that electric charge and magnetism are deeply related. After Maxwell, it became apparent that they are just different ways in which electromagnetic energy interacts with matter as it moves through space. An electron (or any electric charge) can respond to electric fields or magnetic fields. The same electron generates an electric field or, if in motion, a magnetic field.

The key is "motion." If anything could be considered "Maxwell's Law" it would be his modification to Ampere's Law. He made that equation symmetrical with Faraday's Law so that a time-changing magnetic field was linked to an induced electromotive force or voltage. This was the true spark of genius (so to speak), going beyond Faraday's Law to suggest that a time-changing electric field can produce a current.

Maxwell then made a second leap from "time-changing" to "moving," and got a wave. (Faraday had suggested there might be waves associated with induction, but hadn't incorporated electric fields into the idea.) As they say, this changed everything. As Einstein himself acknowledged, this equivalence of motion and changes with time lies at the heart of relativity and its equivalence of space and time. (Pretty deep stuff for a ham radio magazine, I must say.)

Maxwell's original 20 equations were difficult to understand. We owe their current form to Oliver Heaviside, who simplified them to four equations using modern notation in 1885.
[2], [3]

Table 1 explains as text what each of Maxwell's equations says and then states them mathematically. (Rautio's explanations are more detailed and the equations shown here are somewhat over-simplified.) This is the mountain. The following sections cover three basic math concepts that are used to express the equations. These are your climbing gear.

Table 1
Maxwell's Equations as Text and (Over)Simplified Math

Equation 1	Gauss's Law: electric charge in some volume, q_V, generates an electric field,	$E\ \nabla \bullet E = q_V/\varepsilon_0$
Equation 2	There is no independent magnetic "charge" analogous to electric charge	$\nabla \bullet H/\mu_0 = 0$
Equation 3	Faraday's Law: a changing magnetic field, H, creates an electric field,	$E\ \nabla \times E = -\mu\ (\delta H/\delta t)$
Equation 4	Ampere's law: moving charge (current, I) and changing electric fields, E, can both create a magnetic field, H.	$\nabla \times H = I + \varepsilon(\mathrm{d}E/\mathrm{d}t)$

Gradient (⍰)

Let's start with *gradient*, which is pretty easy to visualize. Gradient, represented by the ⍰ or *nabla* symbol, is a measure of how fast some parameter changes with respect to some other parameter. For example, a steep hill has a large gradient of height with respect to distance. (The word "grade" springs from the same root as "gradient.")

In ham radio, one gradient that concerns us is a change in voltage over distance. An insulator's breakdown voltage — the maximum gradient the insulator can withstand — is expressed in volts per inch. This is why high-voltage circuits and components have rounded or smooth edges and points — to reduce the voltage gradient near these surfaces, avoiding arcs and corona.

Figure 1 is an illustration of gradient on a topographic map, which measures *gravitational potential*, also known as "elevation." The two blue lines, A and B, each represent a horizontal distance of 2,000 feet. Along which line is the gradient of elevation (vertical distance per horizontal distance) the greatest from end to end? The heavy brown contour lines are spaced 100 feet apart, so the net gradient along A from end to end is 200 feet in 2,000 feet or 0.1 feet per foot. B touches six heavy brown lines for a gradient of 600 / 2,000 = 0.3 feet per foot. The gradient symbol in Maxwell's equations includes the gradient in all three dimensions, not just two as in this map.

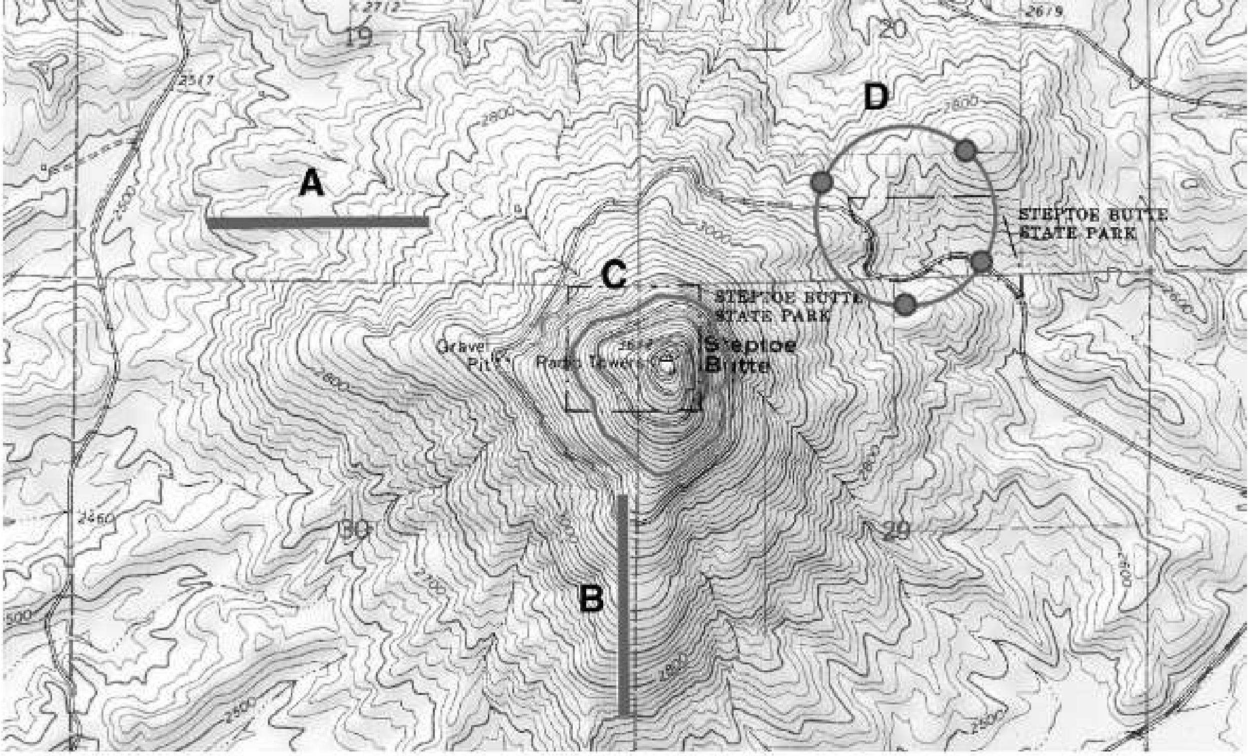

Figure 1 — For this example, we will assume the contour lines representing constant elevations on this topographic map of Steptoe Butte in eastern Washington also represent lines of equal gravitational potential. Gradient is represented by the spacing between the lines. Divergence along a line is indicated by the direction of uphill or downhill across the line at that point. [Map provided by Google, Inc. and **Digital-Topo-Maps.com**]

Divergence (•)

Divergence, represented by the • symbol, can also be illustrated on a topographic map. Divergence describes whether a value, such as gravitational potential (elevation) or electrical potential (voltage), is increasing or decreasing through a curve or across a surface. Start with the contour circuit C, which surrounds the central peak, and imagine a rolling ball as your "gravity-o-meter." What would a ball do if placed on the contour circuit line? Everywhere around circuit C, gravitational potential increases to the "inside" and decreases "outside," so the ball would roll away from the central peak. We would say there is a high positive divergence in gravitational potential (elevation) across the circuit. (If the circuit was drawn around a sinkhole, there would be a high negative divergence.)

The case of the circuit labeled D is not as simple. Part of the circle is on the slope of a nearby peak, two parts are in separate parts of a valley, and some is on the slopes of the central peak. Depending on location, a ball dropped on the circuit would roll toward the interior (1 and 6 o'clock positions, negative divergence), away from the interior (4 o'clock, positive divergence), or along the circuit (10 o'clock, zero divergence). Figuring out whether net divergence was positive or negative would require you to sum it up at each point around the circuit. Mathematically, this is an integration around the whole circuit, and it is shown as an integration symbol with a small circle in the middle (see Rautio's website in Note 3). Like the gradient, divergence in Maxwell's equations includes all three dimensions.

Curl (×)

The final tool in the set is *curl*, and that is something we can't show on a topographic map. Curl is derived from *circulation*, which could be understood as the push from gravity along a closed path such as one of our contour circles. Curl is denoted by the × symbol, which is also used to represent the mathematical *cross product* of two vectors.

Along circuit C in Figure 1, you would get no push anywhere because the gravitational potential at each point (elevation) is the same. Along circuit D, the push might be in one direction then in the other, but around the whole circuit, the net circulation is zero. Otherwise you could go up or down forever like an Escher staircase.

Curl is the amount of circulation per unit of area, and would be experienced as a twisting or turning force. You can experience curl for yourself. Anyone with boating experience has experienced curl when the current vectors at one end of the vessel are stronger (or have a different direction) than at the other. The twisting force shows the curl of the current's vector field across the surface of the water. Whirlpools and hurricanes also illustrate curl. [4]

Gradient, divergence, and curl play a role in our day-to-day radio operating and are illustrated by the online video in Note 5. We'll take a closer look at the equations and find out how they lead to electromagnetic waves in the next column.

Pencils Down

While I'll still write for *QST* and ARRL, this and the following column are the conclusion of a wonderful 15-year run during which "Hands-On Radio" covered topics from simple component characteristics to transistor and op-amp circuit design, using simulator and design software, and now Maxwell's equations. Your response has been great, and I am indebted to all the experts who pay close attention, keeping me on the rails at times with suggestions and (cough, cough) corrections on occasion. Thank you.

Notes

1

All previous "Hands-On Radio" experiments are available to ARRL members at
www.arrl.org/hands-on-radio
.

2

en.wikipedia.org/wiki/Oliver_Heaviside
and Paul Nahin, *Oliver Heaviside: The Life, Work, and Times of an Electrical Genius of the Victorian Age*, IEEE, 2002.

3

J. Rautio, AJ3K, "The Long Road to Maxwell's Equations," Dec. 2014, *IEEE Spectrum*, pp. 36 – 40, 54 – 56, and
www.microwaves101.com/encyclopedias/maxwell-s-equations
.

4

earth.nullschool.net

5

www.youtube.com/watch?v=qOcFJKQPZfo

Experiment #179 — Maxwell's Equations — The Wave Emerges

Using the mathematical equations of divergence, gradient, and curl from last month, let's find out where radio waves come from.
[1]

Maxwell's first equation, also known as Gauss' law for electric fields ($\nabla \bullet E = q_V/\varepsilon_0$), tells us that if we measure where an electric field is pointing and how strong it is (the E-field's divergence or $\nabla \bullet E$) all around some arbitrary point, then we can tell how much electric charge, q_V, is at that point. In a very oversimplified way, we can think of the electric field's divergence as mapping out a distortion of otherwise-neutral space. Multiplying the distortion by permittivity, ε_0, the "electric stretchiness" of space, tells us the amount of charge, q_V, it takes to produce that distortion.

In more visual terms, you can "put a bag" around a point as in Figure 1A, add up the electric field everywhere it crosses the surface of that bag, and deduce how much electric charge must be inside the bag. There must be an equivalent magnetic charge that produces its own type of distortion we experience as, *H*, the magnetic field, right? No.

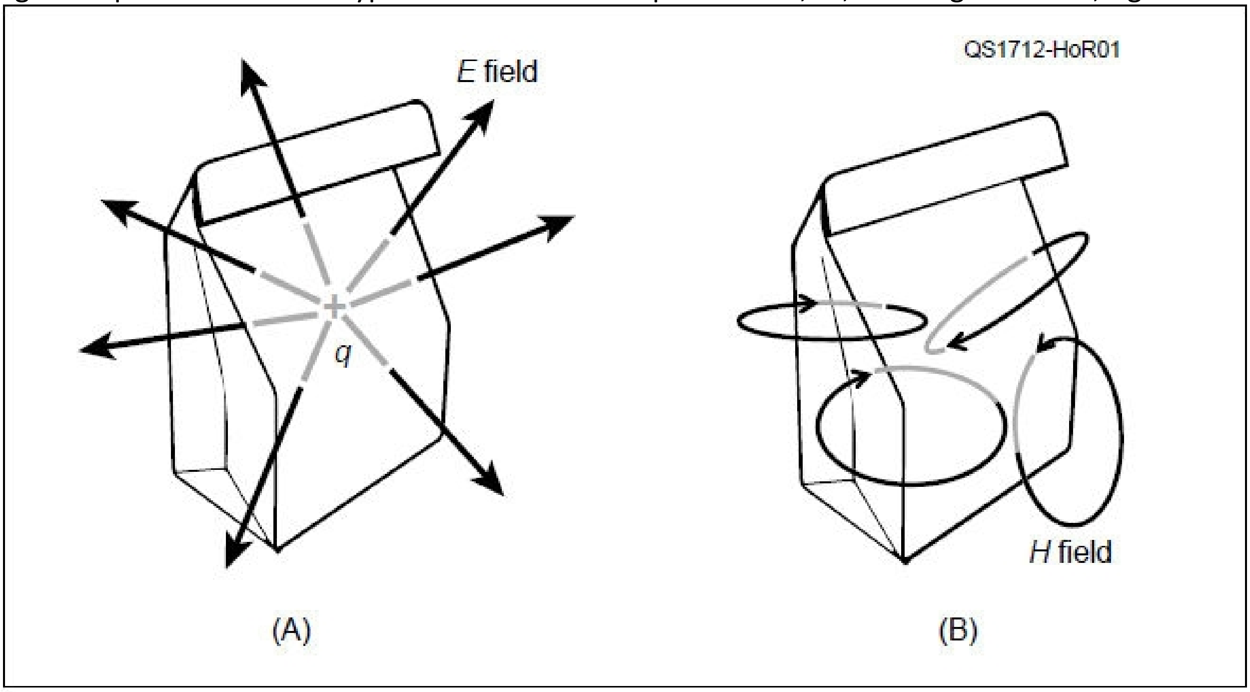

Figure 1 — By enclosing some point with any surface, even a bag, and measuring the *E* field everywhere it crosses the surface as in A, you can tell how much charge is inside. Because the magnetic field forms closed loops, as at B, any loop crossing the surface must go both in and out, so the total crossings are zero and no magnetic charge exists.

Maxwell's second equation, also known as Gauss' law for magnetic fields ($\nabla \bullet H/\mu_0 = 0$), tells us there can never be any "magnetic charge" inside the bag. Where the electric field can be visualized as field lines from an electric charge streaming off into space, the equivalent lines of the magnetic field are all closed loops with no beginning or end. Every loop crossing the bag's surface both enters and leaves, so the net total is always zero, as in Figure 1B. (If the loop is completely inside or outside the bag, it isn't counted.)

Magnetism — Charge in Motion

If there are no magnetic charges, what generates a magnetic field? This is where Maxwell's addition to Ampere's law, his fourth equation, comes in. It has two terms on the right-hand side: $I + m_0(dE/dt)$. The first is current, *I*, which is moving charge, at the point where the magnetic field is created. The second is the rate at which the electric field at that point is changing multiplied by permittivity. To add these two quantities together, they have to be of the same type with the same units. The first term is unambiguous — it's current — so the second quantity must also be current, or at least something equivalent to current. Maxwell called this term *displacement current*.

Displacement current comes from a time-changing electric field — how can that be created? Maxwell's third equation (Faraday's law) says that a time-changing magnetic field will do the job, but we're trying to create the magnetic field in the first place. According to the four equations, the only other way is to change the amount of electric charge at the point where we're trying to create the field.

Regarding the change in the amount of charge in our "bag of charge," the Law of Conservation of Charge is pretty clear — charge cannot be created or destroyed. If I want to change the amount of charge, I have to *move* some charge into or out of the bag, and moving charge is current. To shorten a really long story, the only way to create a time-changing electric field is by moving or *displacing* charge, which is current. That's how the electric field in a capacitor is created, by moving electrons off one electrode and onto the other. What we have just learned is that magnetic fields are the effect of electric charge in motion.

The Wave Equation

We're not quite there yet. We need one more equation. (Serious students of electromagnetism will want to dive into

Fleisch's book on Maxwell's equations.
[2]
) The equation below is the *ideal wave equation*
[3]
(without loss or non-linearities), and it describes all waves:

$$\nabla^2 A = \frac{1}{v^2}\left(\frac{d^2 A}{dt^2}\right)$$

Let's break this into digestible pieces. *A* is whatever field in which the wave is created. We use *E* or *H* fields for radio waves. The right-hand side's first term ($1/v^2$) is just the reciprocal of the velocity of the wave squared. In the second term, the superscript "2" above the *d* and *t* means "do this twice," not "squared." If the rate of change of our field is dA/dt, then d^2A/dt^2 means "the rate of change of the rate of change." (For example, acceleration is the rate of change of velocity which is the rate of change of position. Hold that thought.) The intimidating term on the left ($\nabla^2 A$) describes how the gradient of the field in all three dimensions is changing — the gradient of the gradient, basically.

Reading from right to left, what the equation tells us is, Changes in the rate at which the field A varies, create changes in the field's strength throughout space. The changes are inversely proportional to the speed at which the changes propagate. Solutions to that equation, such as sine waves, describe wave motion, whether as a plucked string, a ripple across water, or a radio wave in space.

Coupling *E* and *H*

The left-hand side of Maxwell's third and fourth equations don't describe the *E* and *H* fields directly. They give us the curl of the fields ($\nabla \times E$ and $\nabla \times H$). I don't have room to show the complete process (it's in Fleisch's book), but with two mathematical tools (Stoke's theorem and the divergence theorem), we can get from the third and fourth equations to a wave equation that replaces *A* in the wave equation with *E* or $\mu_0 H$.
[4]
,
[5]

We are really close now — take three big steps then one final leap. Step 1: Equations three and four show that time-changing *E* and *H* fields generate each other (and if you follow the vector math for waves, *E* and *H* are at right angles in free space, as shown in Figure 2).

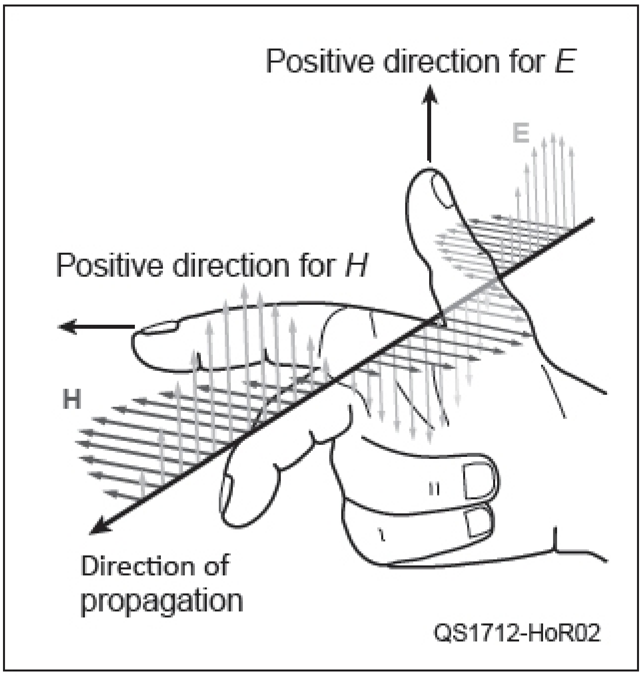

Figure 2 — The right-hand rule shows how to determine direction of propagation of an electromagnetic wave. Point your thumb in the positive direction for the *E* field, your index finger in positive direction for *H*, and your middle finger will point in the direction the wave is traveling

Step 2: The only way to get those time-changing fields is to move charge — change its position with time — and the only way for that right-hand term in the wave equation to not be zero is for the charge to be accelerating (or decelerating).

Step 3: Motion is a change in position relative to an observer (i.e., me), so the resulting changes in the fields also appear to be moving from the perspective of the observer.

And here we are, at last: The wave equation describes those changing fields, and we have our electromagnetic radio waves, generated by accelerating and decelerating charge, better known as ac current.

The coupling of the electric and magnetic fields creating each other explains why an electromagnetic wave is more than just an electric field and a magnetic field that just happen to be at the same place and just happen to be at right angles. They aren't separate things at all. The *E* and *H* fields are two aspects of the *same* thing — an electromagnetic wave moving through space.[6]

Even with all the math, we're still not all that clear on what's really going on. From Jim Rautio, AJ3K:

> We talk about electric fields and magnetic fields as though they are real. Sure, you have seen iron filings move around, but no one has ever seen or touched a field. Like lines of force, fields are just a mathematical convenience that allows us to predict what happens when we do an experiment.[7]

Nor does anyone know what a photon or electrical charge actually is. We just know how to observe their effect on the space we live in.

Having completed our journey through the Land of Maxwell, this is a great place to end "Hands-On Radio" — at the headwaters of all radio waves, whose magic is at the heart of ham radio and the people who enjoy it. 73!

Notes

1

All previous "Hands-On Radio" experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

D. Fleisch, *The Student's Guide to Maxwell's Equations*, Cambridge University Press, 2008.

3

The correct form uses partial differentials, δ, but for simplicity, the simple derivative symbol is used here. A full discussion of the wave equation is well beyond the scope of this overview.

4

The relationship between *B* and *H* is explained at **www.physicsforums.com/threads/in-magnetism-what-is-the-difference-between-the-b-and-h-fields.370525**.

5

It was not lost on Maxwell that the same process also shows *v*, the wave's velocity, is equal to $1/\sqrt{e_0 m_0}$, which just happens to also be *c*, the speed of light.

6

See the electromagnetic wave animation at **commons.wikimedia.org/wiki/File:EM-Wave.gif**.

7

J. Rautio, AJ3K, "The Long Road to Maxwell's Equations," Dec. 2014, *IEEE Spectrum*, pp. 36 – 40, 54 – 56, and **www.microwaves101.com/encyclopedias/maxwell-s-equations**.

Experiment #122 — Battery Characteristics, Part 1

This month, we begin a two part article about batteries and battery characteristics with a discussion of what makes batteries "go" and how the different types behave electrically. Next month, we'll observe the differences for ourselves.

Basic Battery Construction and Chemistry

Viewed simply, the battery is a self-contained chemical reaction vessel in which chemical substances give up electrons that flow through an external circuit to other types of substances that accept electrons. The battery's *separator* keeps the chemicals, well, separated so that the electrons have to make the trip through the circuit to be exchanged.

The electrons do some useful work along the way, converting the chemical energy to electrical energy and then to whatever type of energy the user extracts from the circuit — mechanical, heat, electromagnetic, etc. The reaction continues until all of the available electrons have been exchanged through the external circuit, depleting the chemicals and discharging the battery.

The types of atoms or molecules involved in giving up and accepting the electrons — called the *battery chemistry* — determine the *electromotive force* (EMF) that pushes the electrons through the circuit. Each type of atom or molecule has a certain affinity for electrons: Via a specific chemical reaction, some want to get rid of them and some want to acquire them. The strength of that reaction's electron exchange is the reaction's *electropotential,* which is measured in volts. The difference in electropotential between the chemicals is what determines the *terminal voltage* of the battery. (The list of electropotentials for common materials and reactions is also known as the *galvanic series*.)

Batteries are classified in two groups; *primary* (non-rechargeable) and *secondary* (rechargeable). In both groups, giving up and accepting electrons changes the chemicals into different compounds. (The atoms are still the same types of atoms but the rearrangement of their electrons changes the structure of molecules made from those atoms.) In a *primary* battery, the reaction is not reversible even if a voltage is applied externally to make the electrons flow "the other way." In a *secondary* or rechargeable battery, the reaction will run in reverse if powered by an external voltage, restoring the original chemicals and recharging the battery.

Basic Battery Terminology

Let's start with *capacity,* which is given in *ampere-hours* (*Ah*). Amperes (coulombs of charge per second) multiplied by time results in an amount of charge. (1 Ah = 1 coulomb/s × 3600 s/h = 3600 coulombs) Thus, capacity measures the number of electrons that a battery can cause to flow through an external circuit. Because capacity is an amount of charge, it is abbreviated C for coulomb. Causing large numbers to flow (high current) discharges a battery's capacity quickly and low current discharges it slowly.

Capacity is independent of terminal voltage: A specific pair of chemicals has the same relative electropotential no matter what quantity of those chemicals is contained in the battery. Thus, terminal voltage is the same for large and small batteries of the same battery chemistry. The larger the quantity of chemicals in the battery, however, the more electrons can be exchanged.

Capacity also indicates the amount of energy stored in a battery. Since the terminal voltage is relatively constant and voltage is joules (J) per coulomb, multiplying capacity times ter- minal voltage yields energy, usually in units of *watt-hours* (Wh). For example, a capacity of 2 Ah and a 1.5 V terminal voltage represents 2 Ah × 1.5 V = 3 Wh. (Note that not all of the stored energy can be delivered to the external circuit and that terminal voltage drops as the battery is discharged.)

Another important battery characteristic related to its capacity is the battery's *specific energy,* which is given in watt-hours per kilogram (Wh/kg). Batteries with high specific energy store a lot of energy for a given weight.

In battery literature, you will also encounter *C-rate,* which is the rate at which a battery is charged or discharged measured in terms of its capacity. If a 1000 mAh battery is discharged at a current of 1000 mA, it is being discharged at a rate of 1 C. At 500 mA, the rate is 0.5 C, and at 100 mA the rate is 0.1 C. Because batteries are made in so many different sizes using the same chemistry, C-rate is a useful way to talk about battery performance and maintenance independent of size.

Types of Battery Chemistry

Several battery chemistries account for most needs of the Amateur Radio operator: alka- line, lead-acid, nickel-cadmium (NiCd), nickel-metal hydride (NiMH) and lithium-ion (Li-ion). Table 1 gives the basic characteristics of these battery types. [1]

Alkaline batteries are primary (non-rechargeable) and the rest are secondary (rechargeable). While re-chargeable batteries offer higher specific energies and lower costs over the life of the battery, alkaline batteries do not require a charger, which can be important for emergency situations when ac power is not available. They also have a long shelf life.

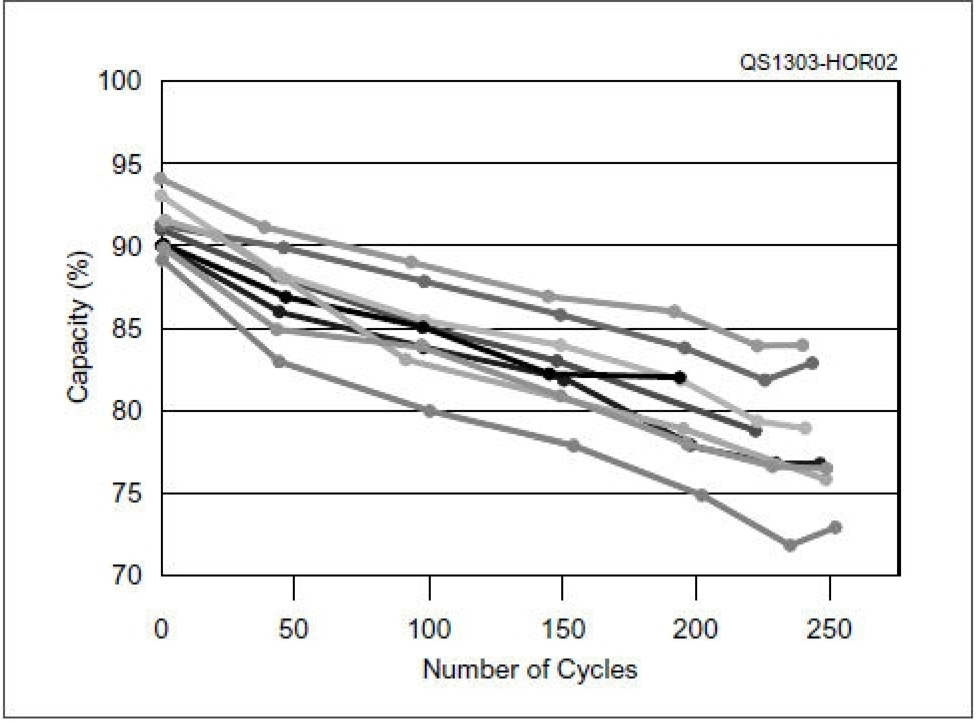

Figure 2 — The effect of repeated charge-discharge cycles on a set of new 1500 mA Li-ion batteries. After 200 cycles, most of the batteries had lost 10% of their capacity. [Courtesy of Isidor Buchmann and Cadex]

Effect of Discharge Rate

Table 1 shows peak discharge rates for the various types of batteries in terms of capacity, however, that is not the recommended rate of discharge during normal use. The amount of energy a battery can deliver is maximized at a much lower rate, shown in the row "Best discharge rate" in Table 1. Note that all types of batteries perform best well below the peak discharge rate. Figure 1 shows the effect on a NiCd battery pack's lifetime at different discharge rates. Other types of batteries are even more strongly affected by discharge rate.

Table 1
Characteristics of Common Battery Chemistries

Specification	Alkaline	Lead-Acid	NiCd	NiMH	Li-ion	Cobalt	Manganese-Phosphate
Specific energy (Wh/kg)	210	30-50	45-80	60-120	150-190	100-135	90-120
Cycle life	1	200-300	1000	300-500	500-1000	500-1000	1000-2000
Self-discharge/month	<1%	5%	20%	30%	<10%	<10%	<10%
Cell voltage (V)	1.5	2	1.2	1.2	3.6	3.8	3.3
Peak load current	<0.5C*	5C	20C	5C	>3C	>30C	>30C
Best discharge rate	<0.1C*	0.2C	1C	0.5C	<1C	<10C	<10C
Toxicity	Low*	Very high	Very high	Low	Low	Low	Low

*evaluation based on public literature
Reprinted from *Batteries In A Portable World*

Effect of Depth of Discharge

Depth of Discharge (DoD) has a large effect on the number of charge-discharge cycles a battery can provide. The more, or "deeper," =a battery is discharged, the more it is stressed. Table 2 provides an idea of the effect on a Li-ion battery lifetime when repeatedly discharged to a specific level. Partial discharges preserve battery life.

Table 2
Effect of Depth of Discharge (DoD) on Cycle Life

Depth of discharge	Discharge cycles
100%	500
50%	1500
25%	2500
10%	4700

Reprinted from *Batteries In A Portable World*

Effect of Repeated Charge-Discharge Cycles

As most users of rechargeable batteries quickly discover, a battery performs like new over a number of charge-discharge cycles and then begins to lose capacity. This is due to changes at the microscopic level in the materials of the battery. For example, when a battery is new, the chemicals are typically in the form of very small crystals that provide lots of surface area to exchange electrons. With each cycle, however, the crystals grow in size and that reduces the total surface area and battery capacity.

Figure 2 shows the effect of repeated cycling on the capacity of a set of identical, new 1500 mAh Li-ion batteries. Even though the battery may be able to support hundreds of cycles, after 200 cycles most of the batteries had lost 10% of their initial capacity.

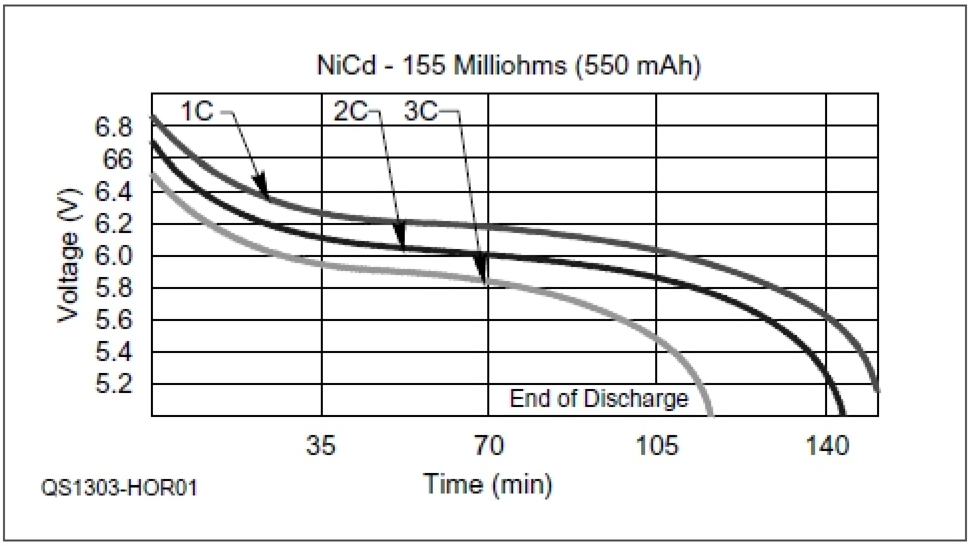

Figure 1 — The effect on a NiCd battery pack lifetime at discharge rates of 1, 2 and 3 C. [Courtesy of Isidor Buchmann and Cadex]

Effect of Temperature

Because batteries are based on chemical reactions that vary in rate with temperature, capacity is also affected. For example, a lead-acid battery loses about 10 to 20% of its capacity between room temperature (23° C) and freezing (0° C). By –20° C, about half of the capacity is lost. Thus it is very important for capacity to be measured at a specific temperature when comparing batteries. Automobile and deep cycle batteries have a *Cold Cranking Amps* (*CCA*) rating that must be specified at –18° C (0° F) for just this reason.

That is not to say that everything gets better with increasing temperature. From *Batteries in a Portable World*, "The optimum operating temperature for a [lead-acid] battery is 25° C (77° F). As a guideline, every 8° C (15° F) rise above this temperature cuts battery life in half. At 33° C (95° F), that battery's lifetime is cut in half and is reduced to 10% at 45° C (107° F)."

Measuring Battery Performance

Next month we are going to both measure some of these effects on actual batteries and introduce you to a new type of voltmeter that has the ability to act as a *data logger*. Once you realize how useful data logging is, you'll find all sorts of uses for it in the shack and around your home.

1

I. Buchmann, *Batteries In a Portable World*, 2011, pp 34-37. Available from your ARRL dealer or the ARRL Bookstore, ARRL order no. 1156. Telephone 860-594-0355, or toll-free in the US 888-277-5289;
www.arrl.org/shop
;
pubsales@arrl.org
.

Experiment #123 — Battery Characteristics, Part 2

In last month's column, I explored some of the basic terminology used to describe and compare batteries. [1]

Material from *Batteries For a Portable World* clearly showed the differences between the common types of batteries that hams use to supply power for radios and accessories. [2]

This month, we'll measure some common batteries by using the *data logging* function of an inexpensive digital multimeter (DMM).

Data Logging Voltmeter

Let's face it, taking regular measurements of a slowly changing parameter is bor-r-r-r-ring. I've done my share of watching a ticking clock and meter or gauge but today there are automated tools to do that job. They never get distracted, forget, or misread the data. The tool we'll use this month is the data logging DMM.

A full fledged data logger, such as a Fluke 2625A Hydra model (**www.fluke.com**) with multiple channels and high speed high accuracy precision measurements, is way beyond the needs of a typical ham. What we need is a single channel voltmeter with an interface to a PC.

The smaller sibling of the 2625A is Fluke's 289 DMM, which has impressive specifications for a voltmeter. Features include a USB interface and companion software so you can store or *log* data on a PC. This is a top-of-the-line DMM with a $600 price tag (the older Fluke 189 sells for a couple hundred less). I love my ultra reliable Fluke DMM but for ham shack data logging, I needed a less expensive solution.

A trip to the Jameco catalog (**www.jameco.com**) turned up the under $100 *house brand* MS8226, Jameco p/n 137462, with decent specs and an RS-232C interface. Searching the usual Internet bargain sites also turned up similar meters, some for as little as $30. You might also get lucky by watching for used meters from Fluke and other high end manufacturers.

The MS8226 has all the usual functions, plus temperature (°C) with an included thermocouple, capacitance (50 nF to 100 µF), frequency (to 5 MHz), duty cycle and true RMS measurements with an unspecified upper frequency limit. The RS-232C interface requires a USB-to-serial converter or running the host software on an older PC with a serial port.

Without making this a product review, I'll just say that the meter works as advertised and includes an easy to understand manual. The software is very basic, but useful as a means of creating time stamped data files of measurements from the voltmeter. Once the data is on the PC, you can export it into spreadsheet format for graphing or analysis as described later.

Winding Your Own Wirewound Resistors

If you don't have a power resistor handy, you can wind your own resistor from common copper wire. The *ARRL Handbook* gives resistance in Ω / 1000 feet. For a 1 Ω resistor, #20 AWG at 10.1 Ω / 1000 feet requires about 100 feet, and #30 AWG at 104 Ω /1000 feet requires about 10 feet. Wind enameled wire on a ceramic, glass or non melting plastic tube. A drop of epoxy will hold the wire in place.

Loading and Testing Batteries

We're going to record battery voltages with a resistive load applied every few seconds over an extended period of time. Comparisons will be made by manually transferring data into a multi-column spreadsheet for graphing. (The spreadsheet used for this column is available on the Hands-On Radio website.)

Remember that batteries store a lot of energy. When choosing a load, plan for the heat that must be dissipated by using resistors with an adequate power rating and keeping them off of surfaces that can be damaged by elevated temperatures. The maximum power dissipation during a load test will be E^2/R, where E is the battery terminal voltage.

Use load resistances that will draw some- what more than the Best Discharge Rate current in Table 1 of the previous experiment. Values around 1 Ω will work well for these tests, drawing a maximum current of 1.5 V / 1 Ω = 1.5 A, and dissipating a little over 2 W ($1.5\ V^2$ / 1 Ω = 2.25 W). With this power dissipation a 5 W resistor can get hot enough to burn you or a workbench surface. (Don't use incandescent lamps as loads — their resistance varies with current.)

If you make your own loads, use several resistors in series or parallel to spread out the heat as shown in the photo of my test set in Figure 1. My loads are made out of paralleled 2 W metal oxide resistors. They are soldered between SO-239 UHF coax sockets made into a frame with #6-32 screws and threaded spacers. A banana plug fits snugly into the SO-239 making a fine high current connection or, at lower currents, clip leads can be used. Typical plastic battery holder contacts and wiring may not be heavy enough to handle the higher than normal discharge currents in these tests. I used copper pennies held in an insulating vise as my fixture contact with an extra heavy clip lead to the load.

Figure 1 — The test setup for battery load tests. Loads are constructed from multiple resistors in parallel. The battery is held between two copper pennies in a vise with insulating jaws. Clip leads are used to connect the battery, load and logging DMM.

AA to AAA Comparison

Before beginning, it's worth noting that what is usually referred to as a *battery* is a single package of chemicals more correctly referred to as a *cell.* A set of cells connected together form a *battery*, which derives from the original meaning of a group of identical pieces, such as an artillery battery of several guns. An assembly of six individual lead acid cells, each producing 2 V and connected in series create a vehicle's 12 V starting battery. In the case of single cell batteries, the word *cell* and *battery* are interchangeable.

Let's start by comparing two fresh batteries that use the same chemistry but have different capacities. I used AA and AAA alkaline batteries sold by Costco under their *Kirkland* brand name. Battery capacity is not specified by Costco but third party testing has found the AA cells to supply approximately 2300 mAh and the AAA capacity is probably about half that.

Figure 2 shows the initial portion of a 30 minute battery comparison when connected to a 1 Ω resistor load at room temperature (about 21° C). You can see that the initial terminal voltages are approximately the same and that the capacity of the smaller AAA battery is depleted more quickly. At the end of the load test, (not shown in Figure 2)the AA cell terminal voltage recovered to 1.38 V and the AAA cell to 1.22 V.

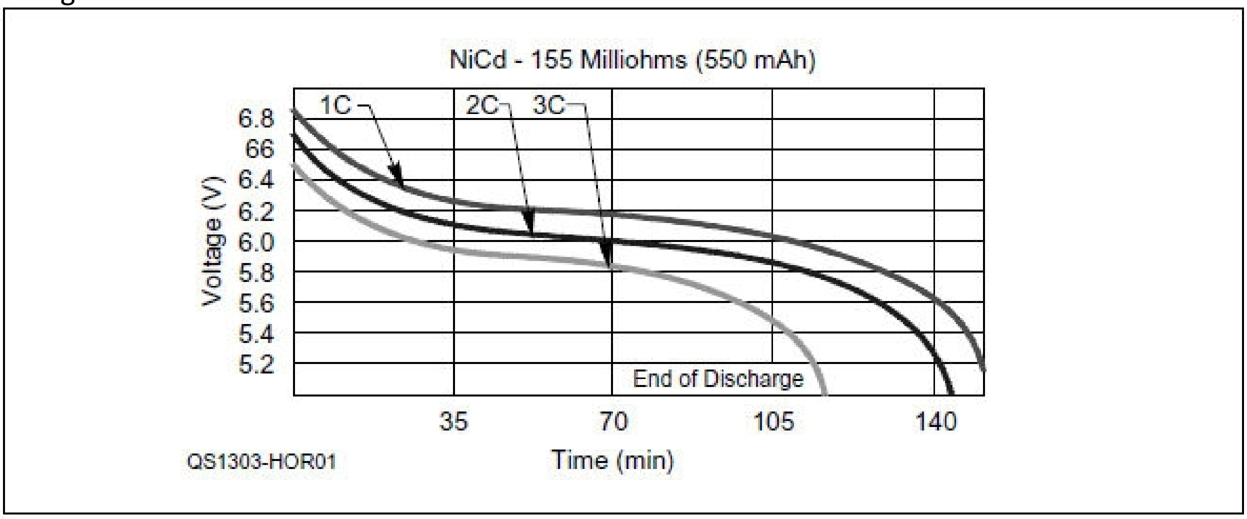

Figure 2 — While AA and AAA alkaline batteries have the same chemistry and open circuit voltage, the capacity of the smaller battery is more quickly depleted under load.

Alkaline to NiMH Comparison

The open circuit voltage of the fresh rechargeable NiMH AA cell shown in Figure 3 is lower than that of a fresh alkaline cell by about 0.3 V. Both drop about the same amount when connected to a 1 Ω load. That difference narrows to 0.2 V after about 6 minutes. The 0.3 V margin can translate to a lot of extra operating time if alkaline cells are used.

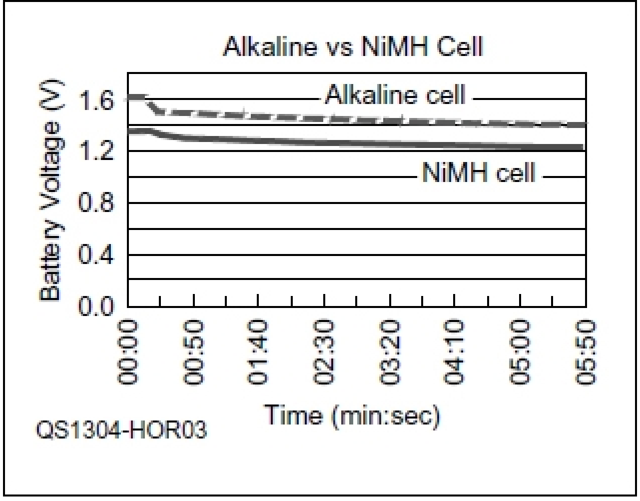

Figure 3 — The alkaline battery has a higher terminal voltage than the NiMH, although both have similar capacities and discharge at about the same rate under load.

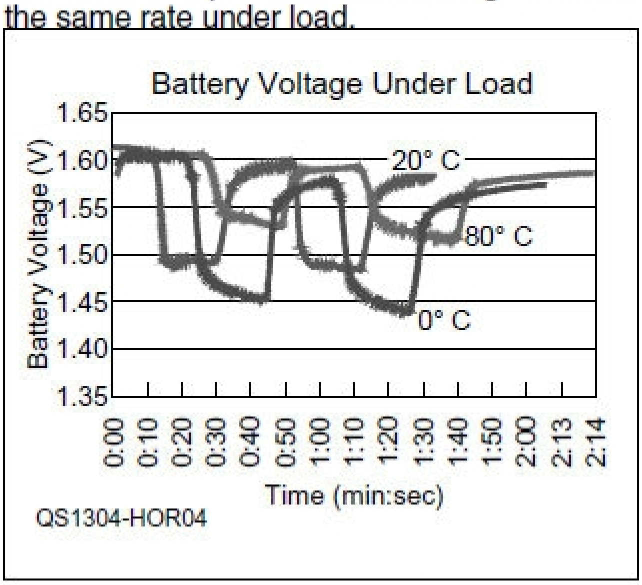

Figure 4 — Low temperatures increase a battery's internal resistance and voltage drop under load by slowing the chemical reactions that supply electrons to the external circuit. High temperatures accelerate the chemical reaction and lower internal resistance.

Temperature and Internal Resistance

A battery's temperature affects its internal resistance quite a bit as you can see in Fig-ure 4. The same fresh battery was tested at 0, 20 and 80° C by connecting it to a 1 Ω load for 10 s with a 10 s rest between load periods. The open circuit terminal voltage was approximately the same at all three temperatures, varying only 8 mV from 1.604 to 1.612 V. The cold battery voltage dropped substantially under load — initially about 0.26 V at 1.6 A load, implying an internal resistance, R_{INT} of 0.26 / 1.6 = 0.16 Ω. At room temperature, voltage dropped 0.125 V for R_{INT} = 0.078 Ω. At 80° C (hot enough to burn the experimenter's fingers!) the voltage drop of 0.064 V indicates R_{INT} = 0.04 Ω, a 2:1 variation with temperature. This can be important when trying to get the most performance from a battery over a wide temperature range!

Other Data Logging Tasks

Obviously, data logging can be put to many other uses, such as recording temperature or current consumption. As with the radio astronomy project, logging can record audio (noise levels in that case) as well. A spreadsheet can convert voltages from sensors directly into physical units and combine different data elements to measure differential temperatures, ratios, minimum and maximum values and so forth. Best of all, a data logger can patiently record data to catch a power dropout or intermittent noise that never seems to happen when you're around.

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at

www.arrl.org/hands-on-radio
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2
I. Buchmann, *Batteries In a Portable World*, 2011, pp 34-37. Available from your ARRL dealer or the ARRL Bookstore, ARRL order no. 1156. Telephone 860-594-0355, or toll-free in the US 888-277-5289;
www.arrl.org/shop
;
pubsales@arrl.org
.
3
www.batteryshowdown.com/results-lo.html

Experiment #142 — Inductors at RF

In one of the many strange-but-true things that happen at RF, that innocent-looking coil of wire or cable has more than one personality as the frequency changes! This month we'll explore the wacky world of inductors and learn how to use a neglected function of a common antenna analyzer along the way.

Inductor Basics

This formula for the inductance *L* of a basic single-layer, air-wound inductor has been in articles and handbooks for generations [1], [2]:

$$L(\mu\text{H}) = \frac{d^2 n^2}{18d + 40\ell}$$

Where *d* is the diameter of the coil in inches from wire center to wire center, ℓ is the coil's length in inches, and *n* is the number of turns. This approximation works reasonably well but there are innumerable corrections. [3]

The formula makes several assumptions that the coil: is made from wire that is not too thick; is not too long or too short; has leads not too long; and has a reasonable *pitch* (the number of turns per unit of length).

Why does frequency matter? The inductor model in Figure 1 gives part of the answer. This parallel-series circuit represents what an RF signal encounters in an inductor. Instead of just inductance (*L* in the schematic), there are three other *parasitic characteristics* shown; C_P, R_P, and R_S resulting from the physical construction of the inductor. *L* is the inductance independent of parasitic effects.

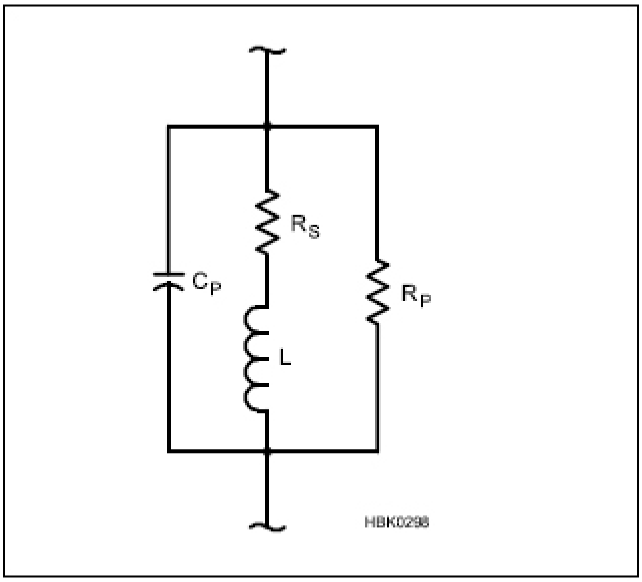

Figure 1 — General model for an inductor with parasitic capacitance (C_P) and resistances (R_P and R_S). R_S can change with frequency due to the skin effect. The combination of C_P and *L* are the cause of self-resonance in the inductor.

R_P is the simplest of the three parasitics, representing *leakage resistance*, resistive current paths "around" the inductor for current. Dirt or grease on a circuit board and dust buildup on the turns of the inductor or the body of an encapsulated inductor are the most common sources of leakage resistance. It becomes significant when there is a high voltage across the inductor, such as might exist in a transmitter or tuning unit. This is a good reason to vacuum out high-power circuits from time to time.

R_S has a larger effect on the inductor's performance than R_P, especially at high frequencies. At dc, R_S is specified as *DCR*, or dc resistance. The resulting voltage drop or resistive heating can be important when the inductor has to carry dc current, such as when an RF choke is used in a bias T or a plate blocking choke.

If the inductor is used at RF, *skin effect* comes into play, restricting current flow to a layer near the surface of the conductor. [4]

This causes R_S to increase with frequency. Inductors used in transmitters and tuning units often carry significant current so it is important to consider skin effect when selecting the size of a coil's wire or tubing. Resistive losses lower the inductor's Q, its ratio of reactance to resistance: X_L / R_S.

C_P has the largest effect on inductor performance at RF. By creating a parallel-LC circuit, the combination of C_P and L means the inductor will resonate without any other external components. This creates the inductor's *self-resonant frequency* or *SRF*. We observed the effects of self-resonance in Experiment #111 on coiled-coax chokes. [5]

The coiled-coax choke makes use of the parallel resonance's high impedance to block current flow on the outside of the coax shield over a range of frequencies.

Where does C_P come from? Figure 2 shows that C_P results from *inter-turn capacitance*. Each spot on the inductor wire forms a small capacitance to every other spot on adjacent turns, even though they are connected together by the wire. Over the entire inductor, C_P is called *distributed capacitance*. Ways to reduce C_P include stretching the coil so that the turns are farther apart or in the case of multi-layer coils, carefully arranging the winding layers and winding the coil in sections.

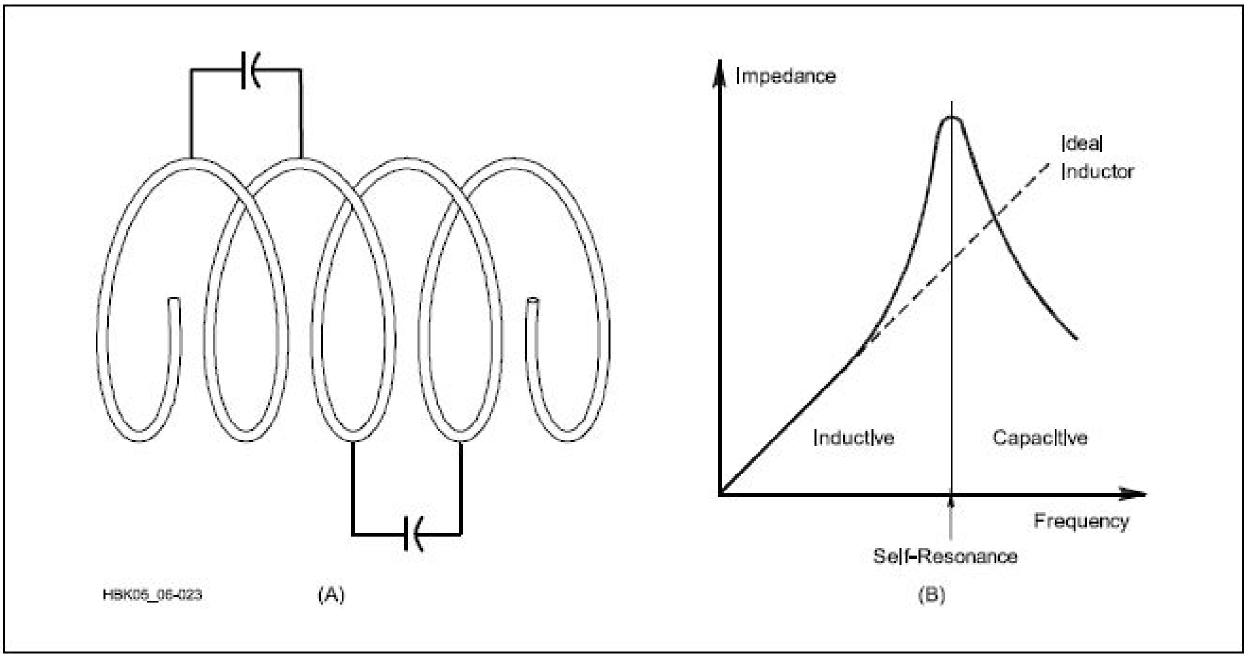

Figure 2 — Inductors have distributed capacitance created by the capacitance between turns of the coil. Over the whole inductor, this capacitance creates C_P as shown in the model in Figure 1. The graph at B shows how the inductor behaves above and below its self-resonant frequency.

Measuring Inductance

If you have access to an antenna analyzer that displays reactance, you can measure an inductor's SRF and see the effects for yourself. As in Experiment #111, we'll use the popular MFJ-259/269-series of antenna analyzers.

Start by obtaining 8 to 10 feet of coaxial cable. Any of the RG-8/213/58/59 family will do — the characteristic impedance of the cable is unimportant as we are only interested in what happens on the *outside* of the shield. Wind the cable around a non-conducting form such as the peanut butter jar in Figure 3. Remove a short section of jacket from each end, twist the shield into a lead, and connect it to the analyzer.

Figure 3 — Using an antenna analyzer to measure the inductance of a coiled-coax choke. The coil is wound around a. 3-inch diameter peanut butter jar. The jacket at each end of the cable is stripped back about 1 inch to allow the braid to be made into a lead and connected to the analyzer using a binding post adapter.

My coil of RG-58 cable has 91⁄2 turns, it is 3 inches (76 mm) in diameter, and is 2 inches (51 mm) long. According to the equation at the start of this article, its inductance should be 6 µH. However, the equation isn't intended to apply to a close-wound coax coil, so I turned to the online inductance calculator by ON4AA at hamwaves.com/antennas/inductance.html. This calculator takes the diameter of my "wire" (4.5 mm) into account, as well, producing an inductance of 5.6 µH at a frequency of 1 MHz.

Connect the coil to your antenna analyzer, using a binding post adapter as shown in Figure 3. (Keep the analyzer and coil away from metal surfaces.) The following instructions apply to the MFJ-259/269 analyzer. Turn on the analyzer and press the MODE button until the display shows Inductance in uH. Clockwise from the upper left, the display shows frequency, reactance value, the label XL, and inductance in μH. The manual explains that the calculation is based on reactance and the analyzer itself can't tell whether the reactance is inductive or capacitive. You have to figure that out — if increasing the frequency causes reactance to increase, the reactance is inductive.

Start with your analyzer at its lowest frequency. (1.7 MHz on my analyzer.) The inductance value of my coil was 6.6 μH. This is not too far from the calculated value which didn't account for the jacket plas-tic's effect on C_P or the extra lead length from the coil to the analyzer. Slowly increase frequency. My inductance value stayed fairly steady near the calculated value until I passed 3 MHz and then began increasing. Why? As Figure 3B shows, the impedance of the resonant circuit of the inductor increases faster than that of an ideal inductor, causing the analyzer to see a "bigger" inductance.

Press MODE until you are back in the analyzer's usual "Impedance R&X" mode. Keep increasing frequency while watching the X value. (Ignore the SWR and Resistance meters.) You'll see it increase faster and faster until it exceeds the meter's ability to measure reactance and it displays Xs = 0. Keep increasing frequency and watch the *Rs* display, the equivalent series resistance value of the impedance. It will continue to increase, exceeding the analyzer's range of 1500 Ω as it approaches the coil's SRF.

Continue to increase frequency and after you pass the SRF, impedance will eventually come back into range and keep dropping as frequency increases, just like in Figure 3B. Switch back to inductance measurement and repeat the sweep through the coil's SRF. Note that above the SRF, reactance drops as frequency increases, showing that the reactance is capacitive. The inductor has changed into a capacitor!

Connect a 1.5 kΩ resistor (carbon composition or film will do) across the analyzer terminals to keep the impedance within the analyzer's range and switch back to impedance mode. You can find the coil's SRF by adjusting frequency to find the maximum value of *Rs*. My coil's SRF was 13.6 MHz where the meter displayed 1488 Ω. This would be a good choke for a 20 meter antenna! Now try spreading the turns apart, reducing C_P, to see how that affects the SRF. You can also try an equivalently sized inductor out of insulated wire. A simple inductor? Not really!

What would happen if I tried to use this coil at VHF? C_P would make the inductor unusable. In fact, even small coils can become unusable at and above VHF due to parasitic capacitance. Capacitors with significant amounts of parasitic inductance can become unusable at those frequencies for similar reasons. Knowing the actual characteristics of your components is important for successful RF design and construction.

Notes

1

H. A. Wheeler, "Simple Inductance Formulas for Radio Coils," *Proc. I.R.E.,* Vol. 16, p 1398, Oct 1928.

2

F. E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, p. 55, 1943.

3

F. W. Grover, *Inductance Calculations*, Dover Publications, 2009.

4

The ARRL Handbook, 91st edition, ARRL, Sections 5.3.4 through 5.3.7.

5

All previous Hands-On Radio experiments are available to ARRL members at
www.arrl.org/hands-on-radio
.

Experiment #147 — Capacitors at RF

Paraphrasing the opening sentence of Experiment #142, "Inductors at RF," when is a capacitor not a capacitor? When it is an inductor.

[1]

This odd fact of life at RF causes a lot of head-scratching and troubleshooting, particularly as the frequency of interest rises beyond the upper HF range. A capacitor's parasitic inductances and resistances are either hidden inside the capacitor or hidden in plain sight. Intriguing, isn't it? It's all part of learning to think like a radio wave.

The Non-Ideal Capacitor

Figure 1 shows a simple test circuit that includes a general-purpose model of a capacitor at RF and some simulated response curves. A nanofarad is not a nanofarad, it seems! The two resistances and the inductance are *parasitic components*. They are consequences of the way the capacitor is constructed and are present to varying degrees in all capacitors.

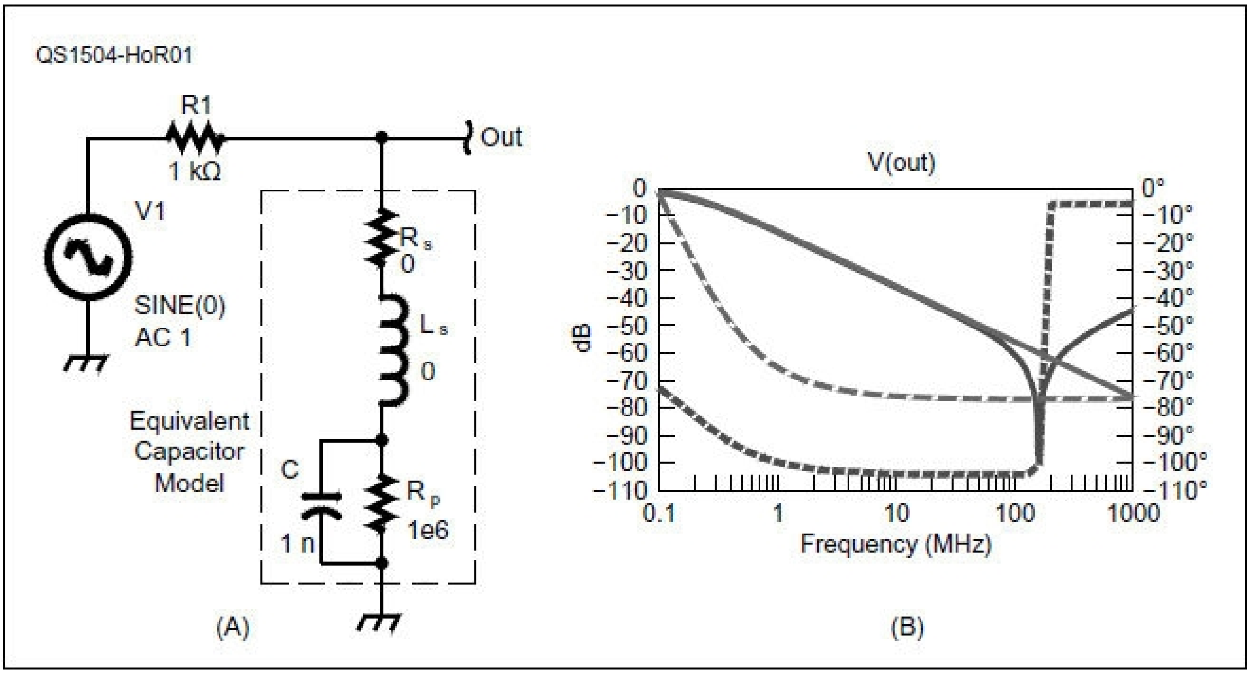

Figure 1 — *LTSpice IV* equivalent circuit or model (A) for a capacitor at VHF and higher frequencies includes both resistance of the leads (R_S) and dielectric loss (R_P) as well as the equivalent series inductance (L_S). (B) shows the effects of parasitics on amplitude (solid line) and phase (dashed line) response with L_S = 0 (blue) and L_S (red) = 1 nH.

- L_S is *equivalent series inductance*, or *ESL*, created by the capacitor's internal construction and connecting leads.
- R_S is *equivalent series resistance*, or *ESR*, and represents the resistance from skin effect of the electrodes and leads along with dielectric losses.
- R_P is *equivalent parallel resistance,* or *EPR,* and represents leakage losses.

Because the two primary sources of R_S (skin effect and dielectric loss) both change with frequency in different ways, they are often modeled as separate resistors in simulation calculations. In this column, we'll just combine them for simplicity.

Figure 2 shows how two common types of capacitors are constructed. At A, you can see a common "roll-type" construction used for inexpensive electrolytics and many plastic film-type capacitors used in applications below 100 kHz. From the view of the end of the roll, you can probably guess that L_S is pretty high compared to other types of capacitors. In fact, from the Cornell-Dubilier Electronics *Aluminum Electrolytic Capacitor Application Guide*, we learn that a typical value for L_S of an axial lead electrolytic ranges from the low tens of nH to around 200 nH.

[2]

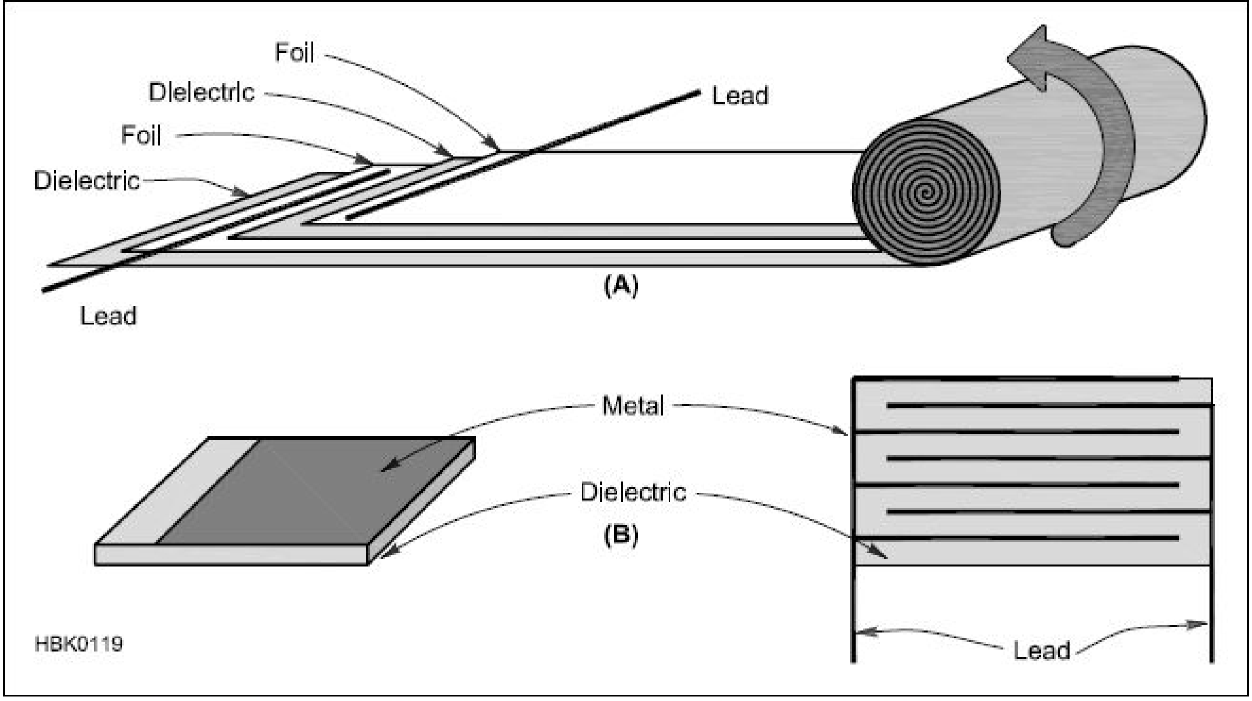

Figure 2 — Two common types of capacitor construction. (A) Roll construction uses two strips of foil separated by a strip of dielectric. (B) Stack construction made from layers of dielectric material (such as ceramic or film), one side coated with metal. Leads or a metal cap are attached to each side.

Figure 2B shows how a typical ceramic SMT (surface mount technology) ceramic capacitor is constructed. The capacitor is manufactured by placing an interleaved stack of metal foil and ceramic in a press and heating it to create a single *monolithic* (literally "one stone") metal-ceramic block. The ends of the metal layers stick out of each side, where a metal cap is applied to form the terminal. This type of capacitor has very, very low L_S — less than 1 nH.
[3]

Disc ceramic capacitors are similar, with leads attached to the metal layers on each end of the stack. Other types of capacitor such as tantalum, silvered-mica, air variable, trimmers, and so on have intermediate values of L_S.

Parasitic Effects

Measuring the effects of these parasitic components at RF is not all that easy, so we'll return to our *LTSpice* roots to have a look. (See Experiments #83 – 85.
[4]
)

Figure 1B shows the difference in frequency response from 100 kHz to 1 GHz between a nearly ideal 1 nF capacitor (R_S and L_S are zero and R_P is 1 MΩ) and one with 1 nH of inductance added. Just like an inductor's series capacitance creates a self-resonance, so does a capacitor's series inductance. As frequency increases through self-resonance, the dotted trace shows the capacitor's reactance changing to inductive, which then continues to increase with frequency.

The ideal component behaves as expected at all frequencies. The non-ideal component, with its series resonance near 2 meters, might not be the best choice for a bypass capacitor at VHF and above!

Circuit Behavior with Parasitics

Figure 3 shows a common-emitter (CE) amplifier, the subject of Hands-On Radio Experiment #1, way back when. The ac gain of a CE amplifier primarily depends on two things — the transistor's gain-bandwidth product (h_{FE}) and the ratio of the collector and emitter impedances, both of which change dramatically with frequency. In the emitter circuit, however, the bypass capacitor is represented by the RF model.

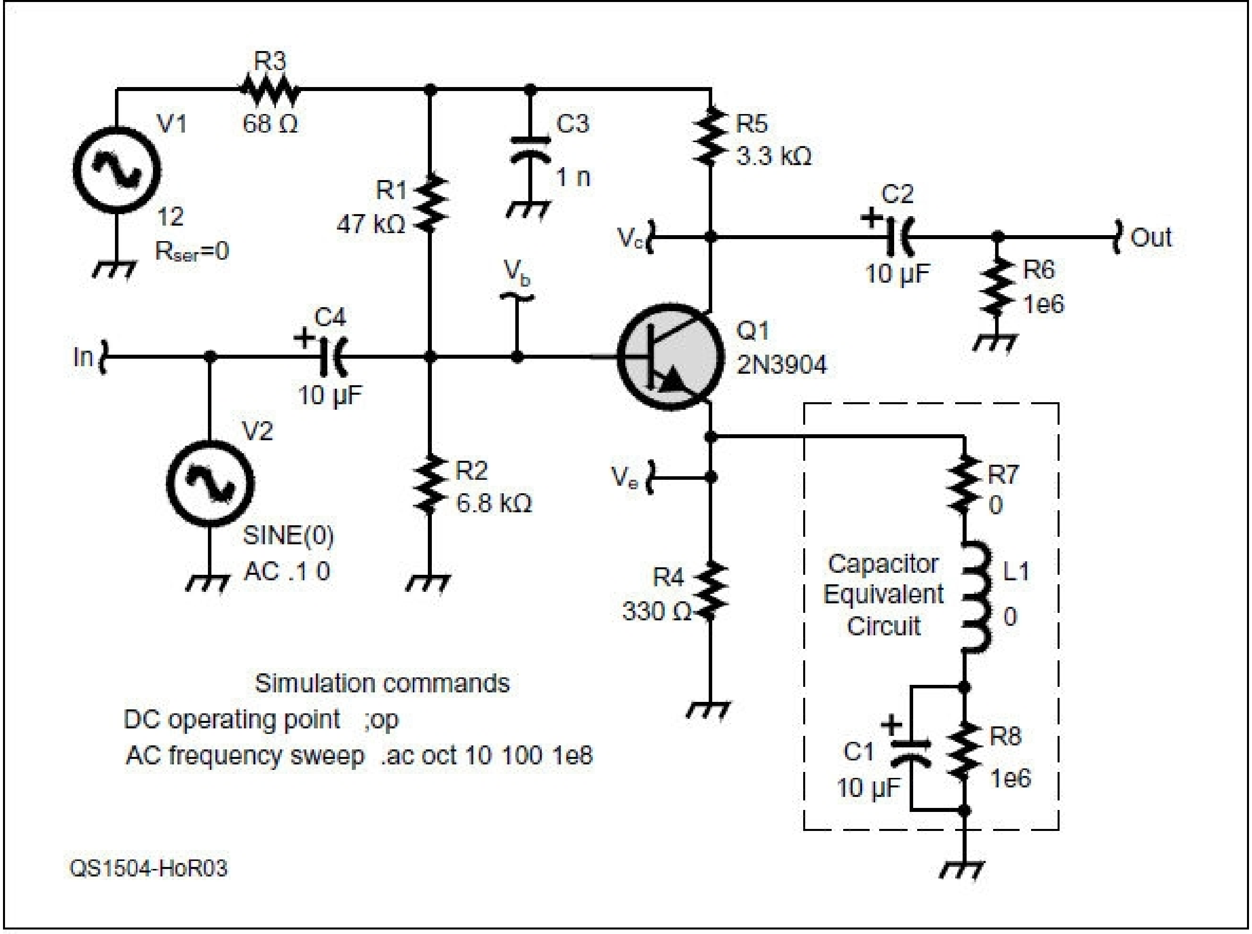

Figure 3 — *LTSpice IV* circuit for demonstrating the effect of parasitic components associated with the emitter bypass capacitor, C1.

If the bypass capacitor is assumed to have a very low impedance, the amplifier's gain is limited by the resistance through the transistors internal emitter resistance, r_e, which depends on dc current flow through the transistor, but is typically around 25 Ω. That means the voltage gain of the circuit will be approximately $A_V \approx -3.3\ \text{k}\Omega \,/\, r_e = -132 = 42$ dB. If the bypass capacitor wasn't there at all, A_V would drop to –3.3 kΩ / 330 Ω = –10 = 20 dB. So if the capacitor parasitic components are present, their effects should be clearly visible. Let's try it!

Start by carefully creating the entire circuit as shown in Figure 3. (The *LTSpice IV* schematic file and two sample response traces are available on the Hands-On Radio web page for the experiment.) The voltage source V1 is a 12 V power supply. The voltage source V2 is the ac sine wave input to the circuit. Set the parasitic component values of R_S (R7) and L_S (L1) to zero and of R_P (R8) to a very high value. Next, perform a dc operating point (DC OP PNT) simulation. Make sure the value of V_C is somewhere near 1⁄2 V1 and the transistor's collector current, Ic(Q1) ≈ 2.5 mA. If the operating point is not correct, the simulation will not produce accurate results.

Create an AC ANALYSIS simulation command with 10 points per decade of frequency between 100 Hz and 100 MHz (1e8 Hz). Run the simulation and click the simulation probe on the OUT and IN ports. In the response window (available on the website), you can see V(in) at the –20 dB level and V(out) near 30 dB, so voltage gain is approximately 48 dB, which is within reason of our previous estimate. The amplifier's bandwidth at the –3 dB points extends from about 1.2 kHz to 9 MHz, more or less.

Now start adding in the parasitic component values and observe the effect on the amplifier's frequency response. For example, changing the value of R_S to 10 Ω drops the gain of the circuit to 43 dB without changing the frequency response very much. Changing the value of L_S to 1 µH shifts the entire frequency response lower and reduces the circuit bandwidth so that the –3 dB points are now at 700 Hz and 3 MHz.

Next, change the bypass capacitor to the non-ideal 1 nF capacitor of Figure 1 and run the simulation from 100 Hz to 1 GHz. At low frequencies where the small capacitance is ineffective, amplifier gain is 20 dB as predicted.
[5]

Above 100 kHz, gain increases to 40 dB, then falls off as transistor gain decreases and series inductance begins to have an effect.

Experiment by changing the coupling capacitors (C2 and C4) to their RF models, changing the transistor type, and adding small inductances in the signal path to simulate lead length. Keep an eye on the circuit's phase response (dashed line on the response graphs) to see how the extra inductances "color" the response.

Notes

1 All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2 **www.cde.com/resources/catalogs/AEappGUIDE.pdf**

3 "Parasitic Inductance of Multilayer Ceramic Capacitors," AVX Corporation, **www.avx.com/docs/techinfo/parasitc.pdf**.

4 *LTSpice IV* is available from Linear Technologies at **www.linear.com/designtools/software**.

5 Remember that the input signal is 0.1 V creating a –20 dB input reference for gain calculations, so an output level of 0 dB represents 20 dB of gain.

Experiment #160 — Transistors at High Frequencies

The first few columns of "Hands-On Radio" covered bipolar junction transistor (BJT) amplifier circuits like the Common Emitter (#1), Emitter Follower (#2), and Common Base (#28).

[1]

For the basic gain and bias equations, transistor behavior was simplified to assume a base-emitter voltage (V_{BE}) of 0.7 V and current gain (β) was assumed to be constant. If you want a more accurate picture of the transistor, particularly at RF, you have to take into account more subtle aspects of how the transistor actually behaves. A good explanation comes from reviewing the models used by simulators — let's zoom in for another look.

Transistor Models

If you look through texts on transistors, you'll find a number of models that describe BJT behavior — there are a half-dozen in *The ARRL Handbook*'s chapters on Analog Basics and RF Techniques alone.

[2]

Why are there so many, and how do you choose the right one? It depends on how the transistor is to be used. Some models are designed for use at dc and some at very high frequencies.

What's the difference between large- and small-signal models? Large-signal models include behavior of the transistor in the non-linear regions, such as cutoff and saturation, and the effects of dc bias. Small-signal models assume the transistor is biased to operate in its linear region. The linear small-signal model is a lot easier to deal with than the large-signal models for RF applications.

The first model you'll encounter is the very simple "beta generator with emitter resistance" model of Figure 1. The current source in this model replicates the current gain of the transistor, β, since collector current, $I_C = \beta I_B$. This model works for ordinary dc circuits like relay drivers or on/off switches. It is too simple to be of much use at RF but it does help understand the basic function of the transistor. In general, if a model does not include capacitance, it is intended to be used at dc and low frequencies.

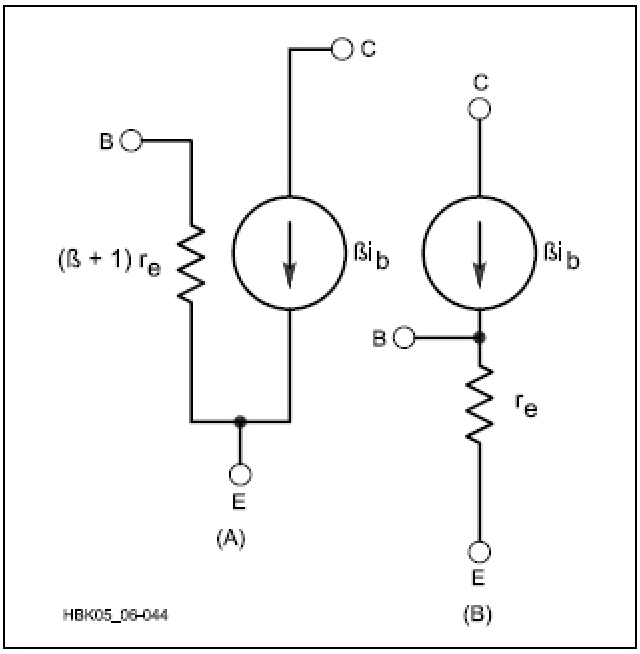

Figure 1 — These two models are equivalent for low-frequency applications. The dynamic emitter resistance, r_e = 26 mV / I_e = 26 / I_e (in mA).

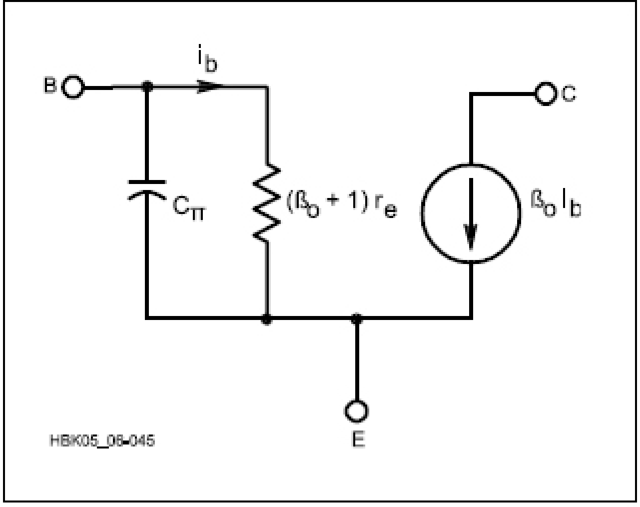

Figure 2 — The hybrid-pi model of the BJT adds capacitor C_π between the base and emitter, causing collector current to roll off at high frequencies. Note that i_b is measured through r_e.

The Hybrid-Pi Model

Commonly used at higher frequencies is the *hybrid-pi* model that adds a single capacitor, C_π, between the base and emitter. BJT construction creates capacitance as discussed below. That capacitance causes high-frequency gain rolloff. Note that the model's base current, i_b, is defined as that going through r_e so that current through C_π does not contribute to the collector current, $\beta_0 i_b$. (β_0 is the transistor's low-frequency current gain.) As frequency increases, the reactance of C_π decreases. This reduces the amount of base current entering the transistor that is multiplied by β_0, causing gain to fall.

The value of r_e is sometimes replaced by h_{ie}, which is one of the h or hybrid parameters and represents ac input impedance. AC current gain is represented by h_{fe} and often replaces β.

3

Dynamic Emitter Resistance

In both of the simple models we've looked at so far, you'll see an internal *dynamic emitter resistance*, r_e, that is dependent on emitter current:

$$r_e = \frac{26mV}{I_e} = \frac{26}{I_e(in\ mA)}$$

This is a characteristic of the transistor itself and should not be confused with an external resistor connected to the emitter which is designated R_E. Dynamic emitter resistance represents the change in base-emitter voltage with the change in emitter current. The change in voltage occurs because of the changing characteristics of the base-emitter junction as emitter current increases.

The numerator value of 26 mV is the room-temperature value of *kT/q*, where k is Boltzmann's constant, T is temperature in Kelvin, and q is the charge of the electron. At very high or low values of temperature or emitter current, the value of 26 mV will no longer be valid, but for most "normal" situations, it's close enough.

The Gummel-Poon Model

In Figure 3, you see the small-signal ac version of the model originally described by Gummel and Poon of Bell Labs in 1970. The Gummel-Poon model improves on the accuracy of the earlier Ebers-Moll model. This is the standard model used by circuit simulators based on *SPICE3*, such as *LTSpice* that we've used in this column previously. With values for all parameters specified by the manufacturer this model can provide accurate results to very high frequencies. The large-signal version of the model also accounts for the variation in current gain with dc current levels.

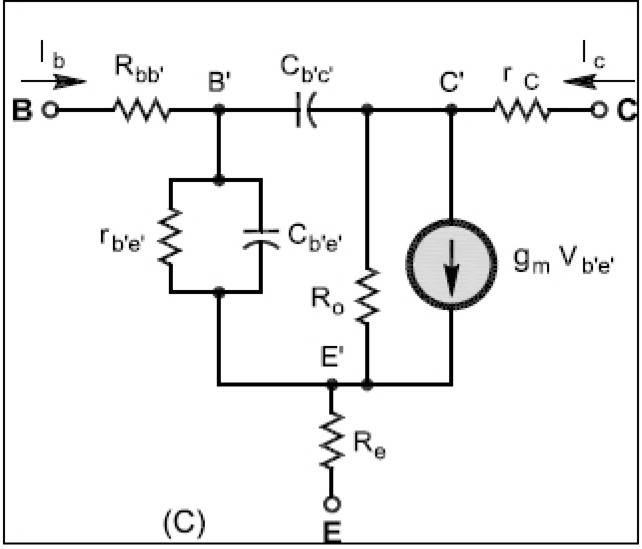

Figure 3 — The Gummel-Poon model is the standard BJT used by *SPICE3* simulation software. The ac small-signal model is shown here.

Figure 4 shows the source of the junction capacitances in the Gummel-Poon model. Figure 4A shows how a planar transistor is actually constructed and Figure 4B gives a more schematic view. Capacitance $C_{B'C'}$, between the base and collector, is created by the reverse-biased base-collector junction. The junction capacitance of the base-emitter junction is labeled C_{je}. The depletion regions at each junction act as the space between two plates of a capacitor formed by the regions of charge on either side of the junction. This is very similar to a reverse-biased varactor diode with the capacitance controlled by bias voltage that moves the regions of charge closer together or farther apart.

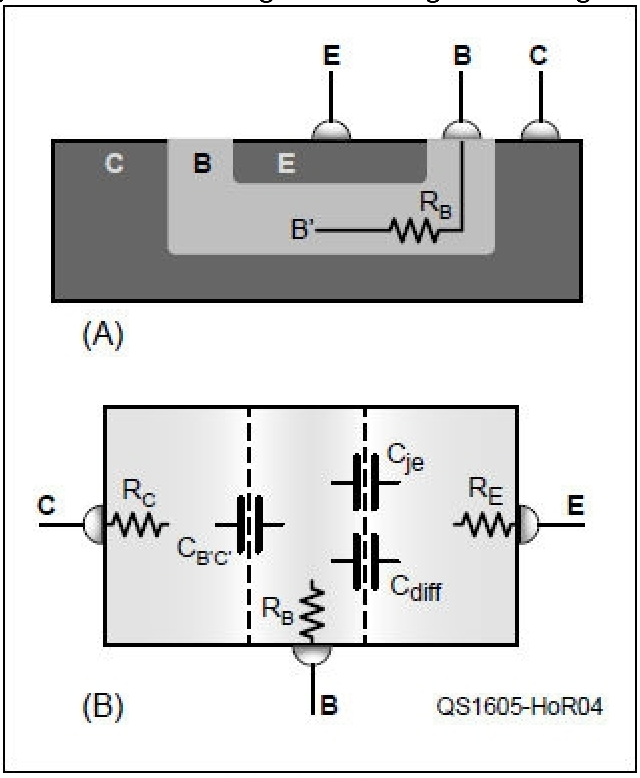

Figure 4 — (A) A typical BJT cross-section showing the three electrodes created by diffusing impurities into the material that serves as a collector. The base resistance R_B models the resistance of the base layer to the point B' underneath the emitter. (B) A schematic representation of the same transistor showing the junction capacitances $C_{B'C'}$ and C_{je} along with the base-emitter diffusion capacitance C_{diff}.

Figure 4B also includes the diffusion capacitance, C_{diff}, which accounts for charge moving through the base-emitter junction when the transistor is biased on. To amplify a signal, that charge must be moved in and out of the junction region, creating the effect of a charge-storing capacitor. C_{je} and C_{diff} act together in parallel and are represented by $C_{B'E'}$, shown in the model and in the hybrid-pi model as C_{π}. (This description oversimplifies the actual behavior of the Gummel-

Poon model and is not meant to be an exact treatment.) Both $C_{B'E'}$ and $C_{B'C'}$ are on the order of a few pF for most common transistors.

Finally, instead of the collector current being modeled with a current gain, the base-emitter junction voltage, $V_{B'E'}$, is multiplied by the transconductance, g_m which has units of amps/volt. This model is more representative of what's actually happening in the transistor and results in more accurate behavior compared to real devices. Next month, we'll experiment with a couple of circuits that demonstrate these effects with real transistors.

Finding Model Values

When models weren't used as widely, the manufacturer data sheets were pretty much the only source for these component values. They aren't measurable from outside the transistor, except indirectly. The place to look is on the data sheet, in the "Small-Signal Characteristics" section, where you can find input and output capacitance. Today's manufacturers usually provide *SPICE* models on their websites so you can get verified models in minutes!

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at

www.arrl.org/hands-on-radio

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2

The ARRL Handbook is available from the ARRL Store at

www.arrl.org/shop/ARRL-Handbook-2016-Softcover-Edition.

3

Hybrid and other two-port parameters are explained in detail by the Talking Electronics tutorial, available at

www.talkingelectronics.com/Download%20eBooks/Principles%20of%20electronics/CH-24.pdf

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Experiment #161 — Transistor Subleties

Last month's column touched on the effects of some transistor characteristics that can be ignored for simple and low-frequency applications, but which become significant in more demanding circuits. This month we're going to tackle two of these effects in different circuits you can build and test: dynamic emitter resistance (introduced last month) and the Miller effect.

Accounting for r_e

I received a nice note from Loren Moline, WA7SKT, who was breadboarding a common-emitter amplifier (see Figure 1) and discovered that the measured gain was less than calculated. In the original "Hands-On Radio" Experiment #1, the equation for voltage gain was given as:

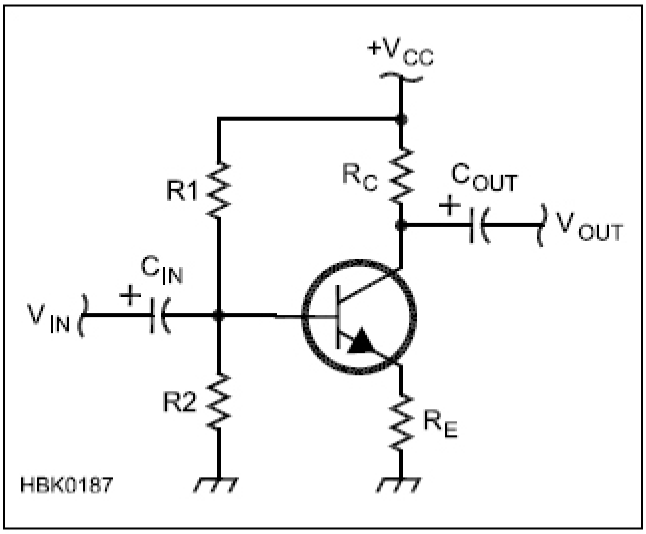

Figure 1 — The common-emitter amplifier.

$$A_V \approx -\frac{R_C}{R_E}$$

This is close enough in many cases. (The minus sign indicated an inverting amplifier.) However, what Loren found is that with R_C = 4.7 kΩ and R_E = 100 Ω, the measured magnitude of gain was only 37 when the calculated gain was 47. Good for Loren not just throwing up his hands and thinking, "I guess that's okay." He dug a little deeper and discovered that the dynamic emitter resistance, r_e, introduced last month, has to be taken into account as well:

$$A_V \approx -\frac{R_C}{R_E + r_e}$$

$$\text{where } r_e = \frac{26\,\text{mV}}{I_e} = \frac{26}{I_e(\text{in mA})}$$

When this equation is used for the 1 mA of quiescent current in Loren's circuit, the calculated gain worked out to a magnitude 37.3, almost the same as the measured value.

Taking the process of accountability one step further, it's also important to take into account the load resistance, R_L in Figure 2. From the ac signal's standpoint, R_L is in parallel with R_C and so the true equation for gain becomes:

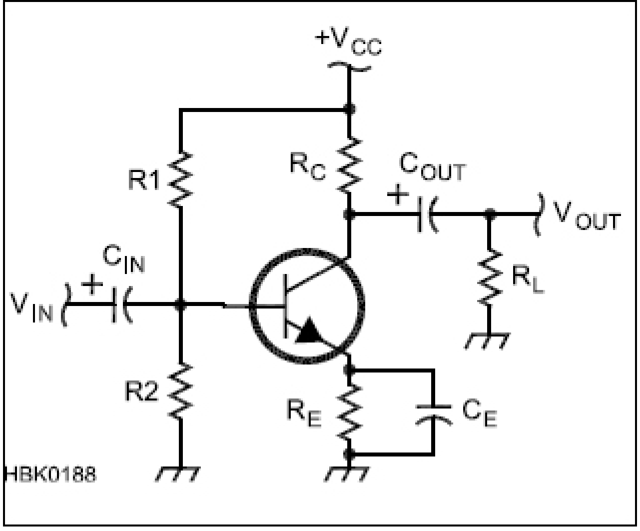

Figure 2 — The common-emitter amplifier with an emitter bypass capacitor.

$$A_V \approx -\frac{R_C \,||\, R_L}{R_E + r_e}$$

How about a little workbench check of all this? Start with the circuit of Figure 1 using Loren's values for R_C and R_E along with 39 kΩ for R1 and 6.8 kΩ for R2. The values are derived in Experiment #1 to set $I_{CQ} \approx 4$ mA. With all of those values in hand, what is the expected magnitude of A_V? (4700 / (100 + 6.5) = 44.1)

Go ahead and build the circuit using an inexpensive 2N3904 or 2N4401 and a 10 µF electrolytic or tantalum capacitor for C_{IN}. Verify that you got the right gain at 10 kHz. You might even want to measure R_C and R_E and use the exact values to calculate gain. (If you lower the frequency of operation, what causes the gain to drop below about 200 Hz? The reactance of C_{IN} begins to become significant — 100 Ω at 160 Hz.)

Next, add the emitter bypass capacitor, C_E, of Figure 2 — a 100 µF component will do nicely. At 10 kHz, a 100 µF capacitor's reactance is how much? (X_C = 0.16 Ω) In parallel with the 100 Ω R_E, the emitter is basically connected to ground. Change R_C to 470 Ω to reduce the circuit's gain. Measure gain and compare it to the calculated value from Eq 2 if $R_E = 0$. (A_V = –470 / 65 = –72.3.) Keep going and add a 10 µF capacitor for C_{OUT} and a 470 Ω resistor for R_L. Calculate gain using Eq 3 ($A_V \approx$ –36.2) and see if your circuit behaves as expected.

Miller Effect

Another internal component that is surprising to the beginning designer trying to get gain at high frequencies is the *Miller capacitance,* C_M. This is partially a real capacitance created by the transistor's internal structure and partially caused by the inverting amplifier configuration. Both bipolar and FETs have some parasitic capacitance from the input (base or gate) to the output (collector or drain for an inverting amplifier or switch circuit). Working with FETs now, that gate-to-drain capacitance is C_{dg}. As Miller discovered in 1920 when working with vacuum tube amplifiers, the gain of the circuit ($-A_V$) also amplified the effect of the capacitance, essentially creating a larger capacitor with a value:

$$C_M = C_{dg}\,(1 + A_V) \qquad \text{[Eq 4]}$$

This *Miller effect* occurs for any inverting amplifier with internal or external capacitance between the input and output: the amplifier's output moving in the opposite direction to the input makes the capacitor look much bigger than it really is. [2]

(This is true for any type of feedback impedance — here we are limiting the discussion to capacitance.)

The Miller effect can be observed fairly easily by turning on an FET while watching the gate voltage. Figure 3 shows an *LTSpiceIV* schematic that we'll use for a circuit simulation that shows the effects of C_M. [3]

The magnitude of this circuit's voltage gain is:

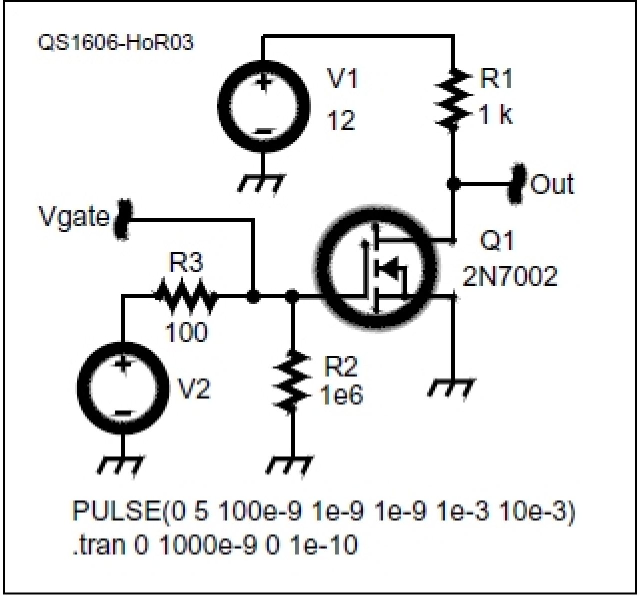

Figure 3 — The *LTSpiceIV* circuit for an FET switch to show how the Miller effect increases the FET input capacitance.

$$A_v = g_m R_d \quad \text{[Eq 5]}$$

For the 2N7002 enhancement-mode MOSFET used here, g_m is specified to be 320 mS. [4]

With an R_d of 1 kΩ, voltage gain would be quite high, A_V = 1000 × 0.32 = 320. So the FET's small reverse transfer capacitance, C_{rss} = 4 pF, is turned into 4 × 320 = 1280 pF by the circuit's voltage gain!

You can see this yourself with a suitable oscilloscope and signal generator. To give you an idea of what you're looking for, though, Figure 4A shows what happens as a fast input pulse is applied to the 2N7002's gate. (The *LTSpiceIV* voltage source values are set to produce a 0 to 5 V pulse beginning at 100 ns with a rise and fall time of 1 ns, a pulse width of 1 ms, and a repetition rate of 10 ms. The simulation's .TRAN parameters configure the simulation to begin at 0 seconds and run for 1 µs with a time step of 0.1 ns.) Voltage traces are taken at the FET drain (V_{out}) and the FET gate (V_{gate}).

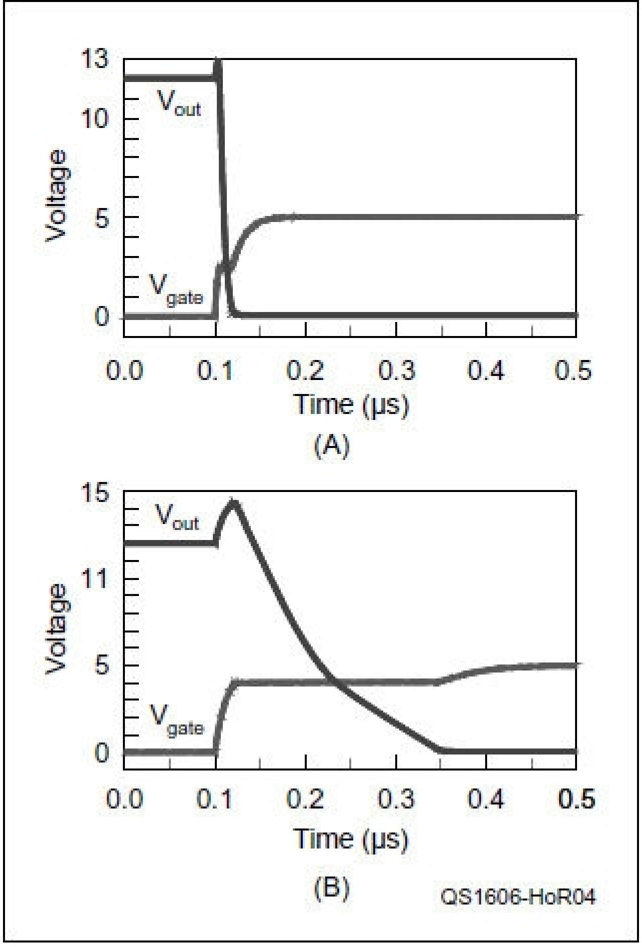

Figure 4 — The gate (blue) and drain (red) waveforms for a 2N7002 MOSFET (A) in the circuit of Figure 3. V_{gs} is held at 2.5 V by the Miller effect while the drain voltage is changing from OFF to ON. (B) shows the same effect on the much larger IRF510 MOSFET with the higher threshold voltage of 4.0 V.

As you can see in Figure 4A, the gate voltage rises quickly but then abruptly flattens out at about 2.5 V and stays there for about 15 ns before rising again toward the final value of 5 V. At the same time, the output voltage, V_{out}, starts with a sharp upward transient due to the input signal feeding through to the output through C_{rss} with the FET still off and gain almost zero.

When the gate voltage, V_{gs}, reaches its threshold voltage of 2.5 V, though, the FET starts to turn on and gain increases to its full value. Now the full Miller effect is in evidence as the rapid increase in capacitance "soaks up" any charge being pumped into the gate by the input pulse. This lasts until the drain voltage reaches zero and is no longer changing. That also ends the Miller effect and V_{gs} resumes its climb to the final value of 5 V.

The 2N7002 is a very small MOSFET designed for light-duty switching. What happens when a larger FET is used? Change the 2N7002 to an IRF510 (right-click on the FET symbol and select PICK NEW MOSFET) and re-run the simulation after changing the stop time in the simulation to 1 µs. Wow! Figure 4B shows that the much larger device with at least 20 pF of capacitance experiences a much larger Miller effect, causing V_{gs} to stay "flat" for 350 nsec. To turn this much larger device on and off quickly — either as a switch or an RF amplifier — the gate drive circuit has to be able to drive a much larger load than for the 2N7002. This is why special gate driver ICs are often used with larger FETs.

You can build this circuit on your workbench if you have a suitably fast oscilloscope and a signal generator capable of producing pulses with fast rise times. I recommend using the "dead bug" or "Manhattan-style" construction techniques presented in Experiment #90 to minimize the effects of stray capacitance and inductance present on solderless breadboards. A 5 V logic gate can serve as a fast pulse source, as well. Use a Schmitt trigger (74HC14 or equivalent) to square up a pulse with longer rise and fall times.

When you start using your scope, remember that you are looking for events that occur in a short period of time, immediately after the FET gate voltage begins to rise. Trigger on the leading edge of the drive pulse — either using a dc threshold or a positive-going slope. If you're using an IC to generate the pulse, trigger off the input to the gate to create a short delay between triggering the scope and the flattening out of the waveform. Set time/division on the horizontal sweep to be something approximately like the simulated waveform's time scale. Then zoom in and look for that curious plateau as suddenly, "It's Miller time!"

Notes

[1]All previous "Hands-On Radio" experiments are available to ARRL members at
www.arrl.org/hands-on-radio
.

[2]
N1090-D "Understanding and Predicting Power MOSFET Switching Behavior," OnSemi,
www.onsemi.com/pub_link/Collateral/AN1090-D.PDF
.

[3]
LTSpiceIV is available for downloading from the Linear Technology website as described in "Hands-On Radio" Experiments 83 – 86 which also discuss how to use the simulator. The circuit schematic file for this experiment is available on the "Hands-On Radio" website's page for this experiment.

[4]
g_m is transconductance in amps/volt with units of siemens (S).

Experiment #169 — Gas Discharge Tubes

In experiment #121, I covered transient protection for the various insults that occur in vehicles and other power systems.

[1]

Protection from a lightning strike, either direct or indirect, requires special, sterner measures across your station, though. First, let's back up and understand what we're protecting against.

Lightning Energy

A typical lightning strike is composed of three to four impulses per strike. Peak current for the first pulse averages around 18 kA (98% of the strikes fall between 3 kA to 140 kA at their peak). For the second and subsequent impulses, the current will be about half the initial peak. The typical interval between impulses is approximately 50 ms. Figure 1 shows a typical impulse, referred to as the *IEEE 8/20 model waveform*. Remember, this is an *average*, and half of lightning strikes have more energy than this waveform!

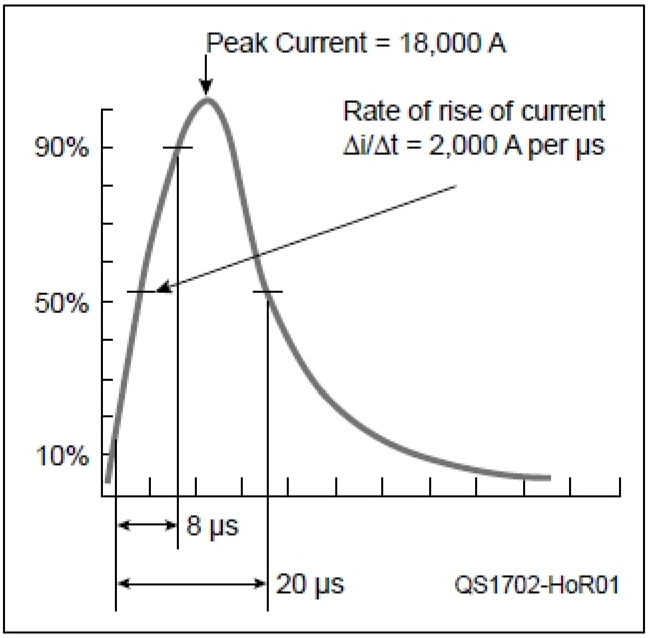

Figure 1 — The IEEE 8/20 model waveform for a typical lightning pulse. This is an average, so many lightning strikes are bigger than the waveform shown here.

Voltages created by this current pulse can be enormous and depend on the resistance (R) and inductance (L) through which the current flows. According to Faraday's Law, the faster that current changes ($\Delta i / \Delta t$, where the Greek letter delta, Δ, means "the change in") through an inductance, the higher the voltage that is created. Higher current through a resistance also means higher voltage, per Ohm's Law:

$V = I \times R + L \times \Delta i/\Delta t$

While the majority of lightning's energy is pulsed dc, there is a substantial amount of RF created by the fast rise time of the pulses. A typical strike rise-time of 1.8 µs translates into a radiated RF signal peaking at 139 kHz. Rise times can vary from a very fast 0.25 µS to a very slow 12 µS, yielding an RF range from 1 MHz down to 20 kHz. When lightning "attaches" to the *air terminal* (where the leader channel reaches the grounded object), the rise time for current can be as short as 10 nS. The result is that lightning's energy extends upward in frequency to 10 MHz and higher.

Clamping a Lightning Transient

What causes the damage when lightning strikes — voltage or current? The answer is both! That's a lot of energy, and a lot of spectrum. The combination of high energy and wide spectrum makes protecting equipment a difficult challenge.

Any protection device that depends on internal resistance to dissipate the energy as heat will still produce high voltages even from indirect strikes. Putting inductance in series with the transient might help at low frequencies, but is incompatible with RF receiving and transmitting equipment.

While disconnecting equipment is nearly foolproof, it's not always possible or practical. We need something that will divert the transient energy away from equipment inputs or outputs. Whatever we employ, it has to be "invisible" until it is triggered. Then it must activate quickly enough not to pass damaging energy, and to remain activated until the transient event is over, then return to invisibility until the next event. It must require no power to operate and be insensitive to RF.

The only component with the characteristics we need and that can withstand repeated transients is the *gas discharge tube* or *GDT*.

GDT Basics

Several examples of GDTs used by amateurs are seen in Figure 2. The photograph of a GDT shows the shape and placement of the metal electrodes. They are typically made of tungsten or some other tough metal that can withstand repeated short arcs. The schematic symbol includes a small dot next to the electrodes. This indicates the presence of gas. A proprietary combination of gases is used, along with electrode spacing, to control the *breakdown* voltage. Once an arc is created, the resistance between the electrodes is quite small until the arc current falls and is *quenched*. The GDT is then inert until the next transient.

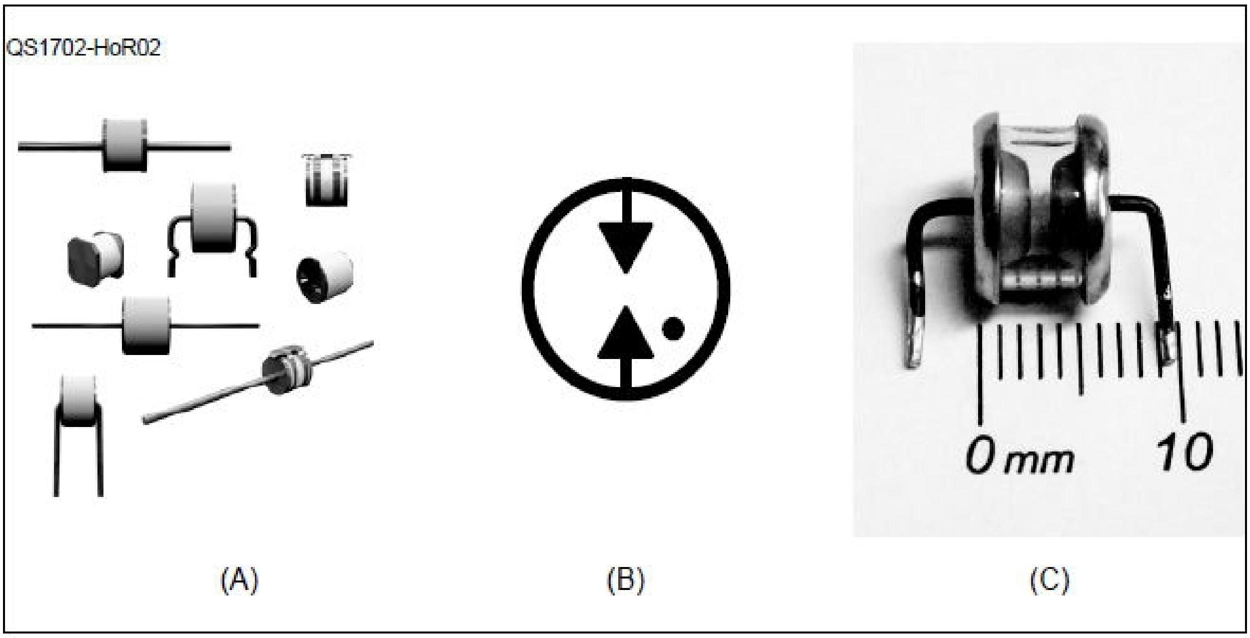

Figure 2 — Typical gas discharge tube components (A), the GDT schematic symbol (B), and a close-up photograph showing the electrode shape (C). [Photo courtesy of Ulfbastel — own work, public domain, **commons.wikimedia.org/w/index.php?curid=2768406**]

The GDT is a type of *crowbar* protection. As stated in the Littelfuse GDT catalog:

A crowbar device limits the energy delivered to the protected circuit by abruptly changing from a high-impedance state to a low-impedance state in response to an elevated voltage level. Having been subjected to a sufficient voltage level, the crowbar begins to conduct. While conducting, the voltage across the crowbar remains quite low…and thus, the majority of the transient's power is dissipated in the circuit's resistive elements and not in the protected circuit nor the crowbar itself. This allows the crowbar to be able to withstand and protect loads from higher voltages and/or higher current levels for a greater duration of time than clamping devices.

[2]

A similar type of crowbar device is a *spark gap* or *air gap protector*. Spark gaps can be easily made from common materials, and ARRL publications have many examples, including the use of automotive spark plugs for open-wire feed lines.

When most amateur equipment used vacuum tubes, a spark gap might have been enough protection. After all, the tubes themselves operated with voltages of several to many hundreds of volts and could withstand even short arc-overs. Transistors and integrated circuits (ICs) are much less forgiving. Once breakover voltage is reached and the arc is established, a typical GDT limits the voltage across the electrodes to 15 V or so, which is well within the safe voltage range for nearly all radio equipment.

Applying GDT Protection

As you can see in Figure 2, there are lots of different styles of GDT packages. The *leaded* and *cartridge* packages are the most common in amateur stations. GDTs are rated by breakdown voltage and energy handling ability. The Littelfuse CG-series is suitable for most amateur applications.

For protecting dc or low-frequency control and telephone lines, the leaded package is the most common. These components can be connected to screw terminals or barrier strips. For rotator control lines, measure the root mean square (RMS) voltage from the control box and multiply by 1.414 to get peak voltage to ground (2.8 for peak-to-peak). Add another 10 – 20% to provide a margin and determine the required breakdown voltage. Remember that telephone line ringing voltage can be as high as 150 V_{PK-PK}.

At RF and especially in circuits that must withstand transmitter output voltages, a cartridge-style GDT is used. Shown in Figure 3 is a set of four lightning protectors with UHF connectors for attaching to feed lines. The GDT is connected between the shield and center conductor of the feed line, with the protector body also mounted on a grounded metal mounting bracket. Several *QST* advertisers offer these protectors for either low-power for full legal-limit operation.

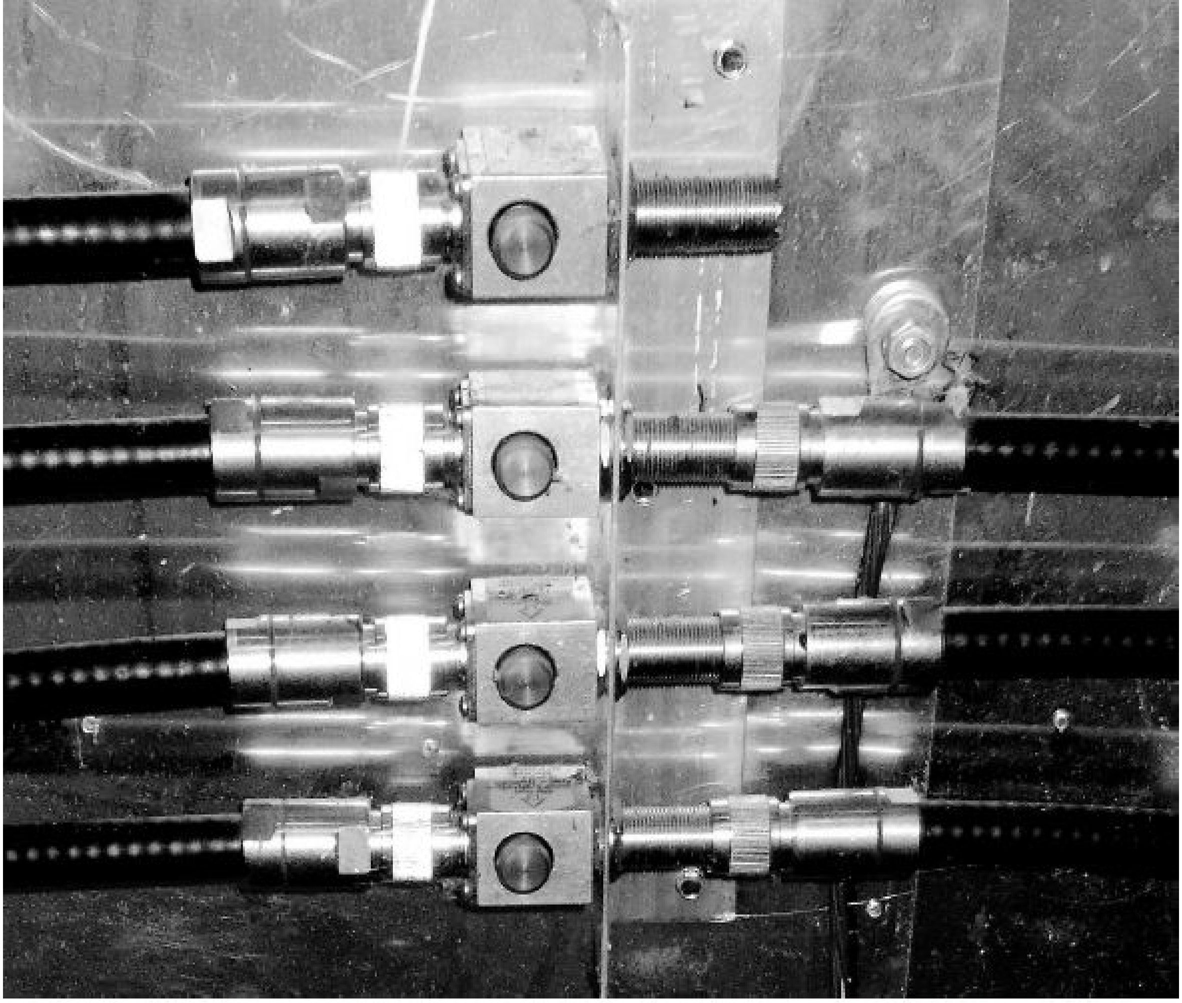

Figure 3 — GDT-based lightning protectors used in the author's station. The GDT cartridge is accessible via the brass screw cap. Each protector is mounted on an aluminum angle bracket that is bolted to the grounded metal sheet.

There are two caveats for using these protectors. You still need a good ground system for full lightning protection, including a lightning protection plan and bonding for all protected equipment. (See the 2002 series of *QST* articles by Ron Block, KB2UYT, at **www.arrl.org/radio-technology-topics** under "Safety.")

You should also be aware that there are two basic versions of the commercial protectors: one that passes dc current and one that does not. If you plan on using a remote device, such as a preamp or antenna switch that requires feed line power, be sure not to purchase the dc blocking version. A call to the vendor will help you make the right choice.

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio.**

2

Littelfuse Inc., *Gas Discharge Tube (GDT) — Product Catalog and Design Guide*, 2015, **www.littelfuse.com**.

Experiment #172 — Wire Characteristics at RF

RF has a way of changing the behavior of even a lowly piece of wire, and this column presents three significant examples. You'll never look at a piece of hookup wire the same way again.

Inductance and Parasitic Effects

Any time a current flows — whether through a wire or not — a magnetic field is created. If the current is flowing through a wire, the familiar "right-hand rule" shows how the field curls around it.
[1]

This field, in turn, also interacts with the current in the wire, creating an *electro-motive force* (voltage). The interaction of current and magnetic field is called *inductance*, because changes in the field "induce" changes in voltage. The relationship is described by Lenz's Law:

$V = -L \Delta i/\Delta t$

where L is the inductance, and Di/Δt is the rate of change in the current, often written in the language of calculus as *di/dt*. The value of L (in units of henries, H) is determined by the size and shape of the conductor. The minus sign shows that the voltage caused by the change in current opposes that change.

Calculating the exact inductance of any piece of wire turns out to be surprisingly complicated, because it depends on geometry so strongly. Even for straight wires, inductance changes non-linearly with length and thickness. For example, 10 inches of copper wire has an inductance of about 0.32 µH for #20 AWG, 0.12 µH for #10 AWG, and 0.11 µH for 0.5-inch rod. That #20 AWG wire has about 0.13 µH at a length of 5 inches and 0.02 µH for a 1-inch piece. Loop and coil inductances are even more complex. (*The ARRL Handbook* has charts and formulas for inductance of straight wires and single-layer coils.
[2]
)

You can measure inductance with an "L-C" meter, of course, but for the true RF experience, SWR analyzers, such as the MFJ-269 in Figure 1, can be used. You can see my simple component test fixture connected to the analyzer — a BNC-to-binding post adapter and a pair of alligator clips with fork terminals soldered to them. Attach the component to be tested to the clip leads and set the analyzer to measure inductance — a 100 Ω resistor exhibited 18 nH of inductance at about 64 MHz, for example. (These are not precision measurements — many factors affect the inductance value.)

Figure 1 — A simple test fixture of alligator clips holding a 100 Ω resistor. The MFJ-269 antenna analyzer is showing an inductance of 18 nH (0.018 μH) at 64 MHz.

Try a piece of wire about the same length as a resistor. (Resistance affects the analyzer measurement.) Follow your analyzer's instructions and note the inductance. Without changing the frequency, shorten the wire and observe the change. Replace the wire with a longer piece and bend it into different shapes or even make a small coil. Bring metallic objects close to the wire — some magnetic (steel) and some not (aluminum) to see their effect. Touch the wire with a finger. Anything in the wire's magnetic field will affect its inductance. Be aware of this sensitivity to nearby objects when building RF equipment!

Skin Effect and Resistive Loss

You may have heard the term *skin effect*. Not only do currents create external magnetic fields around a wire, but currents also create magnetic fields *inside* it! Forces from the interaction between fields and ac currents (similar to what makes a motor turn) move current from the interior to a thin layer near the surface. The higher the frequency, the thinner the layer in which most of the current flows. Inversely proportional to the square of frequency, *f*, the *skin depth* is:

$$\delta = \frac{1}{\sqrt{\pi \mu f \sigma}}$$

where μ is the conductor's permeability, and σ is the conductivity. Similar to RC and RL time constants, δ is the depth at which current falls to 1/*e* (about 37%) of its value at the surface.

Figure 2 shows skin depth versus frequency for several different types of metals from below ac power frequencies all the way to 1 MHz. For a copper wire, most of the current is flowing in a layer about 2⁄3 millimeters thick at 10 kHz and 1⁄5 millimeters at 100 kHz. Restricting current to this thin layer increases the resistance quite a bit. A good rule of thumb for copper and aluminum wire is that skin effect begins to have a significant effect on wire resistance above the frequency:

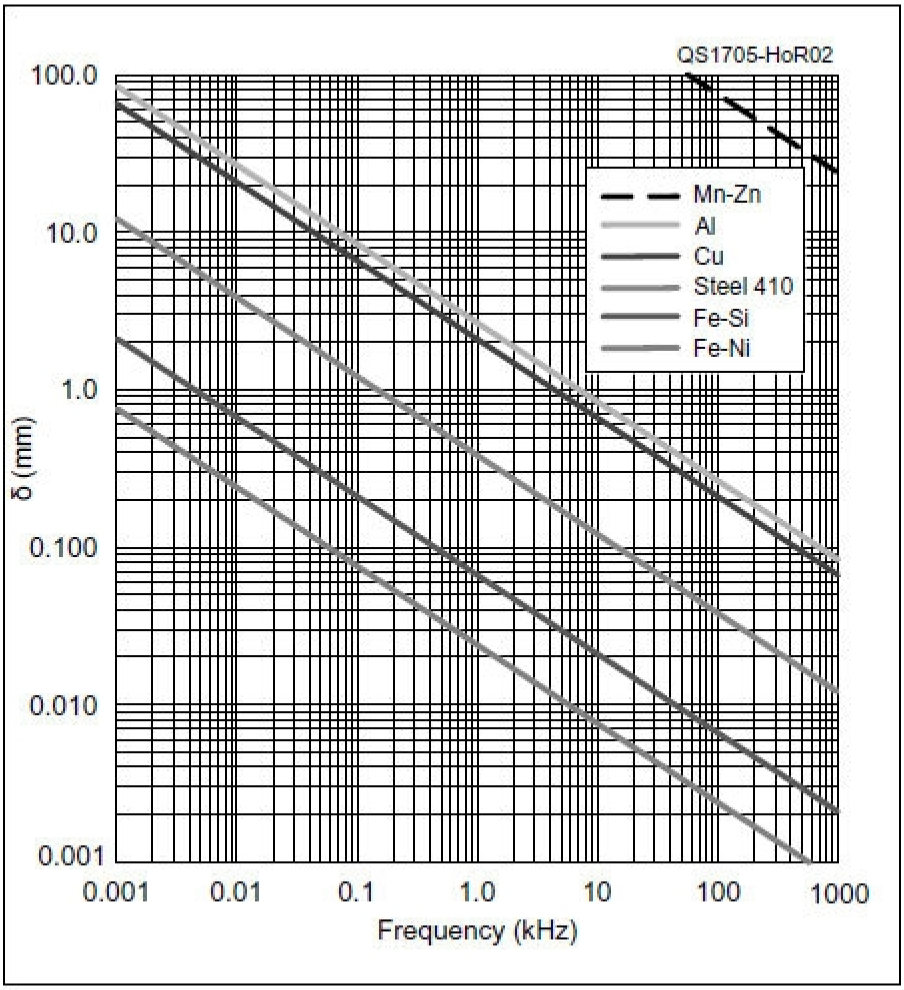

Figure 2 — Skin depth for various types of conductors from 1 Hz to 1 MHz.

$$f(\text{MHz}) = \frac{124}{d^2}$$

where *d* is the wire diameter in mils (1 mil = 0.001 inch). Above that frequency, resistance increases 3.2 times for each decade (10 times) of frequency.

Let's work out an example for #14 AWG solid copper wire at 28 MHz. The wire has a diameter of 64.1 mils and 2.52 Ω of resistance per 1,000 feet. At dc, 10 feet of this wire would have a resistance of 0.025 Ω. The frequency at which the skin effect becomes significant is 124 / $(64.1)^2$ = 30 kHz. 28 MHz / 30 kHz = 933 ≈ 3 decades (10 × 10 × 10), so the resistance increases to approximately 0.025 × 3.2 × 3.2 × 3.2 = 0.81 Ω. Would that be significant? Let's say you wound a coil for a 10-meter band-pass filter from this wire. If the current in the coil was 10 A rms, the coil would dissipate I^2R = 10^2 × 0.81 Ω or 81 W. This is one reason why large diameter wire or tubing is used in transmitting filters.

Effects of Insulation

What if the material next to the thin layer of metal carrying the RF current is not air? Even if the material is an insulator, we can still expect it to affect an RF current if its *dielectric constant* is different from air. Also known as *relative permittivity*, the dielectric constant tells us how much a material affects an electric field relative to that of a vacuum or dry air. (The relative dielectric constant for air is 1.0.)

Insulation's higher dielectric constant increases capacitance along the wire, slowing the RF wave. This is best illustrated by using a model.

I created an *EZNEC* model (www.eznec.com) of a 2-meter dipole made with #14 AWG copper wire, 191⁄4 inches long on each side of the feed point. In free space and using bare wire, the resonant frequency was 146.6 MHz. I then changed to THHN household ac wire — a very popular antenna-building material. THHN insulation is mostly PVC with a very thin outer layer of clear nylon. For #14 AWG wire, the insulation thickness is approximately 15 mils. Without changing the wire length, the insulated dipole's resonant frequency dropped by 2.6% to 142.9 MHz — completely out of the 2-meter band. Table 1 shows resonant frequencies with other types of insulation.[3]

You can observe this effect yourself by making a simple dipole or ground plane of solid insulated wire, measuring the resonant frequency, then removing the insulation entirely or in parts. You'll find that the average effect from most types of insulation is to lengthen the wire electrically by 2 – 3%.

Table 1
Effect of Insulation on Dipole Resonance

#12 copper wire, 19.25 inches on each side of feed point

Material	Dielectric Constant	Thickness (mils)	Res. Freq. (MHz)	Change (%)
None	1.0	0	146.6	—
THHN (PVC)	3.5	15	142.9	–2.6
Teflon (PTFE)	2.1	15	143.9	–1.8
Nylon	4.1	15	142.7	–2.7
HDPE	2.26	15	143.8	–1.9
HDPE	2.26	98*	135.4	–7.6

*This is the thickness of RG-213-type center insulation

Notes

1

The right-hand rule is explained at
www.khanacademy.org/test-prep/mcat/physical-processes/magnetism-mcat/a/using-the-right-hand-rule
.

2

Available from your ARRL dealer, or from the ARRL Store, ARRL Item no. 0628. Telephone toll-free in the US 888-277-5289, or 860-594-0355, fax 860-594-0303;
www.arrl.org/shop
;
pubsales@arrl.org
.

3

www.rfcafe.com/references/electrical/dielectric-constants-strengths.htm

Experiment #173 — The PC Trace at RF

After last month's column, I'm sure you have a new respect for that piece of wire carrying your RF. The same considerations apply to PC boards, too — and more! For high-frequency digital signals and our microwave RF, they can behave like inductors and transmission lines. Let's take a closer look.

Inductance of Straight Traces

Just like the straight wire inductance we discussed last month, you can calculate the inductance of a PC trace,
[1]

but there's a catch: the inductance of the whole connection also depends on the return path for the current. This is because the fields from the current in the return path will interact with those from the current in the main trace. Unless you know the return path, the inductance is unknown as well.

Dr. Howard Johnson gives a formula for inductance of a trace that runs parallel to one or two PC board ground-plane layers: (L per inch) = Z_0 × (delay per inch). You have to know the delay per inch, and Z_0, the characteristic impedance of the transmission line formed by the trace and the ground plane.
[2]

Thankfully, you usually don't need to know the inductance of the PC board trace. Unless, that is, you want to make an inductor.

Making Inductors with PC Board Traces

If you work on VHF/UHF/microwave gear, you'll have seen PC board inductors. They are usually *planar spirals*, like in Figure 1, in the shape of octagons, hexagons, squares, or circles. The inductance is calculated between the two ends, but the exact value depends on where the inner end connection is routed and the arrangement of ground-plane layers.

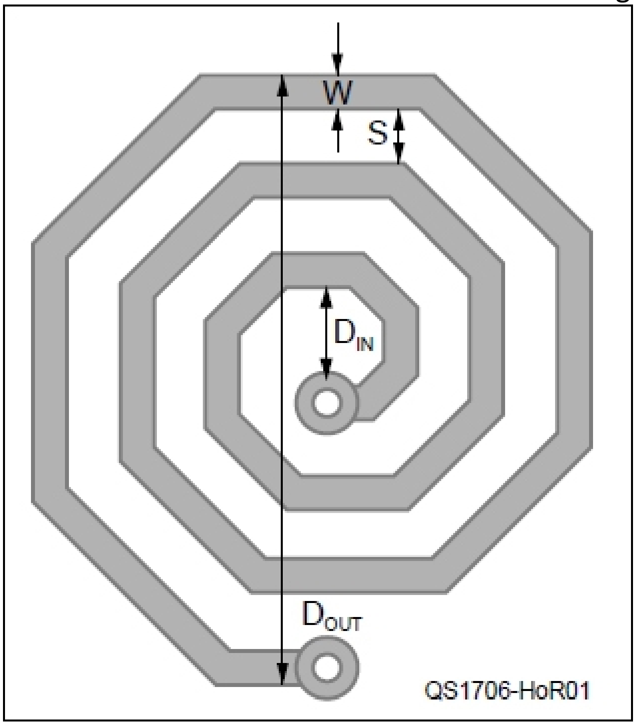

Figure 1 — A spiral planar inductor. The coil diameter (OD – D_{OUT}, ID – D_{IN}), number of turns, trace width (W), and spacing between them (S) determine the inductance. Conductor thickness is relatively unimportant.

There is an online calculator for these inductors at the All About Electronics website.
[3]

A five-turn octagonal coil made with 20-mil traces, spaced twice the trace width, is 0.78 inches across and has approximately 408 nH of inductance. (Three different methods of calculation are provided.) As long as you aren't expecting precision, these coils are very inexpensive and easy to fabricate. Some PC layout software can create this type of inductor automatically, as well.

PC Board Transmission Lines

The dependence on return path may sound like a complicating factor, but the PC board material itself can be used to create a transmission line. There are several variations in which the PC trace forms one of the conductors and ground-plane layers form the other. These are summarized in Figure 2, where ε_r is the dielectric constant of the PC board material. (FR4 is the most common material at and above VHF.)

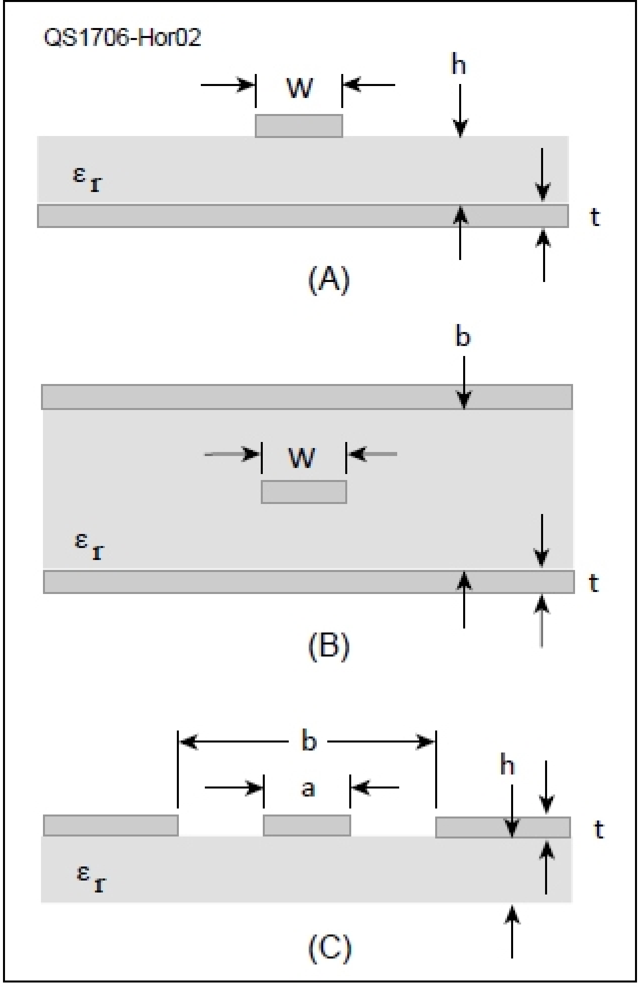

Figure 2 — Types of PC transmission line: microstrip (A), stripline (B), and coplanar waveguide (C). Dimensions shown are used by online calculators to determine the line's characteristic impedance. The PCB material's relative permittivity is ε_r.

Microstrip (Figure 2A) is the most common of the PC transmission lines, consisting of an isolated trace above a ground plane. *Stripline* (Figure 2B) is also common in multilayer boards with the PC trace embedded in the PC board material and centered between two ground-plane layers. *Offset stripline* (not shown) is a variation of stripline in which the PC trace is not centered between the ground-plane layers. Finally, at microwave frequencies, *coplanar waveguide* becomes feasible, as shown in Figure 2C.

In microstrip and stripline, the RF energy is mostly (but not completely) confined to the region between the large surface of the PC trace and the ground plane. Current is spread across the surface of the PC trace at a depth determined by the skin effect.

In contrast, the RF energy in coplanar waveguide is contained between the edges of the PC trace and the edges of the adjacent ground plane. The middle surfaces of the PC trace carry little, if any, current. This increases resistive losses because the current is concentrated in a smaller region, but the waves travel mostly in air and, therefore, have lower losses. This becomes an important tradeoff at microwave frequencies.

The math to calculate Z_0 of these PC transmission lines is ferocious. Because most designs work with 50 Ω impedances, combinations of common copper foil thicknesses, trace widths, and board layer thicknesses have been calculated to produce 50 Ω. Several are shown in Table 1. (For the truly interested reader, see Wadell's book in the notes.
[4]
) The free program *AppCAD* (
www.hp.woodshot.com
) handles many of these calculations for you, along with S-parameters and balun calculations.

Table 1
50 Ω Transmission Line Dimensions (From Maxim Tutorial 5100, see Note 6)

Type of Line	Dielectric (ε_r)	Layer Thickness in mils (mm)	Center Conductor in mils (mm)	Gap	Characteristic Impedance (Ω)
Microstrip	Prepreg (3.8)	6 (0.152)	11.5 (0.292)	N/A	50.3
		10 (0.254)	20 (0.508)		50.0
Stripline	FR4 (4.5)	12 (0.305)	3.7 (0.094)	N/A	50.0
Coplanar WG	Prepreg (3.8)	6 (0.152)	14 (0.356)	20 (0.508)	49.7

Skin Effect and Ampacity

The skin effect, discussed last month for round wire, has an interesting effect on isolated flat traces and strap — not only does it cause current to concentrate at the surface, as in Figure 3A, but also at the edges of flat conductors, as shown in Figure 3B. At lower frequencies, this isn't such a big concern, but at 1 MHz, skin depth in copper is down to 65 µm (0.0026 inches) and 1⁄10 of that at 100 MHz (6.5 µm and 0.00026 inches). With the typical "1 ounce" coating of copper (1 ounce of copper per square foot) on a PC board having a thickness of 1.4 mils (1 mil = .001 inch, 1.4 mil = 0.035 millimeter), the skin effect begins to increase trace resistance substantially above about 5 MHz. This matters both in HF power circuitry and in circuits operating at and above VHF.

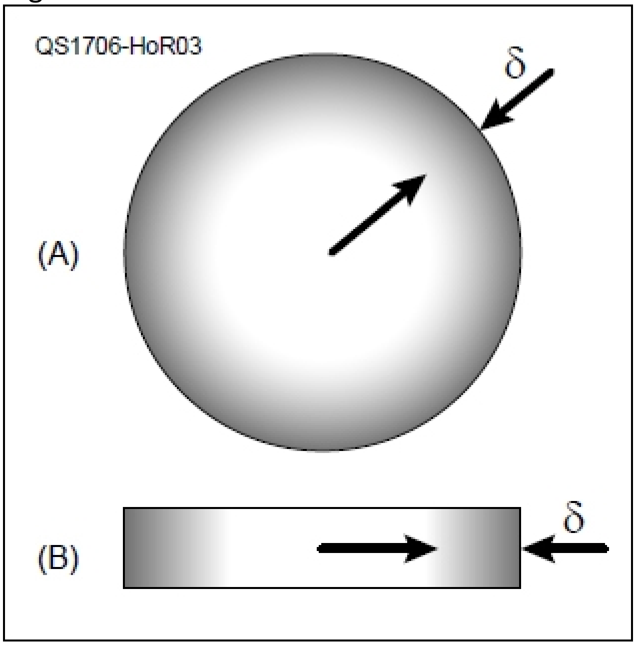

Figure 3 — Compared to the round wire (A), the skin effect in a flat conductor (B) causes current to flow along its edges and not equally distributed across the flat surfaces. The skin depth is shown as δ.

The lesson to learn here is to be careful when using PC boards for RF power circuits. As for coplanar waveguide, the width of the trace becomes less important than the thickness. Because most of us do not have access to electromagnetic simulators and the exact configuration of traces is hard to model, we have to be aware of the effect and conduct high-power tests to look for hot spots, etc.

Clarification about MOVs in Power Strips

My caution in the April column about the use of metal-oxide varistors (MOVs) referred to older strips. Since 2009, an updated UL 1449 standard recognized the shortfalls of using MOVs this way and set new safety standards. (The standard was updated again in 2016.) If your surge-protected appliance or power strip meets these new standards, it's safe to use. Older strips should be replaced, however.

RF PCB Design

We've just scratched the surface of RF PCB layout. As you might imagine, there is a lot of hard-won knowledge you can apply as the megahertz turn into gigahertz. A few guidelines are listed in the notes.[5], [6]

You can collect or bookmark these and other references — don't be afraid to turn up the frequency and have fun with RF!

Notes

1 A 20-mil trace of 1-ounce copper has an inductance of approximately 25.7 nH/inch (chemandy.com/calculators/flat-wire-inductor-calculator.htm).

2 H. Johnson and M. Graham, *High-Speed Digital Design: A Handbook of Black Magic*, Prentice Hall, 1993.

3 www.circuits.dk/calculator_planar_coil_inductor.htm

4
B. Wadell, *Transmission Line Design Handbook*, Artech House, Inc.
5
C. Bourde, J. Fuller, S. Long, “RF Prototyping Techniques,” UC Santa Barbara, 1998 (
home.sandiego.edu/~ekim/otherjunk/rf_proto.pdf
).
6
M. Bailey, “Tutorial 5100 — General Layout Guidelines for RF and Mixed-Signal PCBs,” Maxim Integrated (
www.maximintegrated.com/en/app-notes/index.mvp/id/5100
).

Tests and Test Equipment

Experiment #140 — RF Measuring Tools

This column presents several simple circuits you can build yourself to satisfy a common ham radio need — detecting and measuring RF voltages and currents. We'll start by figuring out how to get access to those RF signals, then measure the value of the voltage or current, and finally turn it into a value measured in decibels. All of the tools are described further in the articles listed in the references at the end of this article.

RF Samplers

How do you tap into a feed line carrying RF without disturbing it (much)? Simple — just get a T connector and hook up your test instrument with a jumper, right? Well, no. If that little bit of cable becomes more than a few percent of an electrical wavelength long, the combination of the terminating impedance and transmission line effects can seriously disturb signals in the main feed line.

The solution is to use a sampler that extracts a very small amount of RF power while not affecting the main feed line very much. Two common methods are used to do this, as shown in Figure 1; the toroidal transformer and the capacitive coupler. [1], [2], [3]

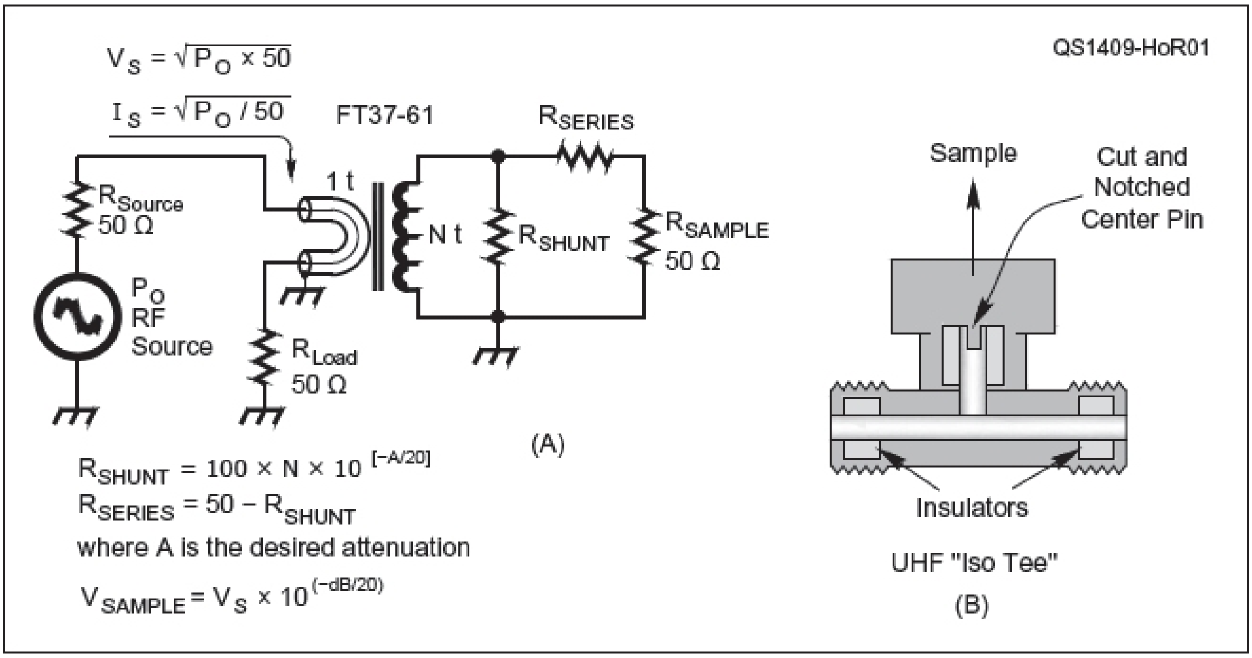

Figure 1 — Two common RF samplers. A one-turn primary toroidal transformer senses RF current at (A). The sampler at (B) senses RF voltages by using the capacitance of a modified T coaxial adaptor.

For the toroidal transformer, pick the desired attenuation, *A*, in dB and the turns ratio, *N*. Then solve for R_{SHUNT} and R_{SERIES}. Assuming the sampler's main line is connected to a 50 Ω load with V_S across it, and R_{SAMPLE} in Figure 1A is 50 Ω, the output voltage, V_{SAMPLE}, across R_{SAMPLE} will be *A* dB below V_S.

The capacitive sampler, known as the *Isotee*, is not designed to have a specific attenuation, but rather to simply pick off a small amount of RF from a main feed line. The Isotee is used to provide a signal for a spectrum analyzer or frequency counter. Be sure to label the coupler clearly to avoid using it as a regular T connector and then wondering why you have a bad connection!

For both the toroidal and capacitive samplers, remember to take into account their frequency response when making comparisons between signals of different frequencies. For example, if you are interested in determining harmonic content of a signal relative to the fundamental, the sampler's response should be consistent well beyond the frequency of the harmonic. You can determine the sampler's frequency response by measuring the sampler's response with a lab-quality instrument.

RF Peak Detector

Without an oscilloscope handy or a true-RMS RF voltmeter, the most useful RF signal amplitude measuring tool is a peak detector. Experiment #53 presented both envelope detector (for receiving AM signals) and peak detector circuits. Figure 2 shows an RF peak detector with both a direct 50 Ω input and an input with 40 dB of attenuation.

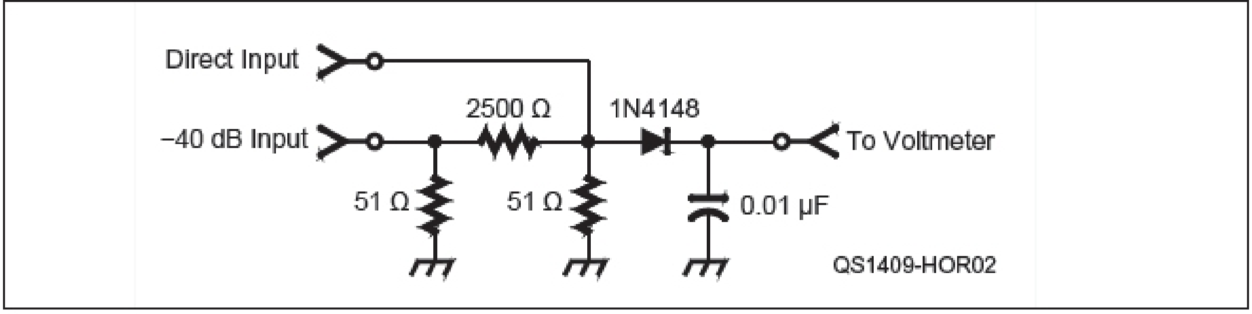

Figure 2 — This peak detector is useful for measuring RF power levels by creating a dc voltage that can be measured with a DVM or other voltmeter. Be sure the 51 Ω resistors are sufficiently rated for the power level to be used.

The meter's response can be calibrated by measuring the input voltage with an oscilloscope or a lab-quality voltmeter. The input power can be calculated as:

$$_{watts} \quad \frac{(\;_{dc} \quad 0.6)}{\times}$$

Where R is the value of the resistors connected to the inputs (51 Ω in the figure), V_{dc} is the dc voltage measured by the voltmeter, and 0.6 represents the diode's forward voltage drop with 1 mA of current. (This equation is from section 7.3 of *Experimental Methods in RF Design*.

[4]

) At lower power and diode current, the voltage drop will be lower, causing a non-linear calibration curve!

A 1N4152 diode can be substituted with a faster response time to extend frequency response. A 1N5711 Schottky diode will have still faster response and a lower voltage drop. And a 1N34A germanium diode will have a lower forward voltage drop (approximately 0.3 V), making the meter even more sensitive. To use the meter for measuring power, be sure the 51 Ω resistor power and diode reverse voltage ratings are adequate for the power level to be used. For a power level of 25 W, the value of V_{dc} will be about 50 V.

RF Current Sniffer

A common reason to go "RF hunting" is to find RF current flowing on feed lines or other conductors. Or perhaps you would like to make some relative measurements of RF current to test antenna or ground radial system performance. Tom Rauch, W8JI, developed the handy RF current meter shown in Figure 3 for this purpose.

[5]

RF current is measured just as described for the toroidal transformer method in Figure 1. To keep leads short, the entire assembly is glued to the back of the 100 µA meter. If a clamp-on transformer is required, try KØLR's design at

lowfer.us/k0lr/currprob/currprob.htm

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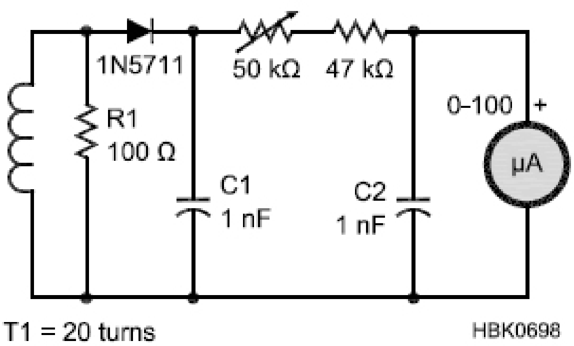

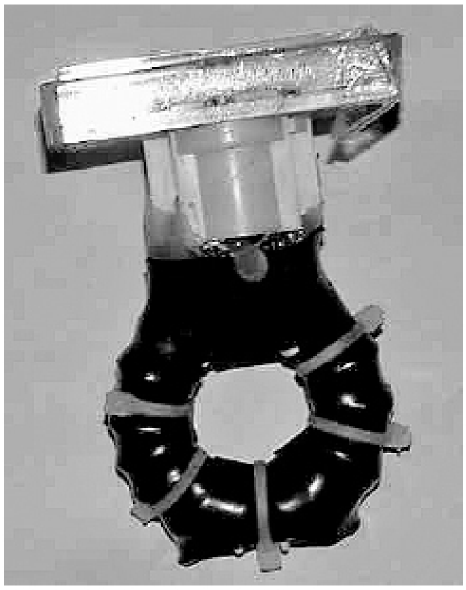

Figure 3 — A toroidal transformer (Mix 2 powdered-iron — T157-2 core) samples RF current, which is then rectified and displayed on a sensitive current meter.

RF Voltmeters

The circuit of Figure 4 was designed by GM4HTU and published in the Summer 2014 issue of the G QRP Club's magazine, *SPRAT*. This interesting circuit uses ½ of an op-amp (U1A) and bridge circuit to convert the input voltage at Pin 3 to a full-wave rectified current through the 100 µA meter. Meter current is equal to the input voltage divided by the value of R1. The meter responds to the average value of the current and can be calibrated for either RMS or peak value. The op-amp should have a gain-bandwidth (GBW) product of several times the maximum frequency signal to be measured. (The NE5532 GBW is 10 MHz.) With these values, the maximum input voltage is about 1 V. The remaining section of the op-amp (U1B) acts as a dc power splitter to provide a virtual ground at its output for the rectifying circuit. The circuit will work fine with a 9 V battery or 12 V power supply.

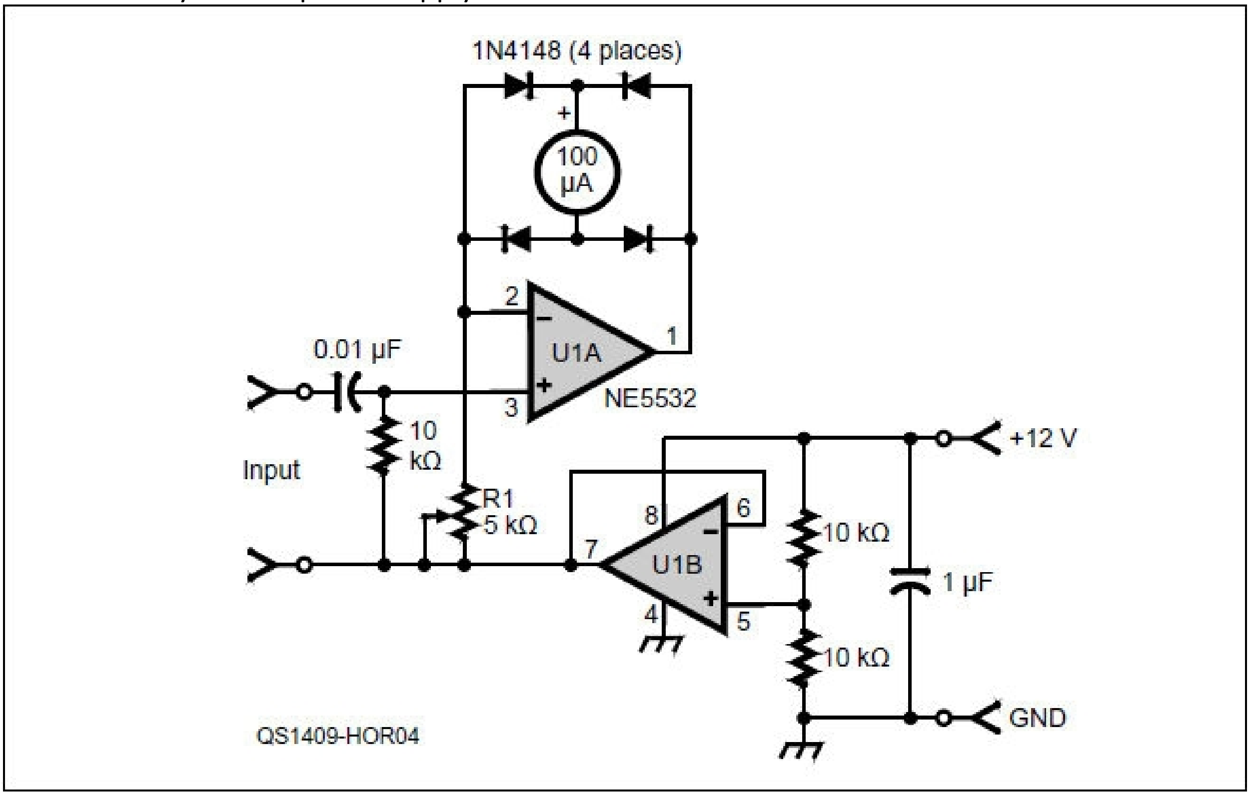

Figure 4 — An RF voltmeter circuit by GM4HTU. U1A functions as a full-wave rectifier that converts the voltage at the input to a full-wave rectified current through the 100 µA meter. R1 is adjusted to calibrate the meter. U1B is a supply-splitter that provides a virtual ground for U1A. The circuit will work at dc supply voltages from 9 to 15 V.

RF Logarithmic Detector

Finally, we often want to measure RF levels in terms of dB. We can use a linear meter and convert the readings to dB mathematically but a circuit that does that conversion for us is much more convenient. Figure 5 shows a circuit that uses the popular AD8307 logarithmic amplifier to provide a linear dc voltage output representing an input voltage of –70 to +10 dBm over a range of 1 to 100 MHz.
[6]

The circuit is described both in a July/August 2014 *QEX* article by Gary Richardson, AA7VM, and in section 7.3 of *Experimental Methods in RF Design*, mentioned earlier.
[7]

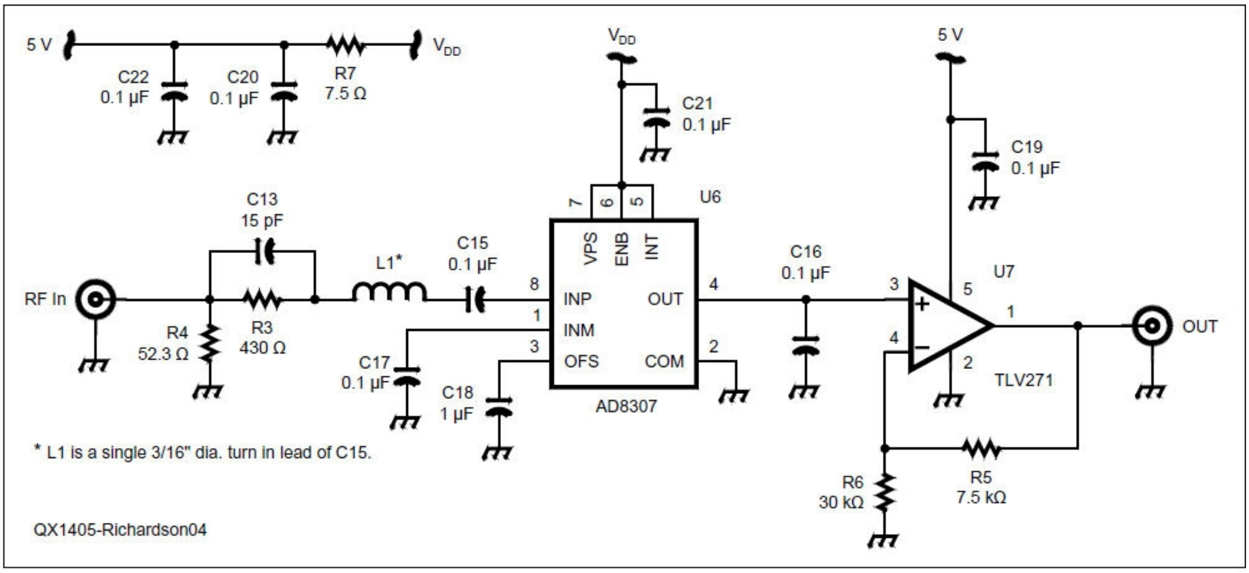

Figure 5 — This wide-range RF detector from AA7VM's July/August *QEX* article outputs a dc voltage based on the input absolute RF power level. The circuit works over a range of 1 to 100 MHz and from –70 to +10 dBm.

Summary

This collection of circuits ranges from simple, passive diode detectors that can be quickly wired together from a junk box and used with an inexpensive DVM, to sophisticated ICs developed for high-performance commercial RF applications. You can build just the one you need or create a whole stable of valuable tools for your RF toolbox!

Notes

1

T. Thompson, WØIVJ, "Technical Correspondence — A High-Power RF Sampler," *QST*, June 2011, p. 52.

2

urgentcomm.com/test-and-measurement/every-toolbox-needs-one-these

3

The Isotee was described in Experiment #103. All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**
.

4

Hayward, Campbell, and Larkin, *Experimental Methods in RF Design,* available from your local ARRL dealer or the ARRL Bookstore at **www.arrl.org/shop** or 888-277-5289.

5

www.w8ji.com/building_a_current_meter.htm

6

www.analog.com/en/rfif-components/detectors/ad8307/products/product.htm

7

G. Richardson, AA7VM, "An RF Filter Evaluation Tool," *QEX*, July/August 2014, pp. 3-6.

Experiment #158 — Test Sets

I have had the experience of mentoring teams of students assigned to work on a design project for an industrial client. We'd discuss the first steps of determining the customer's requirements and writing a specification for what the design was supposed to do.

The next step was enlightening: the development of test methods. "After all," I said, "if you don't agree on how to test the design, how will you know when you're done?"

Hams often just cut-and-try until it's "close enough," and that often works out okay if we're just building one of something. But what happens when that one of something stops working? Or if you need to verify that a piece of equipment is working properly?

Having a test set on hand can be a huge time saver. Test sets are usually an assembly of test equipment or a custom circuit that exercises the important functions of a piece of gear. They are used for troubleshooting, pass/fail tests, and calibration.

The Dummy Load

The most common test set — the dummy load — certainly gets a lot of use. A dummy load is the test set every ham should have. It doesn't have to be "full gallon" capacity, either. Even small dummy loads that dissipate a few watts are useful for various things like testing handhelds, mobile radios, and QRP rigs. Find one that can handle the output of your transmitter or amplifier, and make it easy to use during tuning or testing.

Checking RF Analyzers

Another use for dummy loads is checking your antenna analyzer. As the sophistication of analyzers increases, it becomes easier and easier to misconfigure them. That's why known loads are very handy. They provide a good sanity check on your readings, if nothing else. If the analyzer is broken (or worse, erratic) a dummy load will help you decide what action is needed — none, repair, or replace.

Network and single-port vector analyzers require a set of three known loads to account for the effects of connections to the device or system being tested. The three loads are a short circuit (0 Ω), an open circuit, and the system reference impedance, usually 50 Ω, but sometimes 75 Ω. Each load is applied to the instrument in turn as part of a calibration routine, allowing the analyzer to cancel out the effects of cables, connectors, and construction. This creates what is called a *reference plane*, where the measurement can be made independent of cable lengths and connections.

Making these test loads is pretty straightforward at HF and low VHF frequencies. For the 0 and 50 Ω loads, just solder a jumper or 50 Ω non-inductive resistor across the appropriate connector. As frequency increases, however, parasitic effects of lead length and distributed capacitance start to become significant. If you purchase an analyzer to use above, say, 100 MHz, you should also purchase a set of calibration loads.
[1]

Like the vector analyzer calibration loads, a test set for analyzers and low-power impedance measurement equipment is as simple as a set of known loads. Power resistors are available in TO-220 packages that have low parasitic lead inductance so they can be used into the low VHF range. A typical example is the LTO 100 series of 100 W non-inductive resistors from Vishay (explained in detail at
www.vishay.com/docs/50051/lto100.pdf
).

A wide range of SWR measurements can be checked with a set of resistors having values from 5 Ω (SWR of 10:1) to 50 Ω and then up to 500 Ω. With careful mounting and short leads, the impedances should be consistent up to 2 meters.

It Worked the Last Time I Used It!

One place you really need testing gear is at the hamfest flea market. Many sponsors have taken the welcome step of supplying a test table staffed by some tech-savvy hams. At the table, you'll find a test set of basic instruments so you can check out a radio. This has saved countless dollars from being wasted on inoperative and faulty equipment. Nevertheless, you can't be running over to the test table with every handful of resistors and roll of cable. Having some test equipment to bring along is a great investment. An inexpensive DVM with the capability to test transistors, diodes, and measure capacitance will allow you to sort through components quickly. Bring alligator clip adaptors for the DVM leads as well. A battery-powered grid or gate dip oscillator makes a dandy low-power (if drifty) signal source to see if a receiver is working.

Several basic test sets are also good accessories for the savvy buyer. Start with a low-power dummy load and a set of common coax adaptors, as shown in Figure 1. This will allow you to perform basic functionality tests on QRP and handheld transceivers. You'll be able to use your own VHF/UHF FM handheld to listen for a signal from the radio you're testing, too. If you include an RF peak detector that can be attached to the dummy load with a T connector, you can get an idea of how much power is being generated and whether the output power level is stable.
[2]

Figure 1 — A set of calibrated SMA-series loads in the small box include a short, open, and 50 Ω, along with double-female and double-male adapters. The 5 W 50 Ω dummy load is flat (non-reactive) to well over 1 GHz. A set of coaxial adapters includes cross-family and double-female adaptors for UHF, N, BNC, and SMA connectors. The pair of jumper cables completes the package.

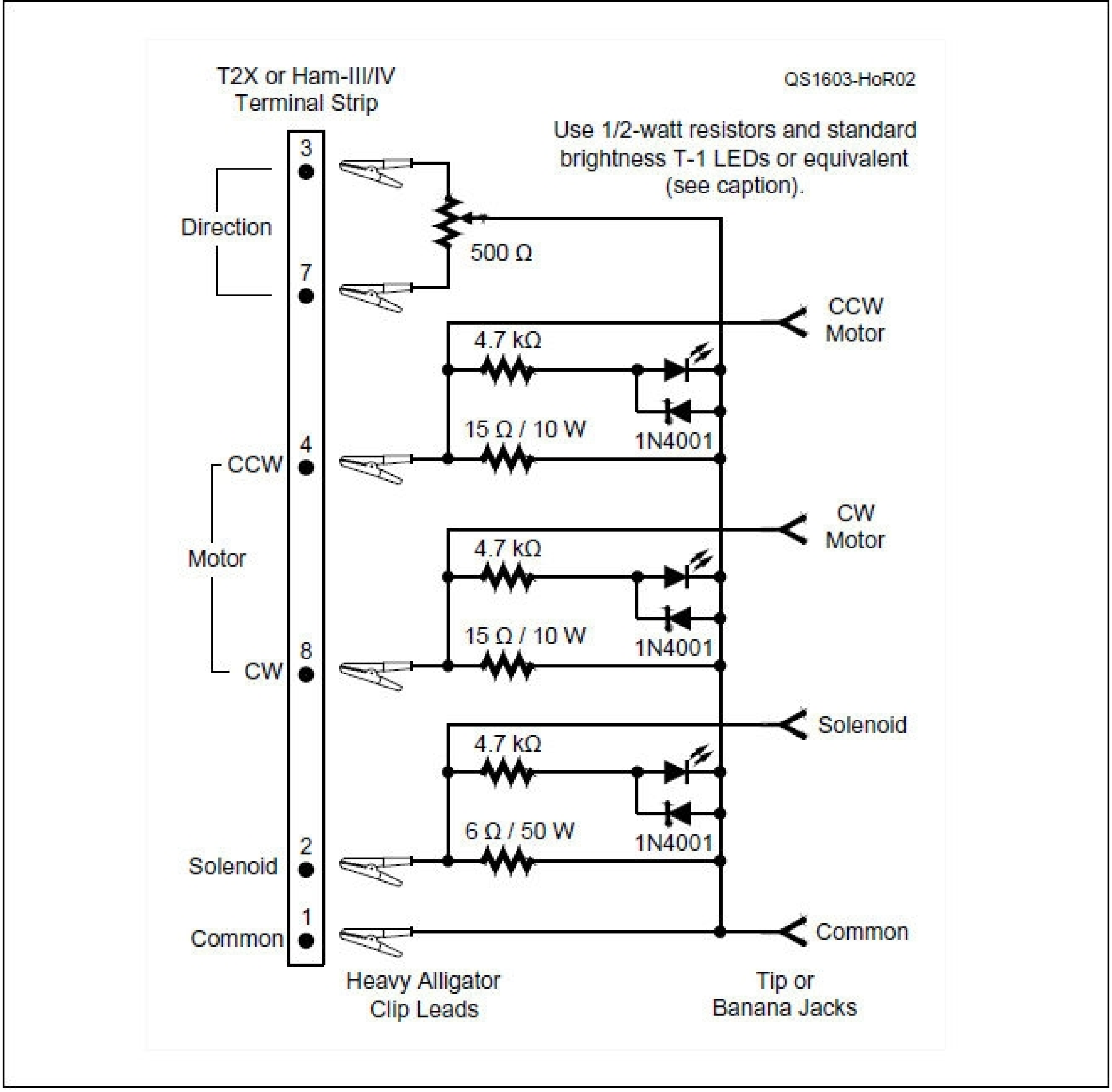

Figure 2 — This test set for T2X and Ham-III/IV rotator control boxes simulates the loads of the rotators. Clip leads are used to connect the test set to the control box. Power resistor ratings were selected to allow testing for several seconds, but not continuously. Select the current-limiting resistor for the LEDs so that a known-good control box results in almost full brightness. A weak or defective control box will be apparent through dim LEDs.

Batteries and power supplies can look fine when measured with the very light load of a voltmeter. You really need to draw some current to be sure the output voltage will hold up when the battery or power supply is loaded. A 10 Ω, 10 W resistor mounted in a metal box will draw almost an amp from a 12 volt battery or power supply, testing the supply's voltage regulator and that the battery (if at least partially charged) can still hold some charge.

A more capable test load is composed of several different resistors mounted on a single piece of scrap aluminum to act as a heatsink. Start with a 1 kΩ, 1⁄2 W resistor, which is a good choice for a light load (resistances don't have to be exact). Then add power resistors of 100 Ω, 10 Ω, and 1 Ω, or similar values. Low resistances can draw large currents, so use banana jacks and #16 wire to handle these loads, and be careful about heating. The goal of this test set is only to provide a short go/no-go test, but hot spots can still develop. Attach a set of tip jacks to plug in the leads for your DVM and you can monitor voltage under load.

Rotator Control Box Test Set

A test set that I've found to be quite useful is the rotator control box test set seen in Fig- ure 2. Having a station with several T2X and Ham-III/IV rotators, there are many opportunities to use it! For testing the rotators, I have a known-good control box — no problem there. Testing the control boxes is another matter, though, so I built the rotator simulator in the figure.

3

The test set schematic is quite simple: the 6 Ω resistor draws 4 to 5 amps typical of an energized brake solenoid. A pair of 15 Ω resistors draw the 2 amps typical of motors. An LED across each load resistor shows voltage through relative brightness. A 500 Ω pot is provided to exercise the metering circuit. A set of heavy-duty alligator clips let me clip to the terminal strip very quickly and tip jacks let me connect a voltmeter. Using this test set, I can quickly tell if a control box is working or not, whether on a hamfest table or at the top of a tower.

Your Test Set

What do you need to test around your station? Is it a special power supply or battery charger? Do you need to calibrate an RF measurement tool? Maybe something to check continuity on that growing box of data cables? Whatever the need, building a dedicated test set will save time and headaches over and over.

Notes

1

The process of calibrating a small vector analyzers is demonstrated in the YouTube video “Vector Network Analyzer’s Calibration Tutorial,” by LA Techniques Ltd, at **https://www.youtube.com/watch?v=lqo3oV2RtlA**.

2

See Experiment #140, “RF Measuring Tools.” All previous “Hands-On Radio” columns are available to ARRL members at **www.arrl.org/hands-on-radio**.

3

Manuals in PDF format are available for download from Hy-Gain at **http://www.hy-gain.com/Categories.php?sub=0&ref=64**. Open the rotator model web page and look for the “[model] Downloads” link.

Experiment #162 — Oscilloscope Triggering and RF

As part of the lab work suggested for last month's column [1], you were asked to take a look at the rising edge of an FET gate drive pulse — a very close look! Being able to observe the subtle changes of a fast-rising pulse's leading edge is not trivial. So this month, we're going to review some of the "triggering" technique required to use an oscilloscope to make these and other high-speed measurements. We'll use triggering to look at an RF waveform in a few different ways.

I assume that you have a scope of your own and can use its basic functions. If you need more background on oscilloscopes, the book *Oscilloscopes for Radio Amateurs*, by Paul Danzer, N1II, is a good introduction to the subject [2] and there are numerous online sources of information, as well. (Note that the labeling on your scope may differ somewhat from the common terms I use in this column.)

RF and Triggering

Triggering on RF signals can be tricky, particularly if they are erratic or intermittent. The result is often a blurry trace on the display. If the RF waveform is synchronized with a cleaner control signal of some sort, it's a lot easier to get a stable display. For example, the CW waveforms in ARRL Product Reviews (see Figure 1) are measured this way — the closure of the key contacts is used to trigger the scope, which then cleanly displays the envelope of the RF output.

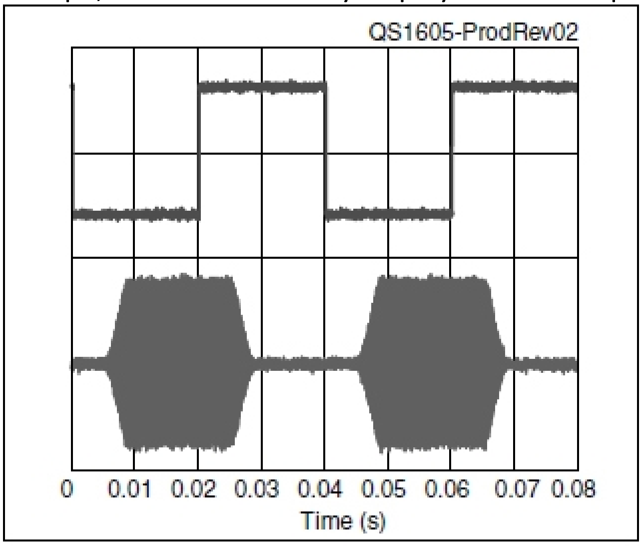

Figure 1 — A CW keying transmitter output waveform (red) synchronized to the key closure (blue). The scope's trigger occurs when the key closure occurs at the left-hand side of the screen.

You can try this yourself with a two-channel scope, a CW keyer, and a paddle. Connect one channel to your keyer output. (If the keyer is connected to the transceiver's KEY input with a phone plug, you may be able to unscrew the plug's shell and attach the scope probe to the exposed terminals of the connector. A splitter adaptor and a spare plug can also provide contacts to the keying input.) Connect the transceiver to a dummy load, turn down the RF power output level, and set the keyer to 30 WPM or so. Choose a quiet frequency on the 30 meter band. Assuming the transmitter is keyed when the KEY input is grounded, set the scope's trigger controls to NORMAL mode with a negative AC slope. Send a string of dits and set the time base to show two or three dits across the screen.

Now pick up some RF with the other probe by wrapping a few turns of hookup wire around the coax feeding the dummy load, as in Figure 2. Connect the probe and its ground clip to the ends of the wire with the scope's vertical sensitivity set to maximum. You will probably have to turn up the RF power level to pick up more than a few mV of RF. Increasing the number of turns of wire also increases the coupled RF into the probe.

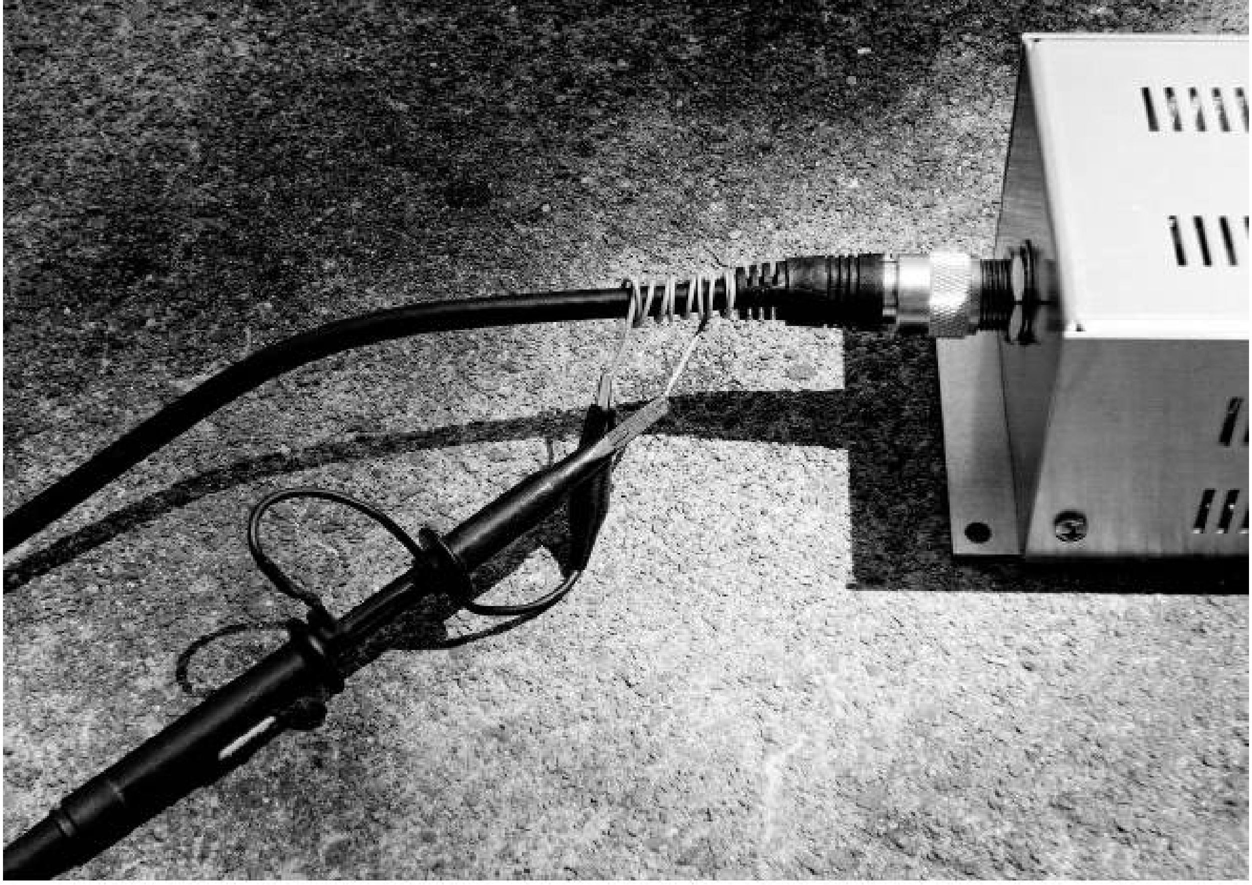

Figure 2 — Enough RF can be coupled out of a coaxial cable via a high-impedance 10x scope probe to display on an oscilloscope. Note that for coaxial cables bundled together, signals can couple between them in the same way!

Alternately, you can use a T connector at the dummy load and devise a way to connect the probe with a short coax pigtail or binding post adapter. Don't connect the scope input directly to the dummy load with coax unless you're sure the power level is low enough not to damage the scope input. It's safer to use a probe.

Key the transmitter again and you should be able to obtain a display very much like the image in Figure 1. (This would be an excellent time to check the transmitter's CW rise and fall time. Values of 4 – 6 msec are brisk enough for good keying, but not so fast that they cause key clicks on adjacent channels.)

Try moving the key closure input to the trigger system's EXT (external) input. The triggering function should work just the same except you'll only see one trace on the display — the RF envelope.

Zooming In

While this provides a good view of your transmitter's keying waveform, it is not possible to see individual RF cycles or the details of how the RF output waveform begins. Switch your trigger source to the channel showing the RF waveform, set the time base to show several RF cycles on the display (100 – 500 ns/div for 30 meter signals), and experiment with the AC slope selection and sensitivity to see if you can stabilize the RF waveform on the display.

It's likely that you'll be able to see some RF, but not with a reliable starting point. The waveform may jump around, or you'll see several waveforms superimposed on each other. This is because the RF waveform is unlikely to start up in exactly the same way each time the transmitter is keyed.

Another option is to generate an "RF Detected" signal of your own. You can make up a simple RF detector circuit, as shown in Figure 3. The output of this circuit follows the envelope of the RF signal. Connect the 40 dB attenuator input directly to the dummy load. You can then trigger the scope from the output of the RF detector either using the AC slope or with the DC level. This will cause the scope to trigger after a few cycles of RF have charged up the detector capacitor. (If the detector will always be used with a 50 Ω dummy load, the left-most 51 Ω resistor can be replaced by the dummy load. Permanently adding this circuit to your dummy load provides a handy peak detector output for other testing purposes.)

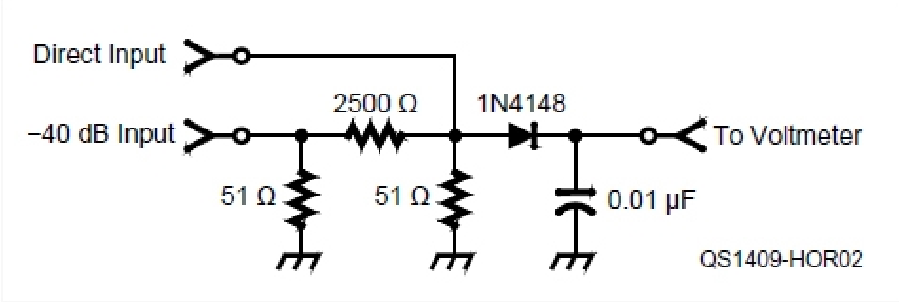

Figure 3 — This peak detector is useful for measuring RF power levels as well as detecting RF waveform envelopes. Be sure the 51 Ω resistors are sufficiently rated for the power level to be used.

Delayed Sweep

If your scope has the feature, this is a great time to experiment with *delayed sweep*, in which the scope waits for a fixed interval after being triggered before beginning the on-screen waveform display. In Figure 1, for example, using delayed sweep would allow you to trigger on the key closure, wait, and then display the RF waveform at a very fast horizontal resolution that could show individual cycles.

With the key closure triggering the scope as before, set the delay to approximately 6 ms before showing the actual RF waveform at 100 – 500 ns/div. The waveform may still be jumpy or blurry, due to the fact that it's unlikely that the transceiver oscillators are synchronized to the keying input.

Single-Sweep Mode

It's a little frustrating not to be able to zero in on those startup RF cycles to see just how your transmitted signal gets going. But there is one more option available on digital scopes those of us who grew up with analog scopes really appreciate — *single-sweep mode*. This option is generally controlled by a knob or switch in the time base area (or menu structure) of your scope controls.

When single-sweep mode is enabled, each triggering event results in just one trace being displayed. You then have to clear or reset the display before another trace will be displayed. The short persistence (how long the trace lasts) of an analog scope makes the one trace very difficult to see at very short sweep times — such as those necessary to view one or two cycles of an RF signal. All sorts of tricks are employed: turning out all the lights in the lab, or putting your head under a hood. (Back in the day, if you were lucky, a storage scope with a very expensive — and easy to damage — display tube was used to hold the image of the waveform.) Today's digital scopes hold the trace onscreen and even allow you to manipulate and measure the waveform on the display.

Go back to triggering the scope directly from the RF signal. Set the scope for single-sweep mode and key your transmitter. It may take a little adjustment, but you'll soon have a nice, bright display of one or two RF cycles just as your transmitter turns on!

Practice and Learn

As long as you are experimenting with your scope, this is a great opportunity to explore some of the triggering system features and really learn how to get the most out of them. Your scope manufacturer may have an online tutorial or manual to follow as well. The scope is your eyes on the RF workbench. The better you know how to use it, the clearer you will see!

Notes

1

All previous "Hands-On Radio" experiments are available to ARRL members at

www.arrl.org/hands-on-radio

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2

www.arrl.org/shop/Oscilloscopes-for-Radio-Amateurs

Experiment #163 — E- and H-Field Probes

At the recent IEEE International Microwave Symposium in San Francisco (**ims2016.org**), dozens of RF professionals checked in at the ARRL display in "University Row." Their interests were far-ranging and included EMC (electromagnetic compatibility), a topic of concern across the industry — to judge from the number of vendors displaying EMI/RFI-related products. One such visitor was Ken Wyatt, WA6TTY. (Ken is co-author of the new ARRL *RFI Pocket Guide*, written with ARRL Lab staffer, Mike Gruber, W1MG.

[1]

)

Ken — an EMC consultant — handed me a complimentary copy of the *Interference Technology 2016 EMC Directory & Design Guide*, which also contained his article, "Assembling a Low-Cost EMI Troubleshooting Kit — Part 1 [Radiated Emissions]."

[2]

Ken's approach typified the mix of commercial and self-built equipment and tools in every ham shack, including some nice-looking and easy-to-make probes for use with a spectrum analyzer, inspiring this month's column.

Probing Questions

Most hams are familiar with the test probes used with oscilloscopes. They present a high impedance (typically about 10 MΩ) to the circuit under test to prevent loading it and changing the voltage. The probes in Ken's article, however, are *E-field probes* and *H-field probes*. Referred to as *near field probes*, they pick up the electric and magnetic field components of electromagnetic (EM) signals, respectively, without directly contacting the circuit at all, and are used with a spectrum analyzer to "sniff" for sources of radiated EMI. A selection of probes made by Ken is shown in Figure 1.

Figure 1 — E-field probes (upper and center right) and H-field probes (center and lower left) are used with spectrum analyzers to detect and compare low-level RF signals. [Ken Wyatt, WA6TTY, photo]

An E-field probe (see the upper right probe in the figure) is the easiest to understand. It consists of a short length of wire at the end of a cable — essentially a short stub antenna. You can make one yourself from coaxial cable by stripping away a bit of the shield at one end — Ken recommends leaving about 1⁄4 inch — and insulating it so it won't come into contact with a live circuit while being used. You can use semi-rigid miniature cable or flexible cable as shown in the figure.

An E-field probe picks up voltage differences between the outside of the probe cable shield and the tip of the probe. Because the probe shield is generally at a relatively consistent voltage, moving the E-field probe around allows you to see changes in the electric field. Without a strict calibration and controlled use, the probe doesn't provide a measure of the absolute field strength, just changes from place to place or before and after a circuit change. The probe's sensitivity also changes with frequency due to the changing electrical dimensions of the probe and of the cable's outer surface. Even with these variations, the E-field probe is a good way to detect and evaluate changes in an electric field, such as around a switching transistor in a switch-mode power supply.

The magnetic or H-field probe (at center and lower left in the figure) is a bit more interesting in both operation and construction. Figure 2 illustrates the probe's basic function, and Figure 3 shows how the probe is constructed. Semi-rigid 50-Ω coax, such as the PTFE-insulated, 0.047-inch dia- meter PE-SR047AL from Pasternack (

www.pasternack.com

) makes a sturdy probe. (Connector-terminated semi-rigid coax is often available as surplus cable assemblies at hamfests and online.)

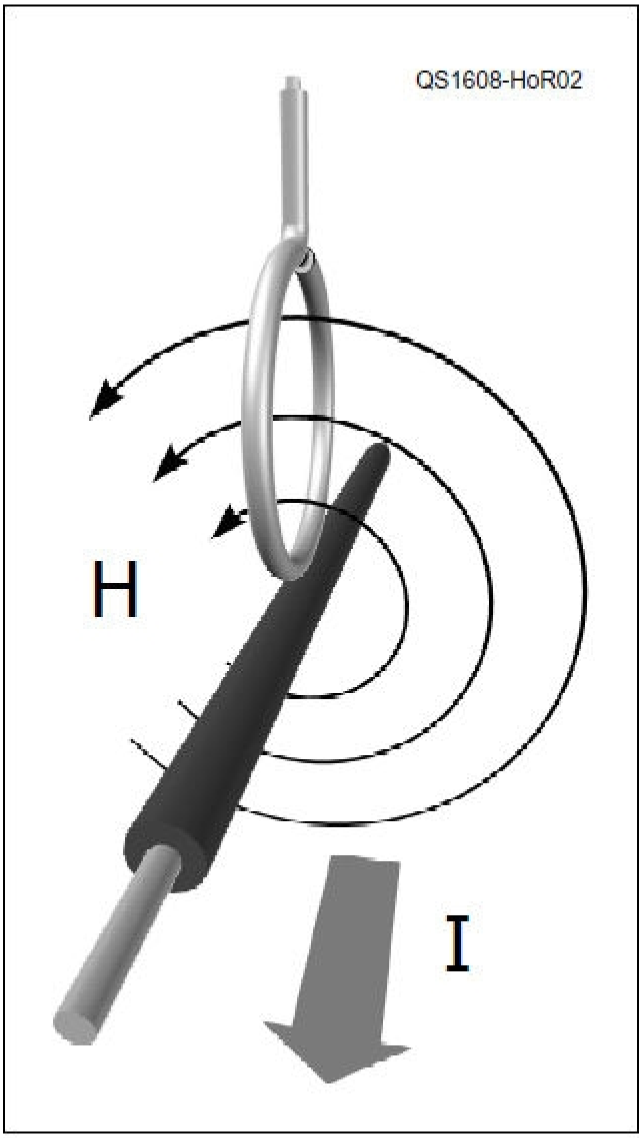

Figure 2 — Because the H field created by current in a wire is circular around the wire, an H-field probe must be positioned parallel to the wire to generate the maximum voltage. When comparing measurements, be sure to position the probe consistently and with the same orientation. [Patrick André, graphic]

The voltage (more correctly, the electromotive force as explained in Experiment #117

3

) generated by the H-field probe is directly proportional to the area enclosed by the loop, the strength of the magnetic field inside the loop, and the rate of change in that field. Because it is current which generates the H field, the generated voltage will be proportional to current, if orientation is consistent from measurement to measurement. Take a careful look at Figure 2 — you'll note that the probe is held with the plane of the loop parallel to the wire or cable to capture the maximum amount of the circular H field and generate the largest voltage. Larger probes generate more voltage, but are less precise about where the current is, and the resonances of shielded versions are at lower frequencies.

The H-field probe can be constructed from unshielded wire, as well, soldering the free end of the wire to the analyzer cable's shield at a connector or some other convenient arrangement. This allows E fields to couple to the H-field probe through capacitance between the probe wire and the circuit generating the E field. This is why the shielded configuration in Figure 3 is usually preferred. The only issue created by the presence of the jacket shield is that a resonance is created by the shorted stub of transmission line, usually in the mid-UHF range depending on loop diameter and cable type. (The alert reader will note that the H-field probe looks like and works the same way as the popular magnetic loop receiving antennas!)

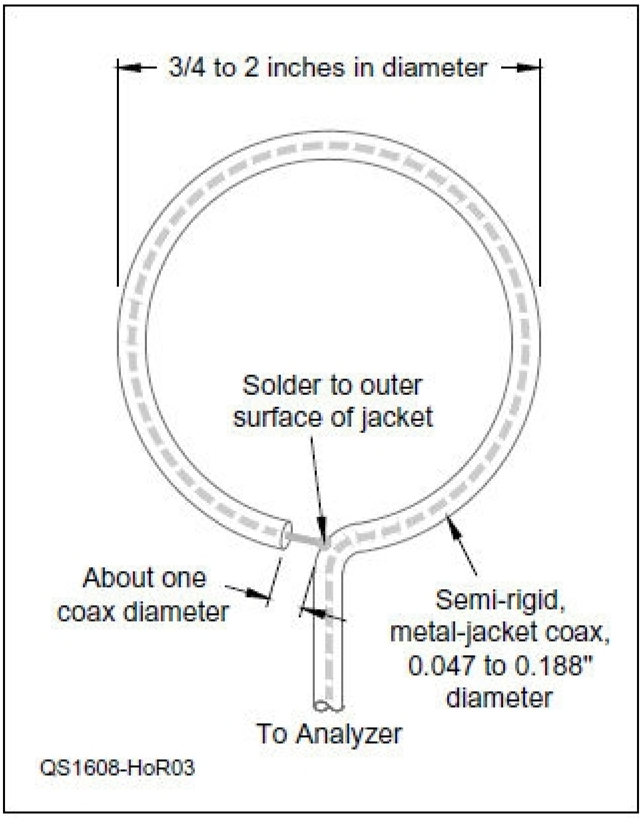

Figure 3 — An H-field probe can be made from semi-rigid coax by removing the jacket and dielectric for a length of approximately one cable diameter. The exposed center conductor is then soldered to the outside surface of the jacket.

The Host Analyzer

What would a ham use the E- and H-field probes to accomplish and what sort of instrumentation is required to make the measurements? Let's start with the second question — do you need a spectrum analyzer? That would help, but could be a bit of a budget problem! There are several options.

If you have a digital scope, you might want to check the manual to see if there is an "FFT," or Fast Fourier Transform, option installed. This function converts a sampled *time domain* waveform into a spectrum analyzer-style *frequency domain* display of amplitude versus frequency. (You can read about FFTs in your scope manual, online, or in *The ARRL Handbook's* chapter on DSP.)

New instruments such as the Rigol DSA815 or Siglent SSA3000X are in the $1200 – 1500 range and the USB-based Triarchy Technologies TSA6G1, which uses a host PC for display and control functions, costs a bit over $600. An inexpensive wide-band SDR such as the AirSpy SDR ($200 from www.airspy.com) with *Spectrum Spy* software can also act as a spectrum analyzer.

With analyzers no longer the expensive, fragile instruments they once were, used analyzers are available for a few hundred dollars, and your club may have a member with an analyzer you can borrow.

Finally, if you are dealing with single signals, such as an interfering carrier, you can use your receiver as a "tunable analyzer." Because you are basically looking only for changes in signal level, the receiver's S meter can provide the amplitude value and the tuning dial gives you frequency.

Using the Probes

Here are two common ham radio uses for E- and H-field probes in ham radio. The first is looking for RF "hot spots" using the E-field probe. A high E-field at the end of an unconnected cable or on an equipment enclosure can couple to adjacent equipment and cables through stray capacitance. Inside a piece of equipment, E-field hot spots can couple signals between parts of a circuit you'd rather stay isolated. It's also common to use the E-field probe simply to "sniff" for RF around a piece of equipment suspected of causing or receiving RFI. Once you have the hot spot identified, you can try different solutions to the problem.

Going to Trade Shows

My encounter with Ken illustrates the value of going to a trade show or exhibition for your profession or a field of interest. You will always learn something new, make or renew a connection, or just see the world from different perspectives, regardless of "which side of the table" you are on. Seminars and tutorials are often available, as well. If admission is charged, look for a "free day," or go late in the show to see if you can get in at less than the full price. Take a few QSL or business cards, and get ready to learn!

The H-field probe is good at finding RF current to diagnose RFI being radiated by equipment and RFI being received by equipment. Using the H-field probe to make A/B or "better or worse" measurements helps you tell whether your fixes are helping, hurting, or missing the mark completely!

For example, you might have a cable picking up your transmitted signal and causing RF feedback in an audio circuit or getting into your living room entertainment center. Armed with your ferrite EMI suppressor cores, you can first locate the problem cable, measure relative current levels with the H-field probe, put on a ferrite core, and re-measure (with the probe oriented just the same as for the first measurement) to see if the current has been affected. This greatly reduces the "Well, let's try another core!" guessing game, saving time and money.

Notes

1

ARRL Item no. 5008, available at
www.arrl.org/shop
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2

Interference Technology, *2016 EMC Directory & Design Guide*, page 70 – 78. Copies are available at
www.interferencetechnology.com
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3

All previous Hands-On Radio experiments are available to ARRL members at
www.arrl.org/hands-on-radio
.

Experiment #168 — Evaluating Filters

A person just learning about radio could be forgiven for thinking radio is mostly just about filters, with the occasional oscillator or modulator tossed in for good measure. Because the filtering function is everywhere, whether analog or digital, we need to know how to describe filters. There are three important parameters that we'll cov

er this month — amplitude response, ultimate rejection, and return loss.

Filter Fundamentals

The basics of filters were covered by two earlier Hands-On Radio experiments: #50 and #51, "Filter Design #1 and #2." [1]

More information about filters can be found in experiments #87 and #88, which are about using *Elsie* (filter design software from Tonne Software,

www.tonnesoftware.com

), and #156, which uses *Elsie* to design a broadcast interference (BCI) reject filter.

Testing a Real Filter

I got an e-mail from Scott Roleson, KC7CJ, titled "Exp #156 BCI Filter — It works!" Scott built a nice version of the filter (see Figure 1) in a die-cast aluminum box, and made the PCB from scratch. The capacitors are silvered-mica with a 5% tolerance. For the inductors, he used miniature encapsulated components.

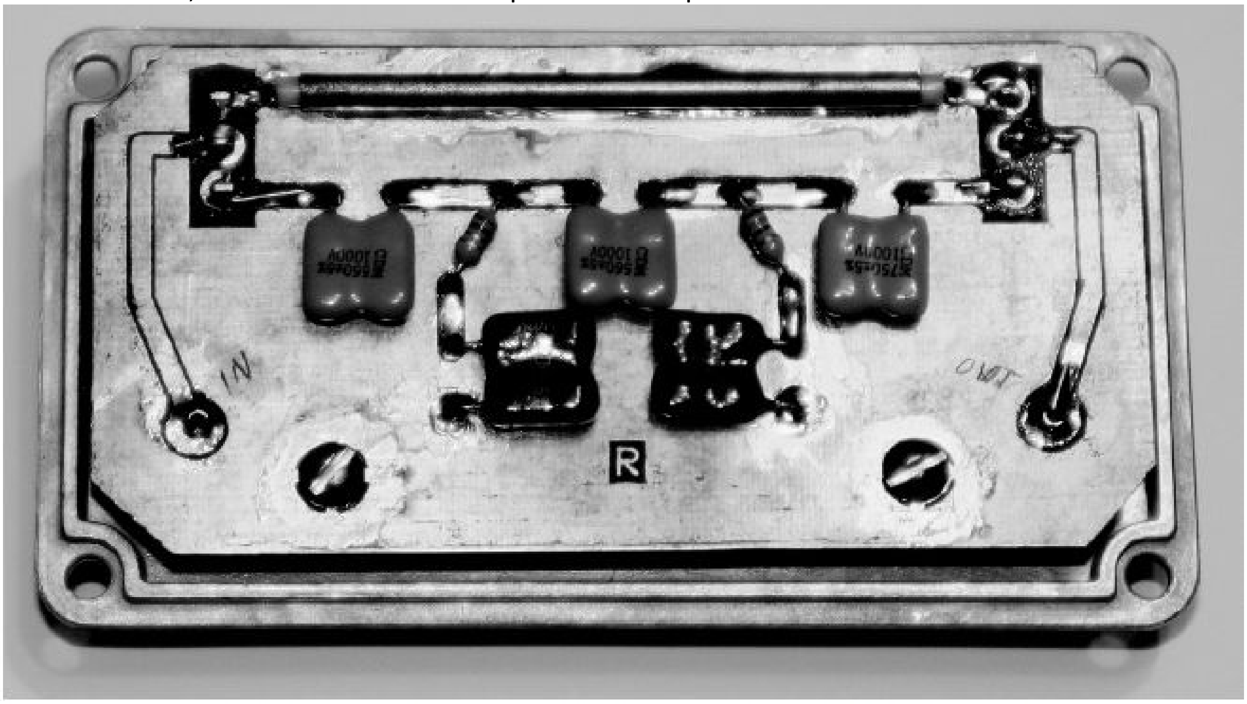

Figure 1 — KC7CJ's homemade version of the BCI filter described in Hands-On experiment #156. The small green components are the encapsulated, subminature inductors. At the top of the board, a short piece of semi-rigid coax forms the filter bypass path.

I set up my vector network analyzer (VNA) to measure his filter's input-to-output attenuation from 1 kHz (referred to as "zero" frequency) through 10 MHz. Along with Scott's version of the Hands-On BCI filter, I also tested a commercial BCI filter from ICE Communications. Both are intended to be used at and above 1.8 MHz, the 160-meter band, where strong local AM BC stations can cause severe receiver overload. Scott's version is receive-only and uses components rated for low power. The ICE filter is rated for 300 W and can be installed in the output of a regular transceiver.

Ultimate Rejection

Ultimate rejection is the attenuation the filter applies to signals far from the cutoff or rolloff frequency. In Figure 2, you can see there are deep notches in the filter's stop band. (Compare the as-built filter to the predicted performance in Figure 3 of experiment #156 — an important verification step.) That's okay — we selected a filter family (Cauer) that obtains a steep rolloff by placing notches at strategic frequencies. But it's not realistic to give the attenuation of those notches (51, 63, and 58 dB right-to-left) as the filter's ability to reject AM BC signals. The ultimate rejection of this filter is measured at the maximum of the two peaks in the stop band, 39 dB of attenuation at about 750 kHz. Did this amount of attenuation satisfy the original design specification for 40 dB at 1.6 MHz? There is plenty of attenuation at 1.6 MHz due to the notch placed there, but across the BC band, we just barely missed by about 1 dB. Pretty good, nevertheless.

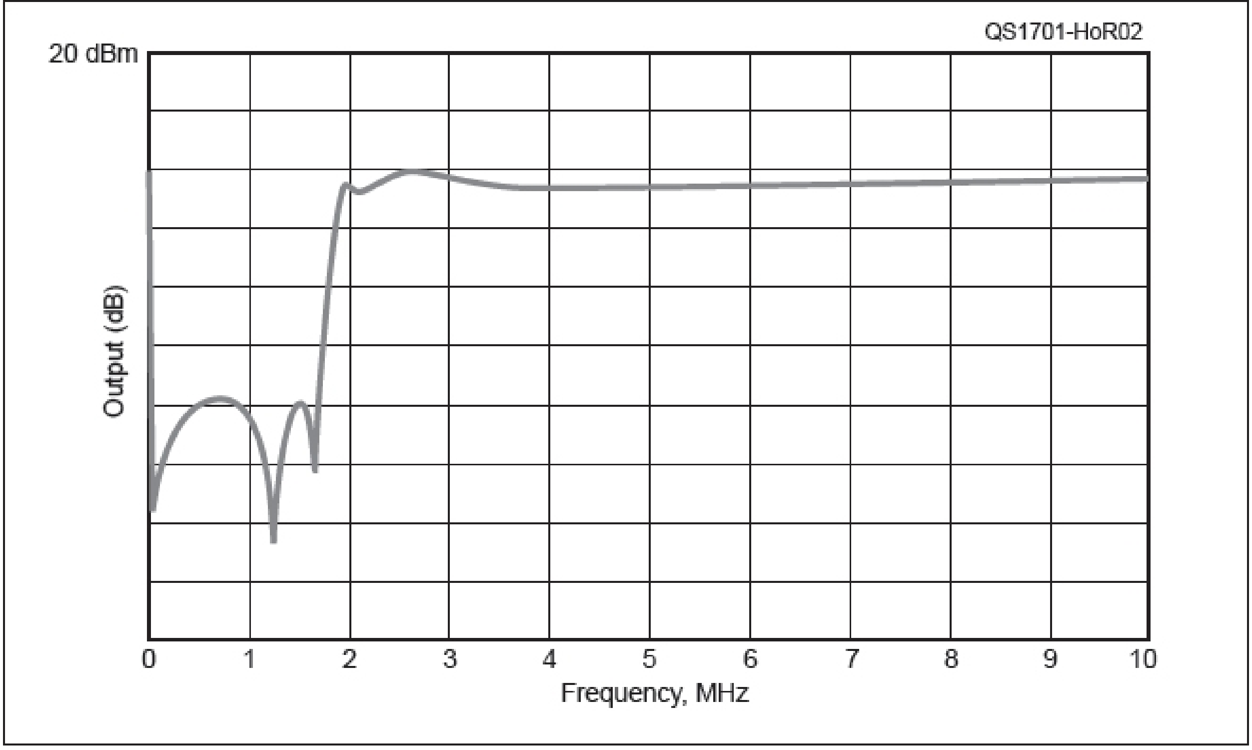

Figure 2 — The input-to-output amplitude response or S21 of KC7CJ's filter. The vertical scale (shown in blue at the upper left) is 10 dB/division. The reference level at the top of the graph is 20 dBm. The frequency range is from 1 kHz (given as "0" where a low-frequency artifact is visible) to 10 MHz at 1 MHz/division.

Insertion Loss

A close look at the filter's response shows that the shape is very close to what *Elsie* predicted, right down to the passband ripple between 1.8 and 4 MHz above the filter's cutoff. Scott's filter rolls off a little higher than expected, hitting 10 dB of attenuation at 1.8 MHz. Because this is a receive-only filter, that extra 7 dB of attenuation at 1.8 MHz (we only wanted 3 dB) isn't a serious deficiency, and the filter will work fine. As the response flattens out, we can see there is about 3 dB of attenuation between 4 and 5 MHz. Similarly to how we measure ultimate rejection, this "worst-case" attenuation in the filter's passband is the filter's *insertion loss* value.

In Figure 3, you see a very different filter response (the blue trace). Because this filter is expected to be used at 100 W power levels or higher, insertion loss must be minimized. From the inset photo of the filter components, you can see that full-size toroidal inductors are used. These have far lower resistance, causing less loss than the subminiature inductors used in the receive-only filter that are wound with very fine wire.

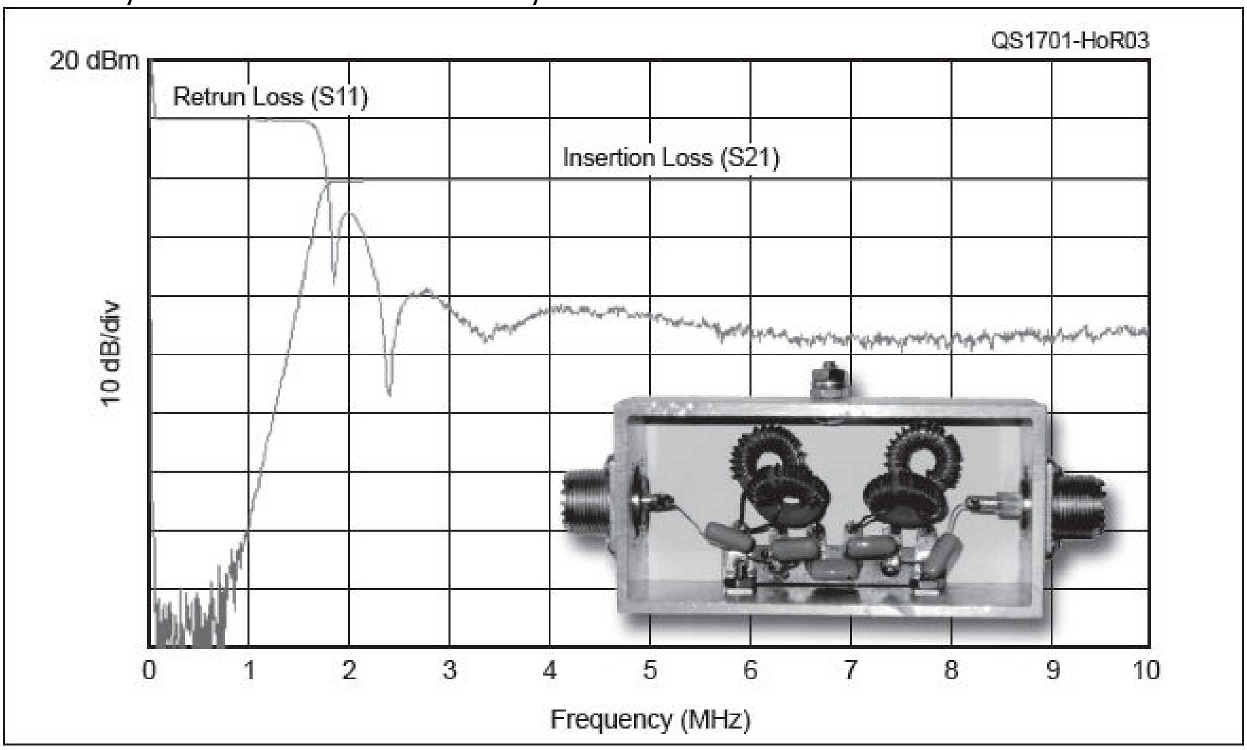

Figure 3 — The frequency response of an ICE 402X BCI transmitting filter. Graph scales are the same as for Figure 2. Both S21 (blue) and S11(red) (Return Loss or RL) are shown. The filter's internal construction is shown in the inset.

The ICE filter's insertion loss is less than 0.5 dB from about 2.2 through 10 MHz (0.5 dB of loss is the same as 10.9%). The tradeoff, as we discussed in experiment #156, is the steepness at which the response rolls off. While the ICE filter has very good ultimate rejection (70 dB), it doesn't reach our attenuation spec of 40 dB until 1.25 MHz, which is well inside the BC band. This is a consequence of selecting the different filter family, requiring extra components to construct.

SWR and Return Loss

VNAs and single-port analyzers like the Array Solutions AIM 4170 (**array solutions.com**), SteppIR SARK-110 (**stepper.com**), and others measure the S-parameter S11, also called *return loss*, or RL. [2]

This parameter is the ratio in dB of how much power is reflected back to the source by the load. (The value of return loss is positive, just attenuation is specified in positive values of dB.)

RL can easily be converted to SWR and vice versa. Higher values of return loss mean less power is reflected and so indicate a lower value of SWR. For example, RL = 6 dB is an SWR of 3:1, 9.5 dB is an SWR of 2:1, 14 dB is an SWR of 1.5:1, and so forth. [3]

Figure 3 also shows the filter's RL (S11) as the red trace. Below 1.8 MHz, RL is very low and SWR into the filter is quite high. But just below 1.8 MHz, RL is 10 dB (SWR approx. 2:1) and is greater than 30 dB (SWR = 1.07:1) for 80 meters and higher frequency bands. While RL is not so important for receiving filters, it is obviously quite important for a transmitting filter.

Making Filter Measurements

It would be nice if we all had VNAs, but even though they are more affordable than before, most amateurs use impedance analyzers and antenna analyzers for measuring return loss or SWR. An oscilloscope and signal generator can be used to make measurements of filter response. If you don't have a signal generator, use a manually-tuned SWR analyzer such as an MFJ-259/269-series unit. Be sure to measure both input and output voltages at each point. Be aware that most filters are designed to be terminated in a specific impedance, such as 50 Ω. Connect the filter to a dummy load when making your measurements, to avoid errors.

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**.

2

S-parameters are discussed in Hands-On experiment #72.

3

A Return Loss — SWR converter is available online at **www.microwaves101.com/calculators/872-vswr-calculator**.

RF Techniques

Experiment #135 — Crimp Connectors

If there is one experience shared by nearly every radio amateur since World War II, it is the chore known as "the putting on of the connectors." In particular, it is the coaxial connector of which I speak, usually a PL-259 and often with a UG-175 or UG-176 adapter. In recent years, the connector installation game has been changed dramatically by reliable, easily installed connectors that attach to the cable by crimping. While not suitable for every application, the crimp connector satisfies the needs of most hams pretty effectively. This month's column discusses PL-259 and Type F crimp connectors, the types most commonly encountered.

Why Not Solder?

There are some very good reasons for using a soldered connector. In a complex station with many strong signals present (such as at a big multiop contest operation) the slight improvement in shield connection resistance can make a big difference in keeping harmonics and intermodulation products off the air and out of receivers on other bands. Another example is for connectors exposed to the weather. A solidly soldered connector that is exposed to the weather may resist water incursion or corrosion a little better than a crimped connector, although both must be thoroughly waterproofed.

Nevertheless, I cannot count the number times I have unscrewed the backshell of a PL-259 (someone else's, of course!) and found the shield loose or hanging by a few wires, balls of solder that make little or no contact with the body of the connector, or even no solder whatsoever! While it would be great if all hams could turn out smooth, filleted, shiny soldered connectors, it would be much more reasonable to provide a simpler method that everyone can perform properly. Enter the crimp connector.

Crimp Connector Basics

The military, telecom, and aerospace firms discovered decades ago that for connections experiencing vibration or those exposed to the air's moisture and contaminants, a "gas-tight" metal-to-metal crimp was actually more reliable than the best soldering job. We can gain the benefit of their hard-won experience.

This column is not a detailed, step-by-step "how-to" for installing crimp connectors. Each manufacturer provides detailed instructions and all the necessary tooling to perform the job correctly. If you do not follow those instructions, you will not reap the benefits of crimping and your connections will be less reliable than soldering. Resist the urge to "cheap out." Learn the right techniques, use the right tools, and you will be amazed at how quickly you can install connectors reliably and inexpensively.

The primary goal of crimp connectors for coaxial cable is to create a reliable connection to the braided shield while maintaining the wire-to-wire connections between the wires making up the braid. Once the shield braid is no longer held in place by the jacket, the wires begin to loosen and oxidize, so the braid begins acting less and less like a solid tube of metal. The crimp connection re-captures the braid wires and holds them firmly against each other between two tubular sleeves (the PL-259) or between the body of the connector and a shoulder gasket that slides into place (the Type F). For the crimped connection to last and to resist vibration and stress, a certain minimum force must be applied and the crimping surfaces have to be properly aligned. That's why it is important to trim the cable correctly and to use the racheting tool to apply the crimping pressure.

The first step toward a successful crimp is proper preparation of the cable. I highly recommend that you use the "automatic" stripping tools — such as the one shown in Figure 1. Each type of crimp connector requires the removal of a specific length of jacket and braid, so use the right stripping tool or cutting die for your cable and connector.

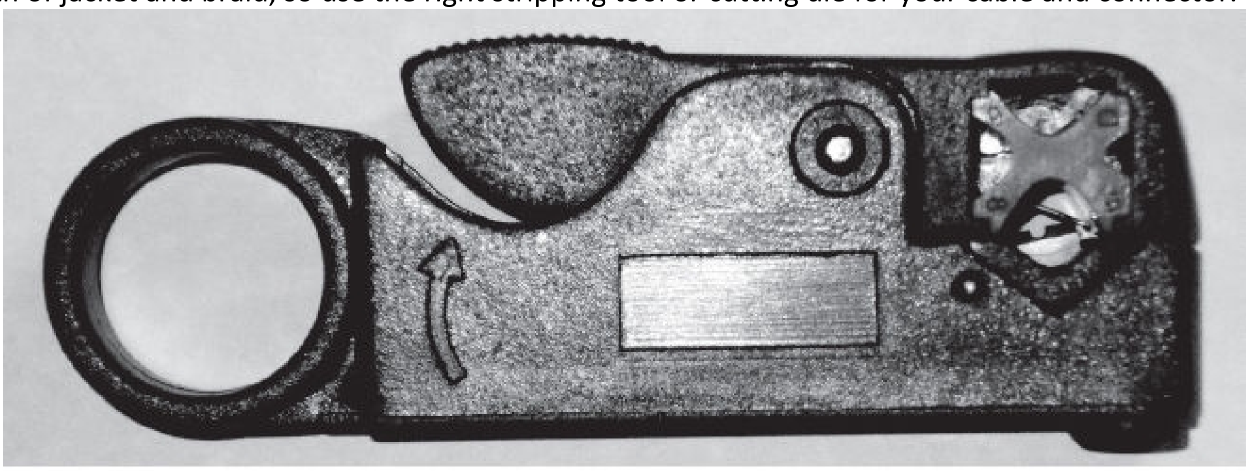

Figure 1 — This inexpensive coax stripping tool removes just the right amount of cable and braid, along with a section of the cable's jacket, without disturbing the braid wires. The orange jig can be rotated to hold different diameter cables.

Start by cutting the end of the cable cleanly and at right angles to its length. If the cable is flattened, use pliers to gently squeeze it back to approximate roundness. Open the stripping tool and insert the cable until it is against the stop or at the specified position.

Close the tool and spin it around the cable. You do not have to squeeze the tool's cutting blades against the cable — the tool will apply the needed pressure. You will first hear the blades crunching through the braid then the tool will begin to turn smoothly and silently. After a couple of smooth turns, stop turning and remove the cable from the tool without opening the tool. Gently pull off the sections of jacket, braid, and dielectric. You can do this with a knife but it takes a lot longer, isn't as precise, and usually messes up the braid.

Crimping PL-259 Connectors

Figure 2 shows the basic steps for installing a crimp PL-259 connector. I have found that on the larger cables, after the initial cuts, you may need to rock the tool back and forth a little as you pull the tool and cut sections off the cable. This breaks any small threads of dielectric surrounding the center conductor that weren't cut. For stranded center conductors, a gentle twist of the tool in the direction in which the strands are wound often helps the cut sections come off more easily. Remember, try to disturb the braid and center conductor as little as possible!

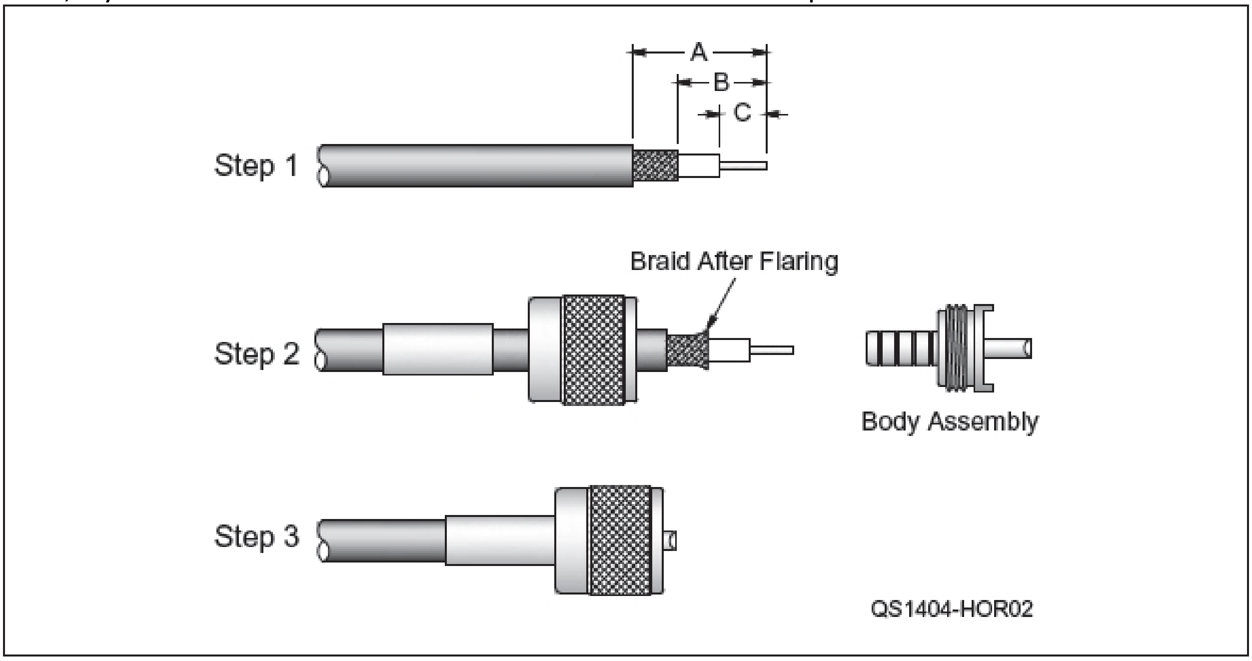

Figure 2 — The basic steps to cutting and installing a crimp PL-259 connector. Each manufacturer will specify the right dimensions for A, B, and C. All connectors are not the same!

Before inserting the cable into the body of the connector, be sure that a stranded center conductor does not have loose strands that might not go into the hollow center pin. Those strands will contact the connector body, creating a short circuit. Twist any loose strands back together with the main conductor. For large-diameter cables with stranded conductors, count the strands when they emerge from the center pin. Now check to be sure you put the outer crimp ferrule and outer shell on the cable!

As you insert the cable into the connector, slide the crimping tube under the braid very gently and straight along the cable. You may have to flare the braid a little bit to get the crimping tube under the braid. Remember, minimum braid disturbance! With the braid all the up to the body of the connector, the center conductor should emerge from or be flush with the end of the connector's center pin. Verify that all of the strands of center conductor got into the center pin, with an ohmmeter check between the shell and center pin for short circuits.

Slide the ferrule up and over the exposed braid. The jacket of the cable should extend into the end of the ferrule, ie — no braid should be exposed. Open the crimping tool and lay the crimping ferrule into the crimping die with the tool against the back of the connector body as in Figure 3. Squeeze until you hit the last click and the tool releases. Inspect the ferrule for equal indentations on all sides and that the cable is not loose in any way. Now crimp or solder the center conductor in the center pin. I like soldering the center pin to seal it against water. You're done!

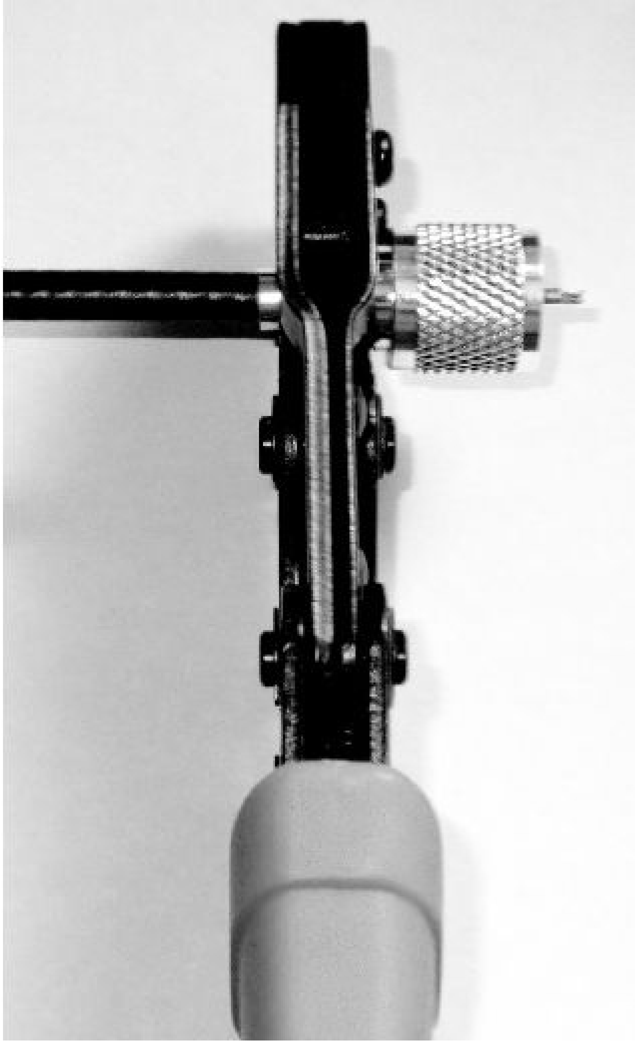

Figure 3 — Proper placement of the ferrule on the cable and in the crimping tool before beginning the crimp. The crimping tool is against the back of the connector body. You can see the center conductor extending from the center pin.

Crimping Type F Connectors

The basic process and cautionary steps are all similar to PL-259 connectors — using the right stripping tool, and crimping with the right tool and die sizes. Designed for use with RG-6 TV cable, there are two types of Type F crimp connectors. Most sleeve-type connectors are one-piece designs. You crimp the connector's outer sleeve onto the inner sleeve with the braid captured between them. There is no separate ferrule to slide over the cable jacket and the center conductor of RG-6 cable is copper-plated steel that forms the connector's center pin.

The compression-style Type F connectors are sturdier and more robust. They capture the braid between the inner body of the connector and a sliding shoulder gasket that is compressed into the body of the connector with a special tool. The braid is flared back over the jacket before the cable is inserted into the connector. The connector and cable are placed into the crimping tool as shown in Figure 4 and the gasket forced into the connector to capture the braid and cable. The plastic gasket also squeezes the cable jacket to keep water out. More detailed drawings for installing Type F and several other consumer-type connectors are available online, such as the Ideal Multi-Media Installation Guide.

[1]

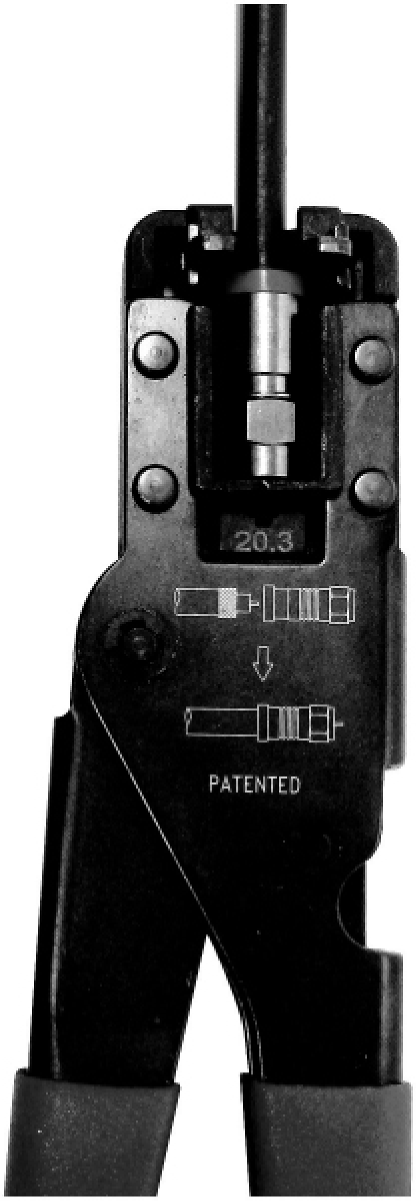

Figure 4 — The special compression tool is used to force the plastic gasket into the body of the connector, capturing the braid and securing the cable jacket. [Photo courtesy of DX Engineering.]

Crimp — Your Style?

The ease of installation is not a panacea. Crimp connectors may not be suitable in high-stress or high-EMI environments. Waterproofing is even more important for crimp connectors and many top station owners use only soldered connectors outside. Nevertheless, the expense of special crimping tools and coax strippers is worthwhile and numerous *QST* advertisers offer kits that work with a variety of cable sizes. In the end you'll be glad you learned how to use crimp connectors!

Notes

1

www.idealindustries.com/media/pdfs/products/guides/p-2804_multi-media_installation_guide.pdf

Experiment #144 — The Myth of RF Ground

Sooner or later, just about every ham who builds an HF station — whether it's at home, in the car, at Field Day, or for portable operation — experiences the rite of passage called an *RF burn*. Although painful, it rarely creates a physical mark, just a certain wariness on the part of the burn-ee. What's happening here?

You just had an exciting encounter with a high-impedance point on your antenna system. Impedance being the ratio of voltage to current, when power is applied, a high RF voltage will be present at these points. Who says you can't feel RF? "But wait," you exclaim, "the antenna is up in the air and connected to the antenna tuner! I'm not touching my antenna system!"

Oh yes, you are! Unless your station is built inside an RF-tight metal enclosure or is otherwise isolated from the antenna and feed line, every coax shield, every enclosure, every unshielded wire…anything connected to the transmitter directly or indirectly should be treated as part of the antenna system. That includes *you* when touching any of those conductors! Take a look at Figure 1, which shows a typical home station. Everything in that figure is part of the antenna system of that station.

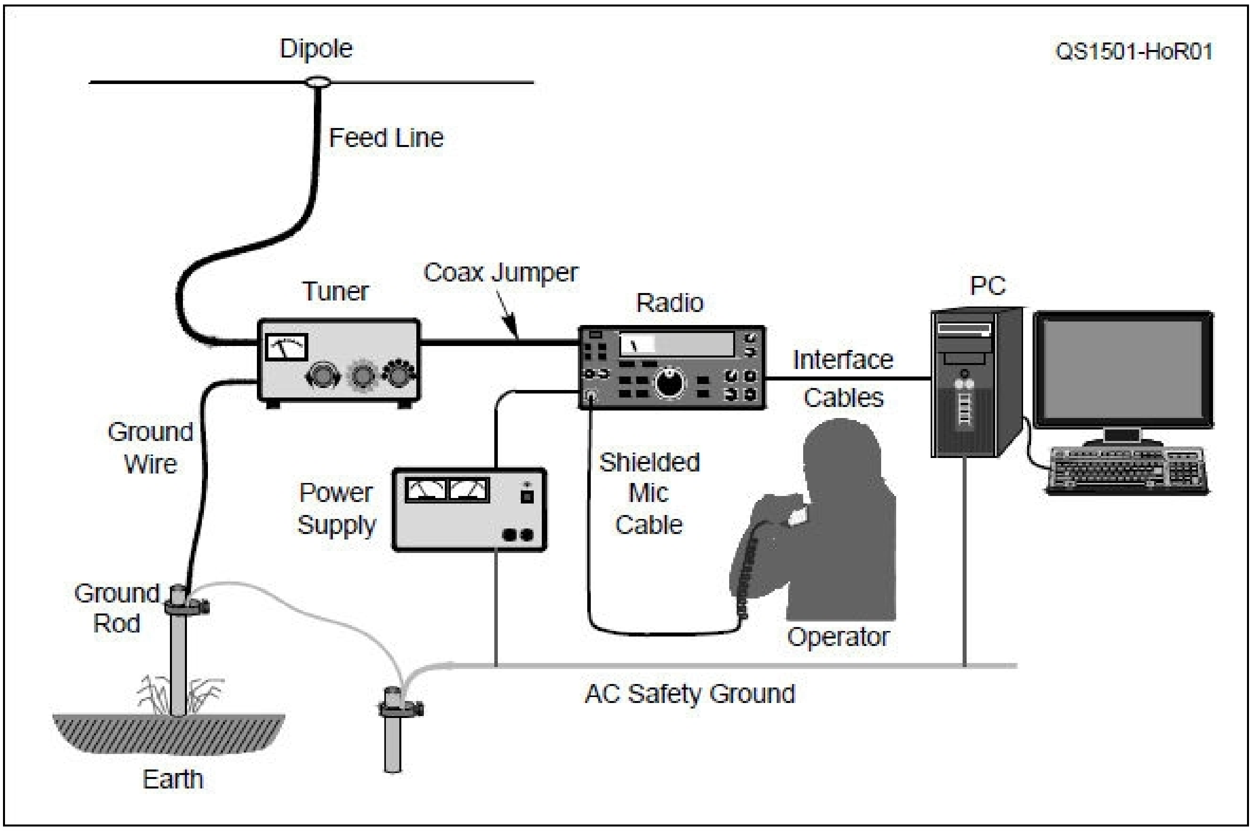

Figure 1 — The complete antenna system for a typical shack. Everything conductive, including the operator holding a microphone, is part of the antenna system.

If the station is operating on 10 meters, at least one potential hot spot is never more than about 8 feet away. Why? Consider the wavelength at 28 MHz. It's 33.4 feet and ¼ wavelength is approximately 8.3 feet. When a conductor is excited by RF, either directly from a signal source or by picking up radiated energy, a pattern of peaks and nulls for both voltage and current is created. Peaks are ½-wavelength apart and so are the nulls, with peaks and nulls offset ¼-wavelength apart. Whether the conductor happens to be a wire, the outer surface of a coax shield, an equipment enclosure, or a "ground" wire makes no difference. It's all a conducting surface as far as the RF is concerned, regardless of what we call it.

While less dramatic than "getting bit," one has to watch out for RF currents, too. Any time the voltage "over here" is different than the voltage "over there," current will flow. When the current flows on the outside of an enclosure or coax shield, it's generally not a big problem. The fun begins when it finds a way into the electronics via an unshielded connection (like a power cord) or an improperly connected shield that conducts the current inside an enclosure instead of keeping it outside. RF that's where it shouldn't be can wreak havoc with a circuit's operation: audio gets garbled, keyboards stop working, control interfaces stop controlling.
[1]

The situation gets particularly interesting when instead of a coax-fed dipole, the antenna wire itself is connected directly to the antenna tuner. This type of connection is often used for portable operating as an "end-fed" or "random wire" antenna with a "counterpoise" (a piece of wire laid on the ground or floor) replacing the ground rod. In this case, the antenna itself consists of everything from the end of the counterpoise to the end of the wire in the air. The equipment and operator are thus all connected to the feed point of the antenna. Imagine the feed point of the dipole in Figure 1 being connected right at the output of the antenna tuner and you get the idea. This also explains why the results of using these directly fed antennas can be inconsistent, because there is so much variation in what the antenna system actually consists of.

Obviously, we would like to control the RF voltages and currents so they don't cause our equipment to malfunction or burn our fingers. The natural tendency is to think, "I'll just ground everything and it will be at zero volts — problem solved!" Not so fast! You're partly right, but we have a failure to communicate, as they say.

Grounding and Bonding

There are grounds and then there are grounds. Consider the actual ground, the soil itself. The Earth acts as a "zero voltage reference" for ac power and low-frequency systems. The *ac safety ground* in your home consists of the power wiring's ground wire (bare or with green insulation), which is connected to the Earth through a ground rod that is connected to the main circuit breaker panel. The ac neutral of a typical two-phase home is also connected to this same ground. (See the *National Electrical Code* and your local building codes for a complete description of what is required in your particular circumstances.)

Any exposed conductive enclosure of an appliance or machine — including your radio equipment — should be connected to the ac safety ground to conduct *fault* or *leakage current* away from you and back to the Earth. It is this current flow that trips Ground Fault Circuit Interrupter (GFCI) circuit breakers.

The purpose of the ac safety ground has nothing to do with RF and it should never be expected to act as any kind of voltage reference above a few hundred kilohertz. That means even if all of your equipment is properly grounded for ac safety, you still have no control over RF voltages and currents. In fact, as Figure 1 illustrates, the ac safety wiring is a part of your antenna system, too.

What if you install a ground rod outside the station and run wide copper strap to it like all the literature tells you? There is another word for ground connections and that is "antenna!" Any ground connection longer than about 1⁄10 of a wavelength begins to act like an antenna, including transmission line-like effects. If the electrical length is close to ¼ wavelength (or any odd number of ¼ wavelengths) the impedance of the wire becomes very high, effectively becoming a resonant open circuit.

Back in the days when most amateurs operated below 15 MHz, a few feet of wire was electrically short enough to serve as a common connection. As operation at shorter wavelengths became more common, the connection to a ground rod got electrically longer and less effective. Hams with shacks on an upper floor had (and have) the same problem at any frequency.

The solution is to stop looking for the elusive "zero voltage connection" at RF. The Earth is not a magic drain into which all of our unwanted RF can be poured via a wire. An electrically long connection to the Earth is useless at RF and often causes RF-related problems. Let's go back to what the problem really is: we have places in the shack where high RF voltage exists and RF voltage differences that cause RF current to flow. These problems can be addressed by *bonding*.

Bonding sounds heavy-duty (and expensive) but all it really consists of is connecting equipment enclosures together with short conductors so they have the *same* voltage. This is partially taken care of by the low-impedance connection provided by shields of coaxial cables between equipment.

However, accessories, computers, and power supplies generally aren't connected together with coax, so we have to provide another path. A common solution (shown in Figure 2) is to provide a wide, flat common *ground bus* at the back of or even under the shack equipment. Each piece of equipment, including computers and other non-radio electronics, is then connected to the bus with a short wire or strap.

[2]

To accomplish the purpose of RF bonding, no other connections are required.

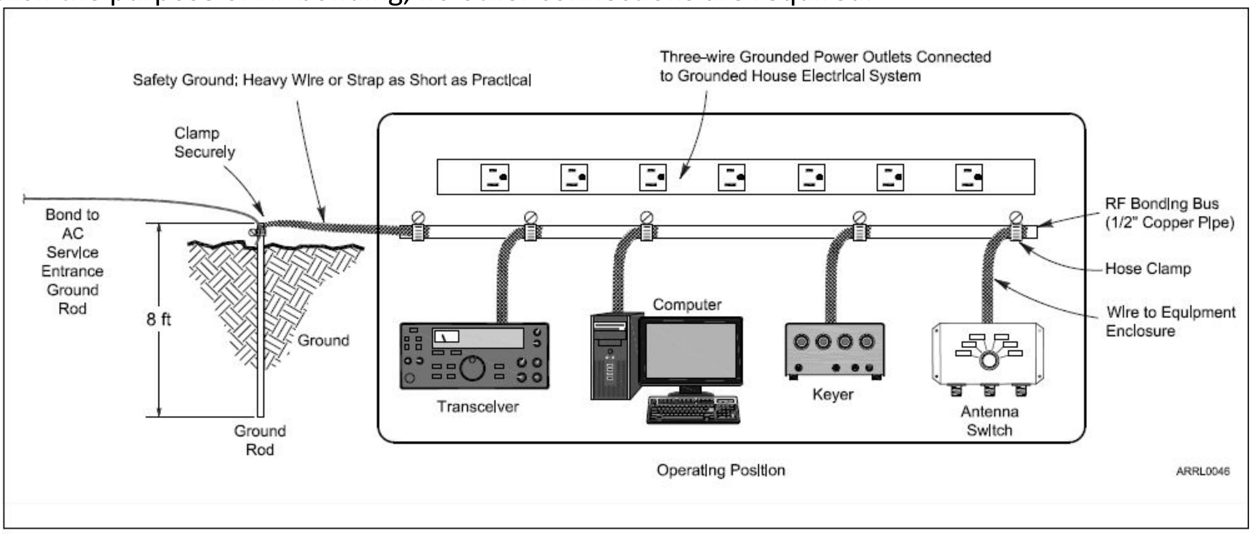

Figure 2 — The ground bus shown in this figure provides effective bonding between the various pieces of equipment at RF. The ac safety ground is also connected to the ground bus, but is of limited effectiveness at RF.

It is a good idea to add a connection to the ac safety ground from the ground bus or to use it as part of a lightning protection system, but the effect of that connection at RF will be unpredictable due to the configuration of the entire antenna system, as discussed earlier. From an RF perspective, keeping all of the equipment as close as possible to the same voltage is the important thing.

Bonding will *not* result in there being zero RF voltage on the equipment. Bonding does keep all of the equipment at about the *same* voltage, so RF current flow between pieces of equipment is greatly reduced with the added benefit of reducing the effect of lightning-caused voltage surges that affect the station ground connection. [3]

Next month we will talk about shielding, both of devices and of cables, and how this affects resistance to RFI.

Notes

[1] Poorly shielded connections and equipment will radiate RF from internal electronics, too. This creates on-the-air interference.

[2] Avoid using coax braid as an RF connection. Once removed from the protection of the confining jacket, the braid's individual strands begin to loosen and corrode, increasing the impedance at RF. Use heavy wire or solid strap, such as copper flashing, for best results.

[3] W. Ronald Block, KB2UYT, "Lightning Protection for the Amateur Radio Stations, Parts 1-3," Jun, Jul, and Aug 2002, *QST*.

Experiment #145 — Grounding and Bonding Systems

Along with effective receiving and transmitting, a station's design must address three important electrical requirements: ac power safety, lightning protection, and equipment-to-equipment bonding. Figure 1 shows that the frequency ranges involved are quite different, requiring different solutions.

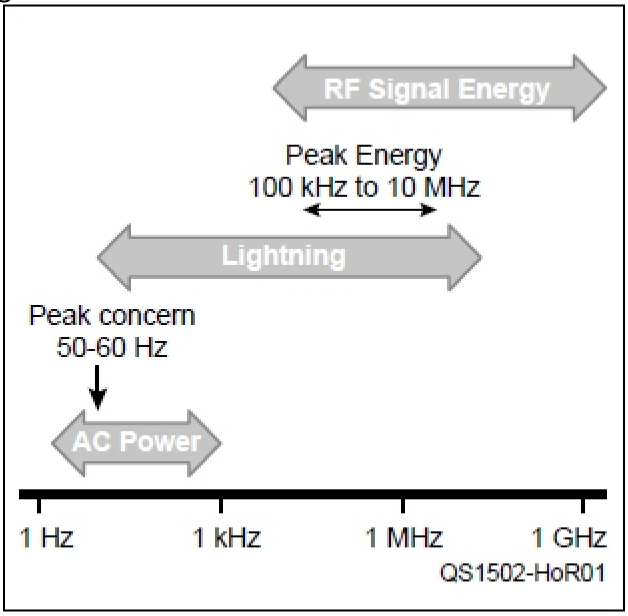

Figure 1 — The frequency ranges of concern to amateurs for ac power, lightning, and RF signal energy.

As you learned in the previous "Hands-On Radio" column, however, few parts of an amateur station exist in isolation, particularly as frequency goes up. Thus, the measures taken to satisfy these requirements can affect and even compromise each other if performed improperly. Standard methods of protection from shock hazards and lightning can be found in the *National Electrical Code* (NEC).
[1]

Low-Frequency AC Safety

Reducing the detailed language of the NEC to simple terms, you must connect all exposed metal from ac-powered equipment to a central, common ground. All equipment electrically connected to an ac-powered device should have a permanent safety connection even if the equipment is unpowered, like an antenna tuner or audio switch. (Don't depend on removable cables for safety connections.)

This connection is usually referred to as the *ac safety ground*, and the connection is made through the "third wire" of your ac power wiring; the bare or "green" wire. Figure 2 shows how power and enclosure grounds are connected through the ac ground wire to the breaker panel's ground bus.

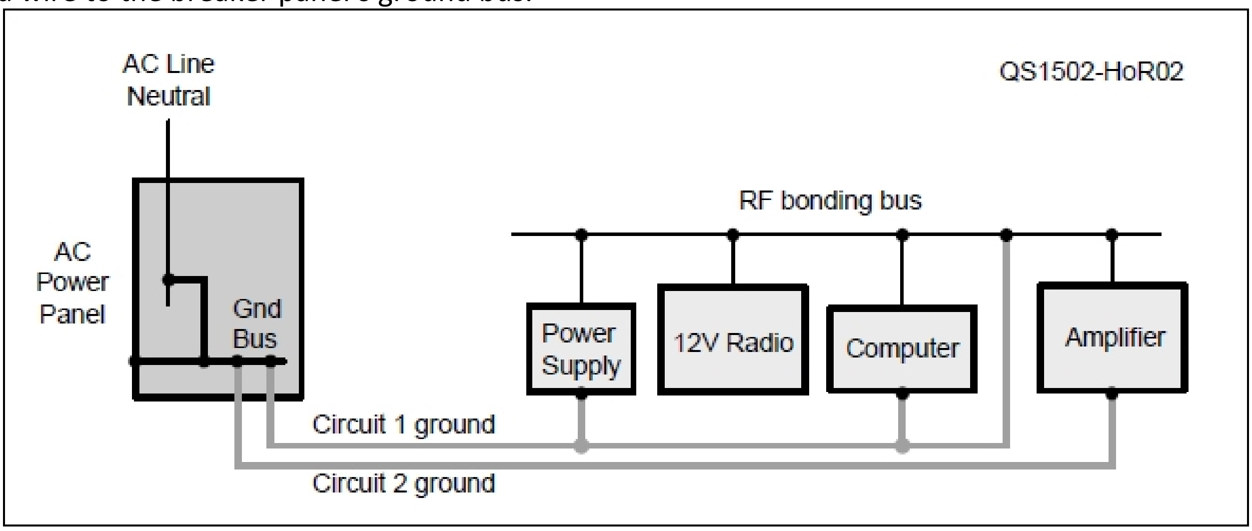

Figure 2 — AC safety grounding routes fault and leakage currents that present shock hazards away from a human operator. Most safety grounding is done through a home's ac wiring.

If a short circuit develops between the ac wiring and the enclosure of a piece of equipment, the resulting *fault current* in the ac safety ground trips a circuit breaker in the hot conductor. *Leakage current* is mostly due to filter capacitance and parasitic capacitance between the *phase* (hot) ac conductor and the equipment chassis or enclosure.

Ground Fault Circuit Interrupter (GFCI) breakers go one step further and monitor the balance of current on the hot and neutral lines of an ac circuit. If an imbalance is detected, it is assumed that the missing current is flowing on the equipment chassis or enclosure, where it can present a shock hazard, and the GFCI trips to remove power.

Since the ac safety ground is concerned with currents at the power line frequency and its first few harmonics, the length of the connection doesn't matter much. At 60 Hz, a 50-foot conductor is only 3 millionths of a wavelength long. Similarly, resistance in the ac safety ground path or an imbalance of a few ohms between circuits doesn't matter much from the perspective of safety. (Imbalances might be significant for signal-level connections.) The important thing is that hazardous current takes a path that doesn't include *you*!

Lightning Protection

Another area of concern also addressed by the NEC is that of lightning protection. [2]

The goal of lightning protection wiring is two-fold. First, a low-impedance path to the Earth and between earth connections is provided for the high currents involved, often thousands of amps. Second, bonding keeps all equipment and circuits at close to the same voltage so that voltage differences and current flowing between them are minimized.

Lightning energy is distributed mostly between 100 kHz and 10 MHz — far higher in frequency than ac power — so that wiring inductance dominates the performance of the protection system. Voltage across a conductor is equal to its inductance multiplied by the rate-of-rise of the lightning's current pulse in amps/second. For example, a straight wire 1 meter long has an inductance of about 1 μH. If the lightning's current rises 20 kA in 2 μs, the voltage between the ends of the wire is 1 μH × 20 kA / 2 μs = 10,000 V! Voltage transients large enough to do significant damage can also be *induced* on your wiring by the magnetic field of nearby strikes as well as being conducted to your wiring by ac power, phone, or cable TV systems.

Figure 3 shows the basic approach for lightning protection. AC safety ground wiring is included, as shown in Figure 2. The additional *service entrance ground* connection from the ac power distribution panel to a ground rod establishes a local earth connection for lightning protection of ac-powered equipment and appliances. In the shack, ham equipment-to-equipment bonding is also shown with its required connection to the ac safety ground.

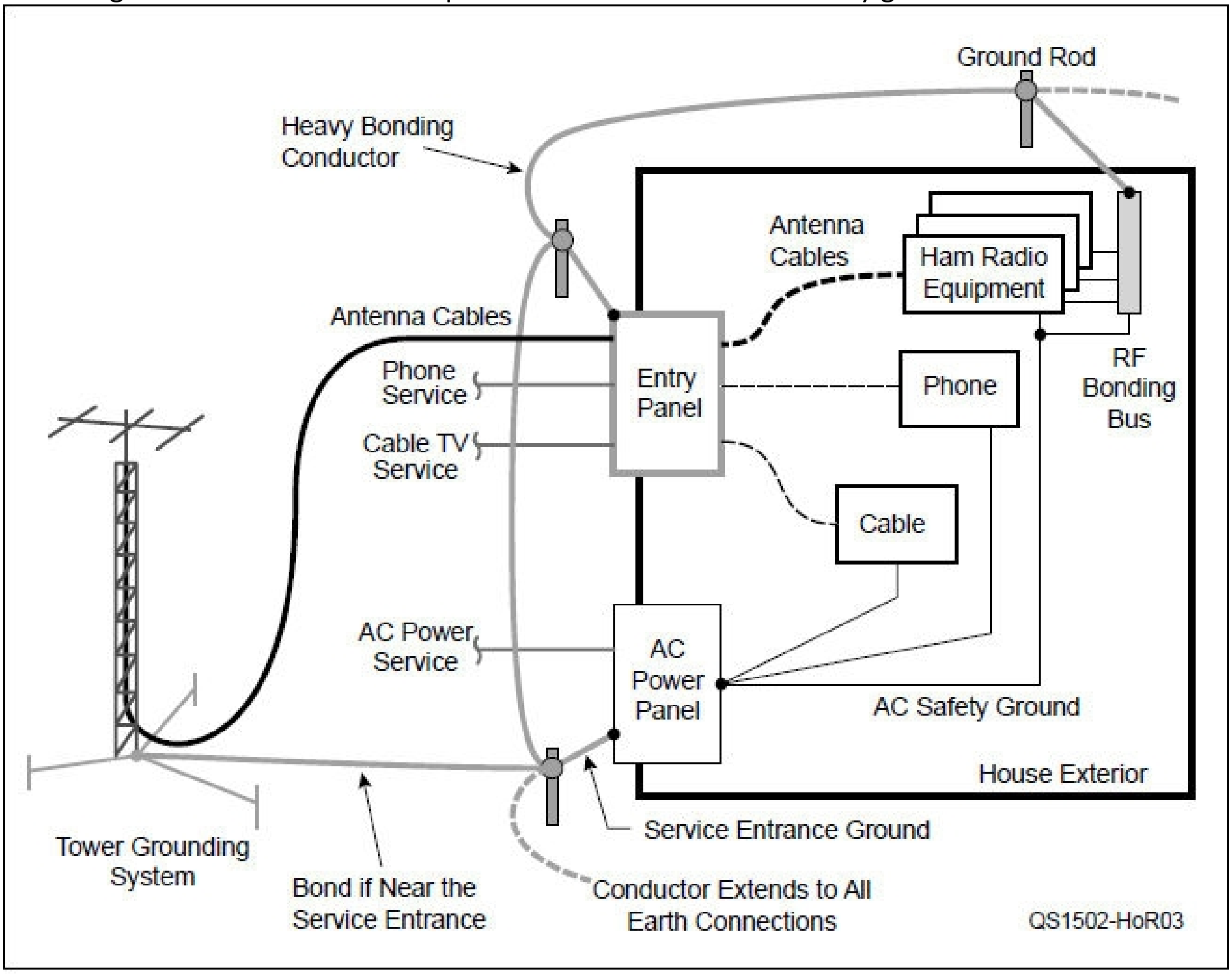

Figure 3 — A lightning protection system attempts to dissipate charge into the Earth outside the home or station. All earth connections are bonded together to provide a low-impedance path for lightning, and to minimize voltage differences and current flow between equipment and circuits.

Any earth connections at the ac power panel, other service entries, or an earth connection to the shack RF bonding bus must be bonded together by heavy copper wire or wide copper strap to provide a mechanically secure, low impedance connection. Using #6 AWG or heavier wire is recommended and a minimum size may be specified by your local building codes.

All antenna cables (including feed lines and rotator control cables) should be routed through an entry panel where they can be protected by lightning arrestors to route high-voltage, high-current surges to an earth connection. This minimizes the voltage between cables and the resulting current that would flow through the equipment connections as a result.

Other services may have their own entry panels and earth connections.

Each separate earth connection for safety, or to manage RF energy, creates a path through the house for lightning and can result in significant voltages from lightning surges or ac power faults. To minimize these voltages, bond all external earth connections together. The ground rods and bonding conductor provide a low impedance path bypassing the house for lightning's charge and keep equipment close to the same voltage at low frequencies. The goal is to dissipate as much lightning energy as possible *outside* while minimizing large voltage differences and current surges that damage equipment *inside*.

The extra ground system shown next to the tower helps dissipate charge, too. This is particularly important if your tower is not located close to the house and its bonded earth connections. If the tower is used as a vertical antenna, a radial ground screen can help spread out charge from lightning along with its RF function. What a great reason to create another antenna!

Tying It All Together

Let's review. All of your exposed connections to equipment powered from the ac line, or connected to such equipment, need to be connected to the ac safety ground. All external earth connections must be bonded together outside your home with heavy wire. If possible, use a single, grounded entry panel for all signal connections entering your house. If multiple entry points are used for different services, connect those panels or lightning arrestors to the outside bonding conductor. Finally, inside your shack, bond the equipment together with short, heavy connections to a common bus.

The Safety chapter in *The ARRL Handbook* and the ARRL's Technical Information Service web page on safety (**www.arrl.org/safety**) both contain lists of useful references, guidelines, and tutorials.[3] W8JI's website (**www.w8ji.com/station_ground.htm**) discusses station grounding and shows examples of an entry system and wiring practices. K9YC has published a slide-show tutorial on grounding (**www.audiosystemsgroup.com/GroundingAndAudio.pdf**) covering a variety of concerns.

Know the Difference

What we often refer to as "grounding" is really three different functions: ac power safety, lightning protection, and managing RF energy. Knowing the difference will help you build and maintain a safe and effective station.

Notes

[1] The NEC is a good set of reference standards but your local city or county building department may have additional requirements based on the specific circumstances in your area. The *authority having jurisdiction* (AHJ), such as a city or county building department, determines what is required at your location.

[2] NEC Article 810, "Radio and Television Equipment," Section III covers "Amateur and Citizen Band Transmitting and Receiving Stations — Antenna Systems." Bonding requirements are discussed in 810.58, and towers are discussed in 810.15.

[3] *The ARRL Handbook*, 92nd edition, ARRL, Chapter 28.

Experiment #146— Notes on Bonding and Shielding

One of the challenges facing the average ham trying to build a station is that the information needed is not collected in one place. Electrical safety (ac safety grounding) is discussed in one place, while lightning protection is discussed in another, and controlling RF is in yet other places. As we've seen, however, the needs for all three have much in common, including the physical wiring. The ham needs to read up on all three and approach station-building with a comprehensive strategy.

Bonding Conductors

One of the more popular materials hams use for "ground" connections is braided strap or shield braid removed from old coaxial cable, but that is *not* a good idea. Here is the problem — the wires in a coaxial cable shield are pressed together and protected from moisture and contaminants by the jacket. Once the jacket is removed, the wires slowly begin to corrode and the weave loosens. Pretty soon all those individual tiny wires start acting less like a large, flat surface and more like, well, individual tiny wires with poor connections between them. This is not very good for conducting RF.

For bonding at RF, use heavy wire or strap, such as copper or aluminum flashing. Solid wire, such as #14 or #12 AWG from home wiring cable, or stranded THHN wire works fine. The important thing is to maximize surface area because of the skin effect at RF.
[1]

For dc connections and for ac safety grounding, which are more concerned with resistance, braided strap is acceptable. Braid should also be used for bonding in high-vibration environments, but that's not necessary for most ham shacks, even mobile installations. For lightning protection and bonding earth connections together, use heavy wire or strap to minimize impedance and for mechanical strength.

Star Versus Bus Configurations

The answer to the question, "Should I use a star or a bus connection?" (see Figure 1) depends on what you are trying to accomplish. At dc and low frequencies, where resistance is the primary consideration, both are equivalent and the star connection is usually more convenient. At the higher frequencies involved for lightning protection, inductance becomes the most important characteristic. A star configuration with reasonably short connections will still provide adequate bonding.

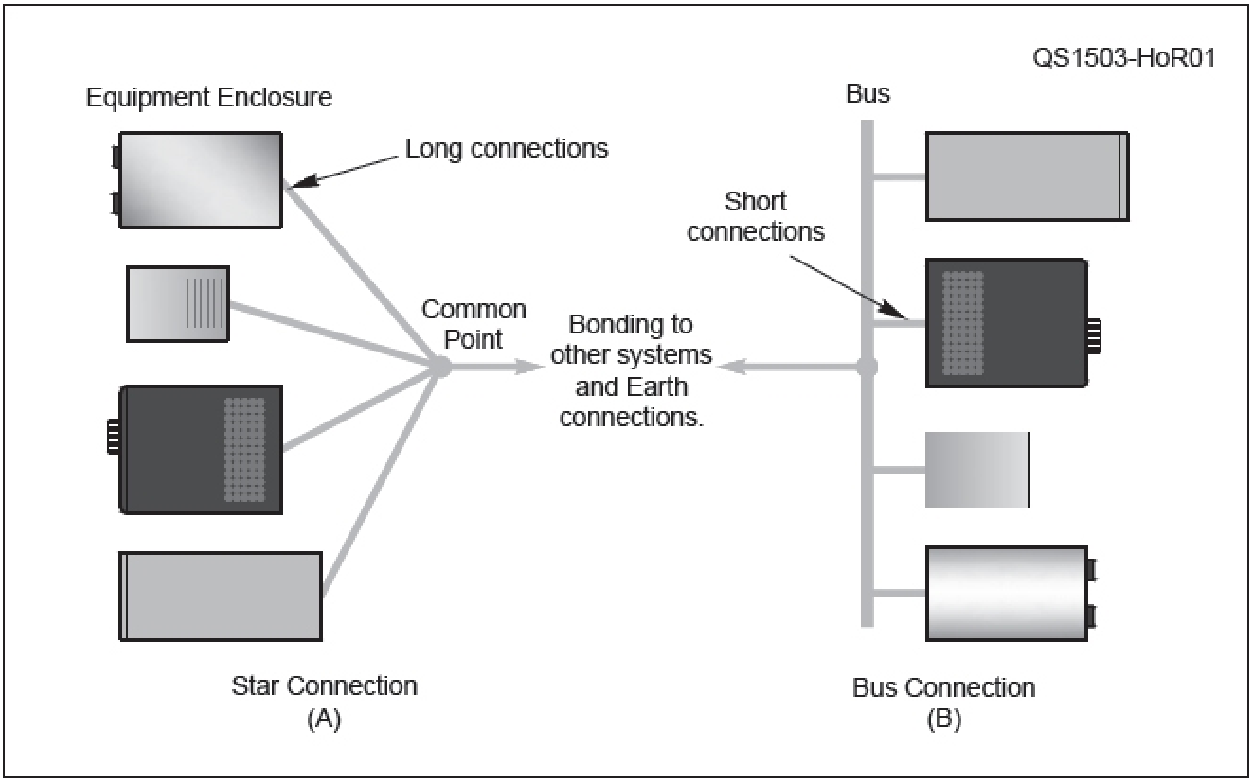

Figure 1 — The star connection at (A) and the bus connection at (B) accomplish the same safety functions for dc and ac power. As the frequency of interest increases, minimizing connection length and loop area makes a bus configuration preferable if compact.

At RF, however, the electrical length of the connection is the dominant consideration. The bus and star are trying to accomplish the same thing — keeping all of the equipment at the same voltage — but the bus does the job with shorter connections.
[2]

The length of the bus must be kept to a minimum. Connecting the enclosures directly together with individual wires is an alternative to using a separate bus conductor.

While trying to address ac safety, lightning, and RF control sounds complicated, minimizing the physical length of bonding connections addresses all of these concerns in the typical strong RF fields of a ham station. Bonding enclosure to enclosure using short conductors between interconnected equipment works for all three needs.

Minimizing Loop Area

The inductance of a one-turn circular loop is directly proportional to the natural log (ln) of the loop's radius: the bigger the loop, the greater the inductance. As inductance increases, so will the voltage between any two points in the loop whether the voltage is induced by a magnetic field or a rapidly changing current pulse.

What loop is this, you ask? The typical ham station is *full* of loops created by the shields of interconnecting cables, antenna system and control cables, ground connections, etc. Figure 2A shows the basic idea, and there are many more loops than the ones indicated in the drawing. Every complete conductive path around enclosures and cables counts as a loop.

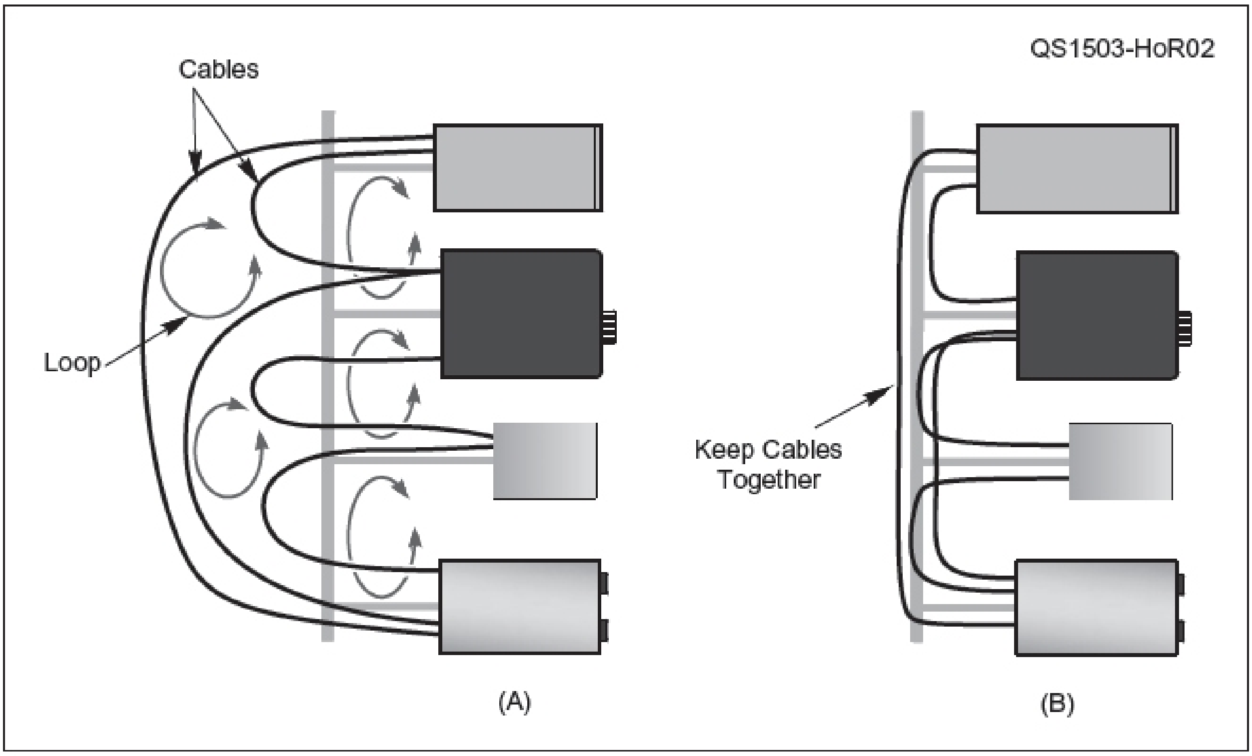

Figure 2 — Each conductive path through enclosures and cables creates a loop (A) that can pick up and radiate signals. Minimizing cable length and loop area by keeping cables together as at (B) can reduce pickup and radiation of RF.

Eliminating these loops is not realistic. Minimizing their area, on the other hand, is a productive strategy. First, use the minimum cable length. If you have two pieces of equipment 6 inches apart connected with a 6-foot cable, replace that cable with a shorter one and coil up any extra length.

Next, minimize the area of each loop. Where several cables run in the same general area, use wire ties or a cable tray to hold them close together. Separate wires for power and ground should be twisted together, as should audio connections to speakers. If you use a bus for bonding equipment enclosures together, consider running your interconnecting cables along the bus as in Figure 2B, further reducing loop area.

Connecting Shields

Most hams understand the need to connect the shields of RF-carrying coaxial cables to metal enclosures, as in Figure 3A. One analogy is that of creating a "water-tight" connection as if the differential-mode RF signal was a fluid to be kept within enclosures and cable shield. This keeps any contaminating common-mode RF currents on the outside of the cable or enclosure. It also eliminates unwanted radiation of the RF signal from the internal electronics.

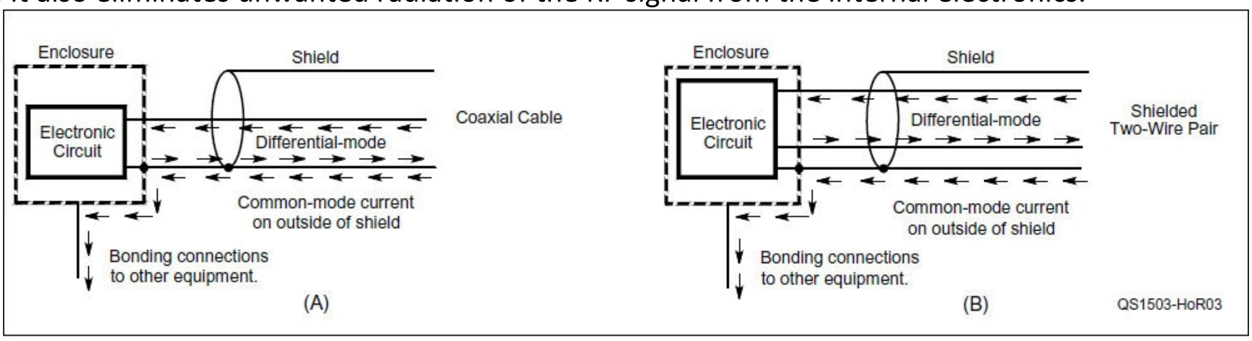

Figure 3 — It is important to keep common-mode RF currents flowing on the outside of cable shields from entering electronic equipment. Similarly, proper shield connections prevent RF signals generated by the equipment from being radiated by the cable.

The same concerns apply to unbalanced data and audio connections. Figure 3B shows a two-wire connection with a shield, such as for one circuit of an unbalanced RS-232 interface. From the perspective of both susceptibility to interference and interference-causing, it is important that the external shield be connected to the enclosure and that any separate signal ground connections *not* connect to the enclosure. Keep internal signals and RF inside the cable and external RF out of the cable. Using twisted-pair cabling also helps prevent RF pickup and radiation.

Unfortunately, it is common practice in amateur equipment for multi-circuit connectors such as mono and stereo phone plugs to have their shield connections (usually referred to as the *sleeve*) isolated from the metal enclosure. Instead, the shields are often connected to an internal common point on a circuit board which is then connected back to the enclosure. Known in pro audio circles as the "Pin 1 Problem" because shields of audio cables are connected to Pin 1 of the standard XLR connector, this routes RF current on the outside of the cable shield to the internal electronics where it can do the most damage. At the same time, RF noise and harmonics from the electronics are routed to the outside world.

Worse, shields are sometimes left completely unconnected. This allows the RF current to enter the cable and flow into the electronics of whatever is connected. For a demonstration, turn on your HF transceiver and connect it to a dummy load. Even a short cable will do. Turn off any preamp or front-end filtering.[3] Note the level of background noise on, say, 20 meters. Now unscrew the shell of the PL-259 at the transceiver, leaving the center pin connected. The dramatic increase in noise is from signals being picked up by the cable shield and entering the cable to become differential-mode signals at the open end of the shield.

The ARRL's Safety page (**www.arrl.org/safety**) and RFI page (**www.arrl.org/radio-frequency-interference-rfi**) list many references and detailed articles to teach more about these topics.

Notes

[1] *The ARRL Handbook, 92nd edition*, (Newing-ton: 2014), "Chapter 5: RF Techniques," p 5.1.

[2] See the Hands-On Radio web page (**www.arrl.org/hands-on-radio**) for Experiments #145 and #146.

[3] If the receiver has a high-impedance input, this demonstration may require a 6 to 10 dB 50 Ω attenuator at the inputs.

Experiment #149 — Accidental Mixers

When reading about RF interference and intermodulation (aka "intermod"), you will often encounter references to "bad connections" and junctions between dissimilar metals that create rectifiers. Since the discussion usually takes place in the context of sites with multiple transmitters for commercial, broadcast, public safety, and/or ham repeaters, most hams file the information in the "It Can't Happen Here" folder. I discovered that it certainly *can* happen "here," meaning "at my station" and, of course, my education came by accident and from a completely unexpected source.

A Bad Case of RF Grumbles

A few years ago, I put up a new tower and proudly topped it with a two-element 40 meter Yagi (a 40-2CD), a first for me. Like most projects, this one proceeded in stages — the tower went up first, then I put the antenna together and mounted it on a stub mast without a rotator yet installed; that would come later. In the meantime, a "rope-tator" did the job. All was well, the band was full of signals, I worked a bunch of DX, and things were going my way. Next, I installed the Tailtwister (T-2X) rotator — no more rope-tator.

The very next morning, I eagerly switched on the rig but — *zut alors!* — instead of a normal background crackle, the band was full of S-3 grumbles and garbled voices from end to end. What was this? I looked inside and outside the shack, checked connectors, accessories, and equipment settings, and found nothing. Resigning myself to a long detective adventure, I sat down at the rig and pressed the rotator control box's Brake Release switch and the grumbles changed to a 120 Hz buzz. Eh?

Sure enough, each time the brake solenoid's 24 V ac power was applied, the new noises changed to a buzz. I checked the rotator cable, but all connections were solid and resistance measurements through the rotator itself were in the normal range. Staring at the schematic (Figure 1), it finally dawned on me what was going on. The rotator cable was acting as a terrific random-wire antenna, happy to pick up the local AM broadcast station signals and convey them to the control box. (My location on Vashon Island, Washington, was home to seven multi-kilowatt stations within 5 miles of my station!)

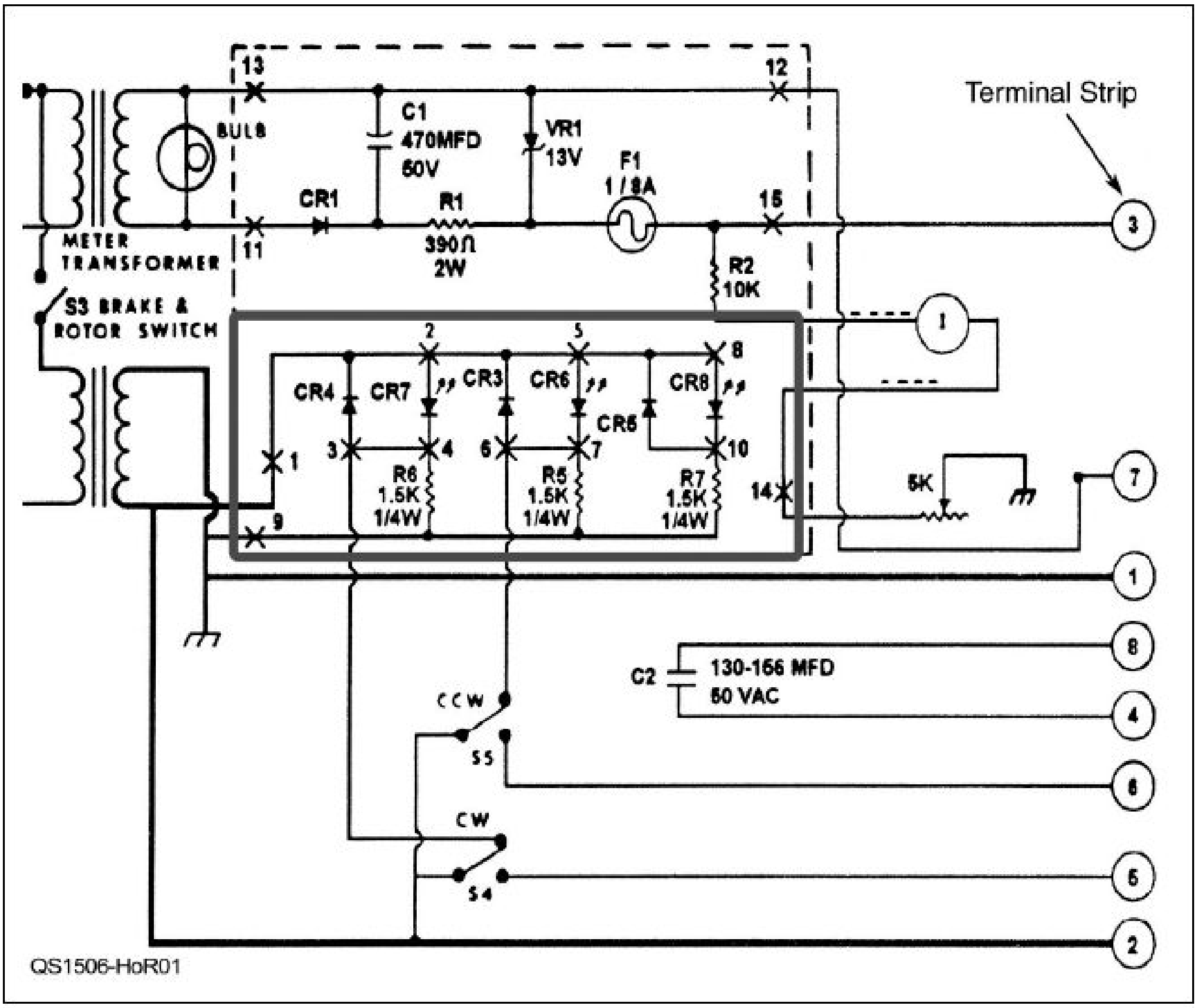

Figure 1 — Schematic of Hy-Gain Tailtwister T-2X rotator control box. Rectifiers CR3 – CR8 are connected directly to the external control cable and can act as mixers or harmonic generators for RF picked up by the cable.

The Accidental Mixers

Inside the control box are eight rectifiers, CR1 through CR8, plus a Zener diode, VR1. The problem was being caused by CR3 through CR8, which are connected to the rotator control cable through the direction control switches S4 and S5. Volts of AM broadcast RF were applied directly to those rectifiers, which happily mixed all of those signals together and re-radiated the mixing products across the HF spectrum via the same cable. (I could hear them all the way up to 20 MHz or so.)

With the brake solenoid OFF (S3 open), the rectifiers acted as passive mixers. With the brake solenoid ON (S3 closed) the rectifiers were either cut off or forward biased by 60 Hz ac from the power transformer that turned on the various indicator LEDs — except at the zero-crossings that happened twice per cycle or 120 Hz. During the short window of time when the 60 Hz ac was near zero volts, the diodes could act as mixers again and so the grumbles I was hearing on 40 meters appeared as pulses at 120 Hz.

Now that I knew what the problem was, the solution was simple. I soldered a 0.01 µF, 50 V disc ceramic capacitor across each diode in the control box. No more grumbles! The capacitors routed the RF around the rectifying PN junctions that were acting as mixers and harmonic generators. Since my location was so hot with BC RF, I took the next step and added another capacitor from each terminal strip position to the metal control box enclosure's outer surface.

Accidents Waiting to Happen

I also realized that the T-2X control box was not the only device in my station with unbypassed rectifiers. I had several accessories, some commercial and some home-built, with LEDs or rectifiers connected directly to control or power cables. A few minutes work with a screwdriver and a soldering iron and all those diodes were "cold" at RF.

Since I hadn't noticed problems before, what difference did adding the capacitors make on the remote switches? Did I notice any improvement? As far as normal listening, the additional bypassing didn't change background noise levels dramatically. However, what I *did* notice was that some of the harmonics from my transmitted signal were now much weaker on some bands.

I am a fan of SO2R (single operator, two-radio) operating in contests (
www.dxcoffee.com/eng/2011/09/07/tutorial-so2r-operating-mode
or
www.k8nd.com/Radio/SO2R/K8ND_SO2R.htm
), in which I'm calling CQ on one band and tuning for new stations on another. It's pretty obvious when you tune across your own harmonics! In some cases, these diodes had been generating harmonics by rectifying RF from the transmitter and re-radiating it via a control cable. The investment of a few pennies per capacitor reduced some of the harmonics quite a bit. (Some were generated by the transmitter directly.) The same improvement applies any time your station is being used for transmitting and receiving at the same time. Multioperator station owners and EOC station managers, are you listening?

Where else besides antenna switches and rotator control boxes can you find these "accidental mixers?" Many places, in fact, and you can see some in Figure 3. All of these circuits are often attached to external control or power cables that act as antennas to pick up RF and re-radiate mixing products or harmonics from rectified RF. Attaching a bypass capacitor across the diode renders it inert at RF.

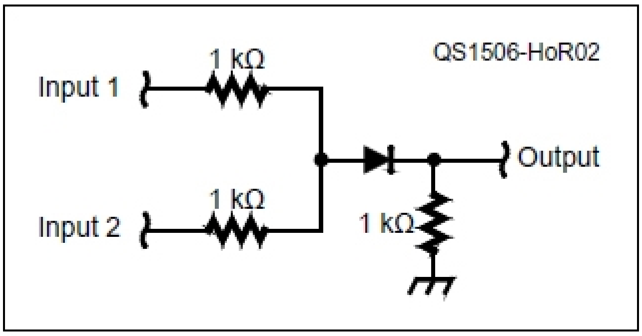

Figure 2 — This simple single-diode mixer illustrates what happens when one or more ac signals encounter a non-linear device, such as a PN junction. Any type of diode can be used. The different I-V characteristics will change the amplitude of harmonics and mixing products.

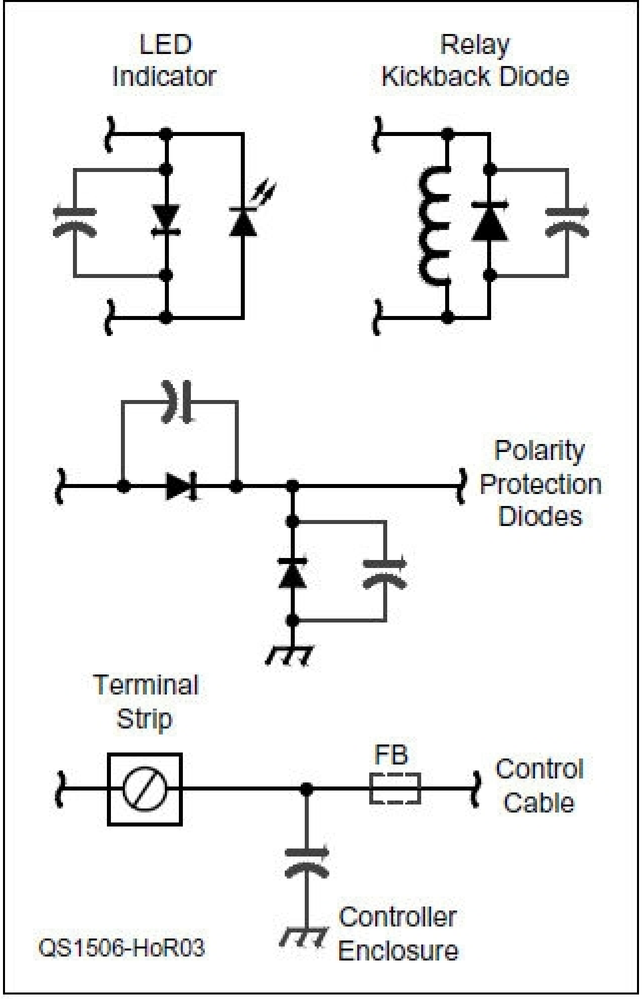

Figure 3 — Several examples of rectifiers in applications where they might be subjected to strong RF signals with methods for connecting bypass capacitors. The ferrite bead shown (FB) can be one or more beads or a choke wound on a toroid or split core.

Look around your shack for power connections and control cables, then be sure the equipment includes measures to block or bypass RF that might be present on those cables. You can also block RF where a control cable enters an enclosure. Connect the bypass capacitors and ferrite beads, cores, or chokes on the *outside* of the enclosure so that RF never gets into the equipment.

Seeing For Yourself

You can see the effects, too, by connecting the circuit in Figure 2 to some signal generators. Use a silicon 1N4148 diode at first. Apply an audio signal of at least 1 V RMS to Input 1 and use an audio spectrum analysis program such as one of those available at www.dxzone.com/catalog/Software/Spectrum_analyzers. Vary the amplitude and frequency of the signal to see the effect on harmonic content. Now connect a second signal to Input 2 and look at the mixing products — yikes! (The output of an HF receiver tuned to an AM carrier makes an impromptu audio signal source.) Try different types of diodes, such as germanium (1N34A) or Schottky (1N581x series), or an LED to see how various types of PN junctions influence the resulting spectrum.

Field Day Stub

The strong signals encountered at Field Day can bring a normally well-behaved receiver to its knees. Phase noise from a transmitter can wipe out a nearby receiver, stressing out operators. A simple transmission-line stub filter could make all the difference. Check out "Hands-On Radio Experiment #22 — Stubs" for a nifty and inexpensive switchable feed line stub band-pass/band-reject filter that works on 40 – 10 meters![1]

[1] All Hands-On Radio experiments are available to ARRL members at www.arrl.org/hands-on-radio.

Avoiding Field Day Accidents

Even if you never transmit and receive at the same time from your home station, taking your gear to Field Day can be a wholly different story! At a multi-station site with antennas close together, RF pickup can be quite strong. Station configuration is often, well, “Field Day-style,” meaning more or less a lash-up, with RF current flowing everywhere. Making your gear “RF-proof” is a great idea *before* you get to the Field Day site or the public service event. For the cost of a few capacitors, you can save yourself a lot of headaches!

Practical Station Practices

Experiment #130 — Communication Speakers

Okay, okay — enough with the math and the phasors and the coordinates! Several columns dedicated to phase rotation and spinning around at the carrier frequency is enough to make anyone a bit dizzy, the author included.

[1]

I'm sure we all need something more on the order of drilling and soldering, so let's return to the workbench and cobble together an accessory that has a home in every mobile station — the communications speaker. But we'll jazz it up a bit.

We all get speakers built into our mobile rigs. However, they are often chosen simply because they will fit in the box and not because they are the best solution for competing with wind and road noise in a vehicle, usually while trying to understand the limited fidelity speech of another operator, who may also be driving.

An external communication speaker is substantially larger and able to reproduce speech with better fidelity at volumes that can overcome ambient noise. Many vendors offer fine products in this regard and if what you need is only one speaker for one radio, that's probably the right solution.

But as our friends and family well know, one radio is rarely enough! Getting separate speakers for each rig leads to clutter, as well as a volume arms race as each radio is turned up louder and louder to be heard.

The most common multi-radio mobile installation has a pair of rigs, perhaps a ham radio and a scanner (or, as in the author's car, a VHF/UHF mobile FM transceiver and an HF rig). If you only have a single speaker, it's simple to add a switch and select one or the other. Yet it's pretty common to have both radios on at the same time — perhaps you are operating HF and keeping an ear out for a call on the repeater or vice versa.

Passive Mixers

The simplest way of being able to satisfy the requirement of listening to either or both radios at the same time is to substitute a *balance* control for the A/B switch. By adjusting the balance control, you can listen to either radio A, radio B or a combination of both. Balance should not be confused with *pan* (from "panoramic") which refers to positioning a particular audio source in multiple audio channels (
thedawstudio.com/Tips/PanPots.html
).

Figure 1 shows a very simple way of being able to listen to either or both radios in a single speaker using a passive balance control. The speaker outputs of most radios can supply a few watts of audio power into the typical communications speaker impedance of 4 to 32 Ω. A fixed resistor is in series with each speaker output to isolate the individual radio audio outputs from each other. The variable resistor is connected so that when the wiper is at either end, the speaker gets full output from one radio and very little from the other.

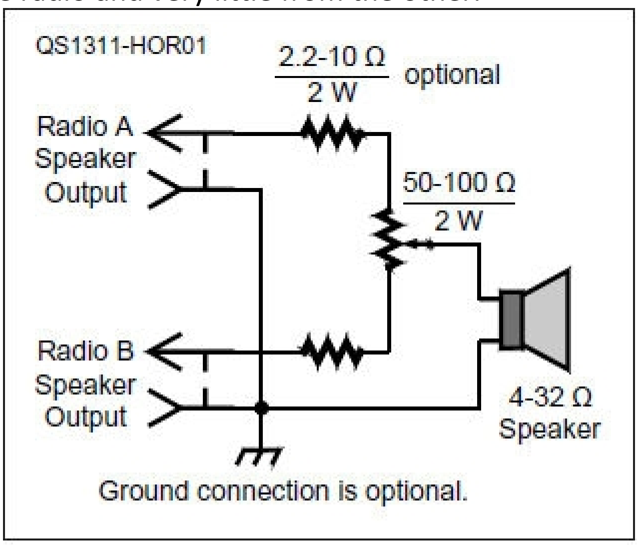

Figure 1 — A basic passive mixer isolates the radio audio outputs with the fixed resistors. The radio volume controls determine overall level and the potentiometer acts as a balance control. Small-value resistors insure the radio outputs aren't connected directly together.

Because of differences in radio audio outputs, available speaker impedances, and your personal volume preferences, a range of values is shown on the schematic. If you want a lot of volume, choose lower values for the fixed resistors, with the tradeoff being higher minimum volume for the undesired channel. Wirewound resistors are fine in this low-frequency application. This simple circuit can be installed inside the speaker housing, too.

It may take a couple of tries before you get the right combination of volume settings on the radios and resistor values in the mixer. Because the radios are different, the fixed resistor values may need to be different, too. In fact, the fixed resistors can be replaced by potentiometers if you like.

Active Mixing

A more flexible method of controlling the volume from more than one source through a single speaker is to use an *active mixer* — an amplifier that combines audio from multiple inputs with each level adjustable in the output. There are many types of active mixers, ranging from a simple summing circuit based on an op-amp [2] to sophisticated designs with two or more output channels that have pan, balance, and frequency equalization on each input. These may have a place in the well-rounded shack at home but we're talking about your mobile station, so let's not go overboard.

You probably already have a type of mixer in your vehicle — it's part of the audio entertainment system. A standard feature on most vehicle audio systems these days is an AUX (auxiliary) input with a three-conductor, stereo and a 1⁄8 inch phone jack mounted somewhere on the dashboard or console. Plug in your stereo audio player or smartphone and away you go. The audio from your ham rigs can be plugged into the stereo, too, if you make sure to keep the signal levels down. Figure 2 shows an audio attenuator circuit suitable for use with the audio system's AUX input.

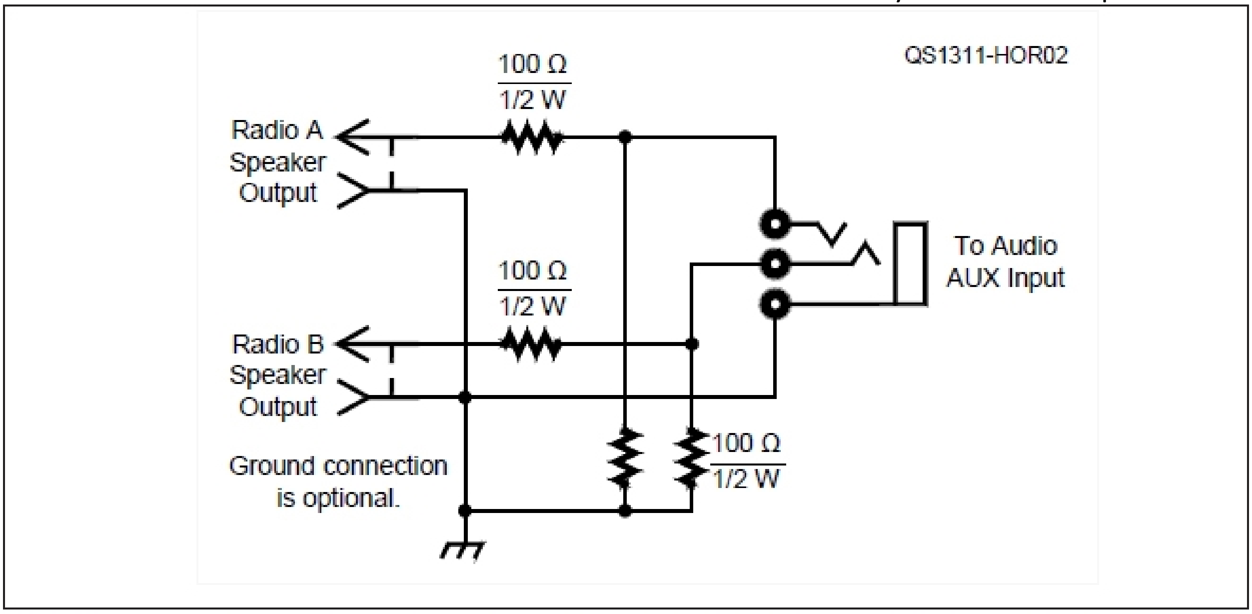

Figure 2 — This fixed attenuating divider limits signal level into a vehicle audio system's AUX input. Audio system volume and balance controls are used to adjust audio volume.

In this circuit, both radio outputs are still connected to a fixed resistor but now they are not connected together in the output balance control. Instead, a resistive divider limits the signal level into the AUX input. If the load connected to the resistive dividers is 32 Ω, typical of small headphones or earbuds, the voltage from each channel is attenuated by about 14 dB. How did I determine that? Since 32 Ω // 100 Ω ≈ 25 Ω (// is used to indicate "in parallel with"), the output is reduced by 20 log (25 / (100+25)) = 13.9 dB. If you need more volume, decrease the input fixed resistor value. A high audio system input impedance reduces attenuation to 6 dB.

With the output of the divider connected to audio system's AUX input you can listen to one radio in the left channel and the other in the right channel. The fidelity of my car's stereo system is a lot better than that of the speakers in the radio! Non-hams find the audio system's output a lot easier to listen to, as well. Hams have gotten used to really poor mobile audio with lots of distortion and no bass. You might be surprised at how good a radio can sound if its output isn't trying to overdrive a minimal speaker over the road noise.

Customize It!

Don't stop here — add more features. You can use a splitter at the AUX input for your audio player, but why not add a parallel jack and switch on the speaker housing? Add a headphone jack or adjustable resistors for independent level setting. Don't be afraid to experiment with different resistor values and configurations.

In my vehicle, I wanted to be able to switch the speaker on or off independently of the audio system so I could listen to both radios and some entertainment at the same time. Figure 3 shows how I have my circuit configured and Figure 4 shows a photo of the final product. The circuits of Figures 1 and 2 are connected "in parallel" to the radio audio outputs so that they can act independently.

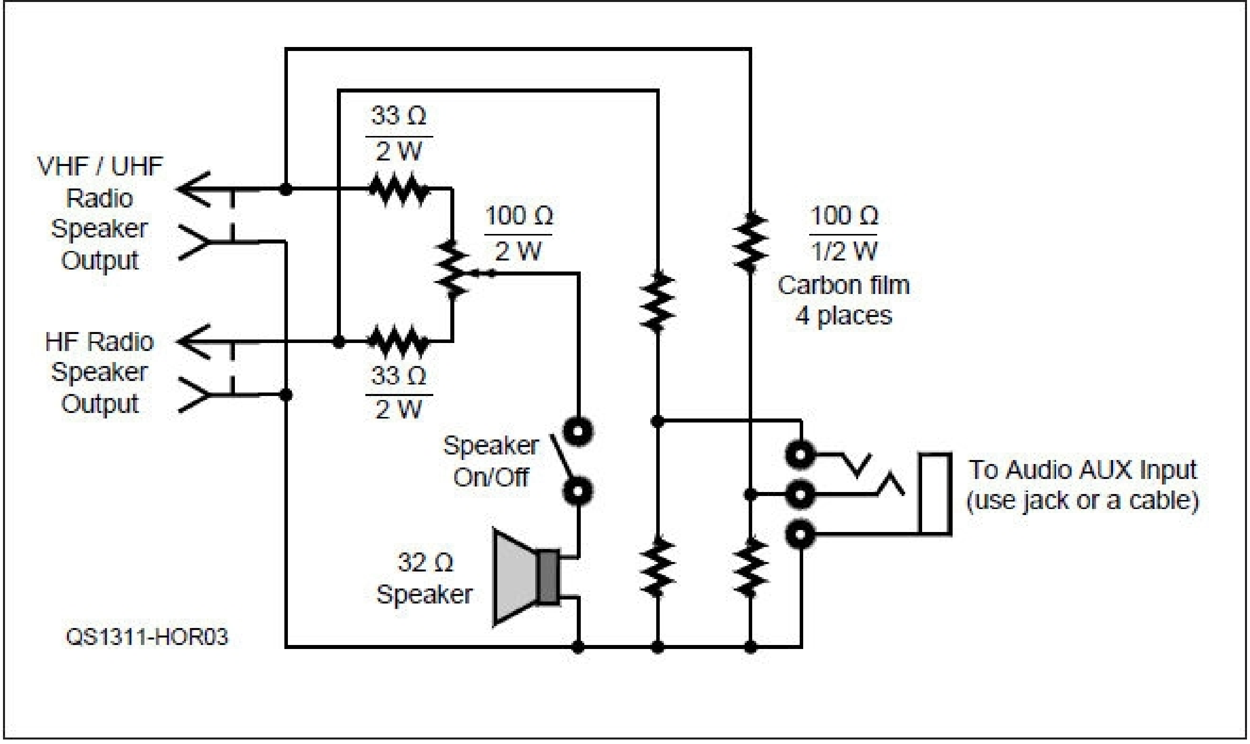

Figure 3 — The author's combined system that allows monitoring through the speaker or through the vehicle audio system.

Figure 4 — The completed speaker with the volume control, on-off switch, and cables for connection to the radios and vehicle audio system (white connector). [Ward Silver, NØAX photo]

I used what I had in my junk box. These values were "close enough" for reasonable radio volume settings. It got the job done and let me proceed with hamming it up. As a bonus, I have to say that listening to a big CW pileup with the widest IF filters when it's played through a powerful audio system is some kind of amazing. It's not quite opera, doc, but it's close!

Notes

1

All previous Hands-On Radio experiments are available to ARRL members at

www.arrl.org/hands-on-radio

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2

Silver, Ward NØAX, “Experiment 3: Basic Operational Amplifiers,” QST, April 2003, pp 63-64.

Experiment #148 — Proof of Performance

"You want truth? You can't handle the truth!" goes the courtroom outburst from Jack Nicholson in A Few Good Men. The same can occasionally be said for finding out about how our station really performs! Until we get an unexpected surprise on the air or in our mailboxes in the form of an Official Observer report, many of us might not really know if our stations are doing what we think they are doing. How to find out? The broadcaster's proof of performance can be a powerful friend to the amateur station owner, and it's not as difficult to undertake as you may think, either.

What is a Proof of Performance?

Every broadcast and commercial station is required to regularly certify that it meets the technical standards required by the rules for the service under which its license is granted. Performing that ongoing certification — called a proof of performance — is part of what allows these powerful stations to transmit without constant engineering supervision. As we all know, things break, wear out, come loose, get water in them (or let the water out), corrode, are damaged — the ways in which radios and antennas can become defective are seemingly without number. Without an occasional inspection, it would be illogical to assume everything is functioning just fine. In fact, it's probably more reasonable to assume the opposite!

Radio magazine has scanned and published online two 1970s-era proof of performance manuals; one for AM stations and one for FM stations.
[1]

These manuals were typical of the documentation that guided thousands of "chiefs" (Chief Engineers) as they conducted the necessary testing, "…quite possibly the most important engineering assignment that most radio station chief engineers have…first of all, it demonstrates to the FCC that the station is at least able to meet minimum technical standards or…far exceed them. A satisfactory proof also assures the broadcaster that [the] facility is doing what it is supposed to do, at least technically." The manuals go on to show the required test equipment and its configuration, what tests are needed and how to perform them, and provides forms for the necessary data.

The goal of a broadcast facility proof is to make sure that the entire programming chain, from the microphones and audio inputs to the transmitted signal, were in good order. Such testing checks noise, distortion, and frequency response of all audio handling equipment. (TV stations have similar requirements for video channels, even the digital signals.) Similar RF tests are done to be sure there are no spurious emissions. AM stations might even take a field strength meter a known distance from the antenna and check to be sure enough signal was being radiated. All this for a station just using one channel!

Proof of Performance for Hams

A mini-version of what the professionals do should be part of your station checkup and inspection, too. The bigger your station, the more comprehensive your proof should be, and the more it will tell you. Certainly, this sounds like a bother, but when would you rather find out that, say, your signal is way down on a band or that your microphone is intermittent? During a contest or public service drill? Or before, when you can do something about it? The latter, of course.

All ham stations are somewhat different, making it impossible to create a simple manual for all to follow. Nevertheless, there are many common elements and techniques. Figure 1 shows a typical station as a block diagram from the power supply and audio sources at the left all the way to the radiated RF at the right. If you draw a circle around each "box," you'll cross one or more arrows. Those arrows and the action or signal they represent are what you test when conducting the "proof."

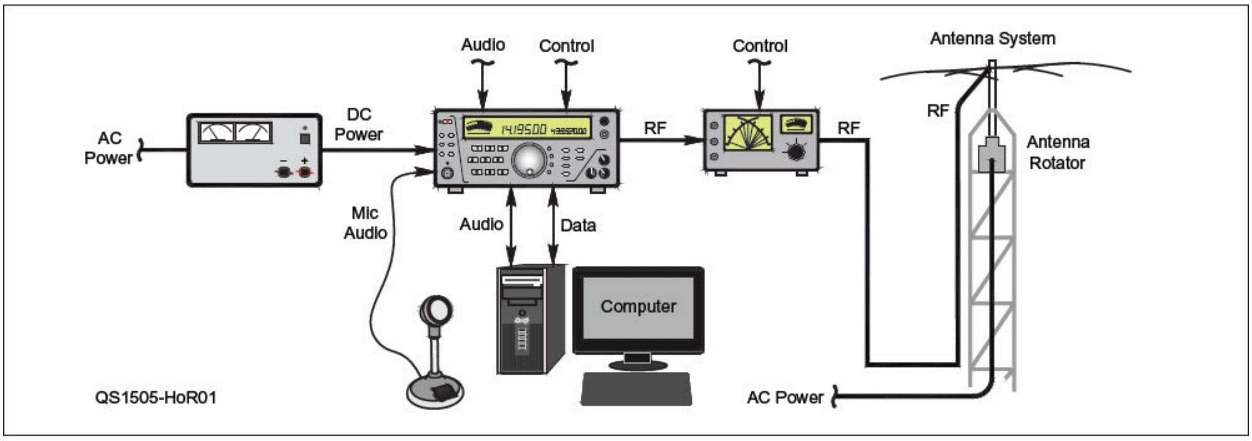

Figure 1 — A typical station showing the basic connections between pieces of equipment. Proof of performance testing consists of verifying that each interface and associated function works properly and in agreement with the equipment's specifications.

A Power Supply Proof

Let's start simple and look at the power supply. What that power supply is supposed to be doing is listed in Table 1. Notice that not only is a performance specification given, but certain test methods or equipment are listed, too, such as a voltmeter or testing under load. Let's give the supply a proof test like you might before taking it to Field Day. Open your shack notebook, (you do have a shack notebook, right?) note the date, copy Table 1, and start taking notes.

Table 1 Typical Power Supply Specs and Test

Spec	Value or Type	Test
AC input connector	IEC/UL, 3-wire, #14 gauge	Visual
AC input	90 to 150 V AC @ 60 Hz	Voltmeter
AC power control	Rocker switch	Manual
DC output connector	Binding posts	Visual
DC output	12 to 15 V DC at up to 30 A	Voltmeter on output under load
DC output	Ripple or noise	Voltmeter or oscilloscope
DC voltage control	Front panel rotary control	Voltmeter on output under load
DC overload protection	AC fuse	Visual and resistance check
DC current limit	30 A	Output jumper test

Crossing each of the power supply arrows in Figure 1 and starting with the ac input, check each item in turn. Yes, even the power cord! (You'd be surprised how often power cords get "temporarily" replaced with a light-duty cord, but are then forgotten about and are never seen again.) If your inputs are okay, proceed to the other arrow — the dc output — making sure all voltages are within spec, voltage adjustments work, and the output connectors are in good shape. Record that you tested all these things, the equipment you used to make the tests, and any values noted.

Load testing the power supply takes a little more planning. Assuming you don't have a large power resistor, a good method is to set the transceiver on full power, FM mode, and connect it to a dummy load. While monitoring the power supply output voltage, press PTT and measure output voltage. Keep the load on for a few seconds or cycle it while watching for voltage drops. Check the output connectors and cables for heating up. Again, record the results.

Expanding Your Proof

A proof of performance doesn't need to be incredibly detailed, but it should test or exercise the equipment features that you expect to work when you need them! An audio test can simply be an on-the-air radio check with a friend. Data interfaces can be verified to function properly by making a contact or by exercising a control or monitoring function. Table 2 lists typical equipment and some of the tests you can perform for it. Start simple and modify the list and tests as needed.

Table 2 Typical Equipment and Tests

Equipment	Function and Test
Transceiver	Input power: check cable and connector for voltage drop
Transceiver	Major controls and menu settings functional and correct
Transceiver	Computer interfaces functional
Transceiver	Microphone audio acceptable, drive level settings correct
Transceiver	Headphone and speaker audio acceptable
Transceiver	Keying interfaces work, no key clicks or spurs
Transceiver	RF output: power levels okay at full and minimum power, all bands, all modes
Transceiver	Harmonics and out-of-band spurs meet specifications, all bands, all modes
Transceiver	Modulation: signal quality check on, all bands, all modes
Antenna Tuner	Switches and controls: functional, not intermittent
Antenna Tuner	Preset adjustments are correct, no intermittent adjustments
Antenna Tuner	Full power test
Antenna Rotator	Direction calibrated
Antenna Rotator	Rotator works from limit to limit with no dead spots or jumps in indication
Antenna System	All SWR readings agree with known-good values recorded previously
Antenna System	Rotatable antenna on-air pattern checks as expected
Antenna System	Full-power test, no intermittent behavior or faults
Antenna System	All connections secure, waterproof, free of corrosion
Computer	Software configured and updated properly
Computer	Data interfaces functional
Computer	Audio output levels correct, free of noise and hum/buzz
Computer	Audio input levels correct, free of noise and hum/buzz

Don't forget that your test equipment needs to be checked, as well. If you have a directional RF power meter or antenna analyzer, check it against a friend's to be sure yours is giving reasonable readings. If the equipment is battery powered, proof of performance time is a good opportunity to change out the batteries for fresh ones.

Performing proof testing once a year or before an important operating event is not unreasonable. Perhaps this would be a good team exercise to make sure your EOC or Field Day station are ready to go. As a consequence, you will all become more familiar with the equipment and will no doubt notice something that needs fixing or doing. Make "proofing" your station a habit — you won't regret it the first time you discover a problem that would have taken you off the air or damaged something expensive!

1

www.radiomagonline.com/deep-dig/0005/proof-of-performance-manuals/31594

Experiment #151 — Quist Quizzes

In every issue this year, *QST* will be featuring an item from its first century. "Hands-On Radio" gets into the act this month by featuring *Quist Quizzes*. These tiny technical tidbits were sprinkled into various issues of *QST* for a decade between 1951 and 1961.[1] Each featured an interesting head-scratcher, usually technical, for readers to puzzle out. The topics are often still of interest today. Let's take a look at some of my personal favorites!

January 1959

Many of the Quist Quizzes involved circuitry puzzles, such as the one seen in Figure 1. There are three resistors in series, but you only know three things: R_1 has a value of 2 Ω, R_2 dissipates 2 W, and R_3 has a voltage drop of 2 V across it. Given that the power supply is a 10.5 V battery, how can you puzzle out the value of R_2 and R_3, as well as the current, I?

September 1956

The problems weren't all busywork math, either. Most were intended to address problems typical of the operating styles and equipment of the day. While most of that equipment is long gone, some of the same challenges are still around today. For example, when using a vacuum tube amplifier, you should turn on the filament transformer first and only after the tubes have warmed up, turn on the high-voltage transformer. You don't want to get that order reversed, so a foolproof circuit is needed to guarantee the high-voltage transformer is turned on last and off first — no matter in which order the two switches are thrown. This sounds complicated, but you can make it happen with two ordinary switches — no microprocessors were around when this quiz was devised. Think about this one for a while and see if something doesn't toggle an idea!

November 1956

Here's another problem that occurs today as we wire up our stations with multi-conductor cable. Sooner or later we splice together pieces of cable with different color codes (or no color coding) and then wonder, "Now which wire is which?" Figure 2 shows the general idea.

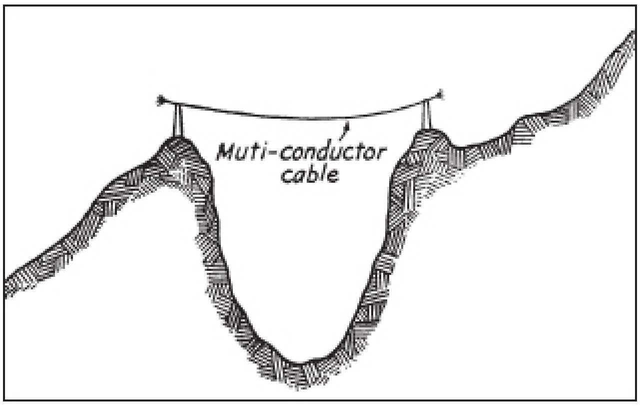

Figure 2 — Tracing out a multi-conductor cable often feels like it's strung across a mountain gorge, but you can do it in one round trip!

Let's assume you have an odd number of conductors. How about nine? Not wanting to make many round trips back and forth from end to end, especially if one end is at the top of the tower, how can you minimize the number of trips while only using a continuity tester? Surprisingly, the problem can be solved in a single round trip! And no foxes, geese, bags of corn, or river crossings are involved!

December 1958

Occasionally, the problem was a mechanical one instead of electrical, but just as interesting to solve. Updated for today's modes of transportation, here's a situation that occurs on a regular basis: A 6 meter enthusiast has a $1/4$-wavelength vertical antenna made of solid $3/4$-inch aluminum rod 1.7 yards long. She really wants to take this on a plane, but it won't fit in the overhead compartment, and excess freight charges for anything measuring more than 3 feet on any side make that an expensive proposition. Nevertheless, our resourceful VHF operator found a way to take the antenna without busting the budget. How did she do it?

March 1959

We all want to know the answer to this one — you have a roll of electrical tape that is 0.05 inches thick, wound on a 1-inch core and with a 6-inch outer diameter. How much tape do you have?

October 1961

Let's finish with one of the last of the lot. This was sent to *QST* by Harold Lanier, W4IFH, revising a Martin Gardner puzzle from *Scientific American* (another source of classic columns and conundrums): Radio operator A told operator B to look for him on a certain frequency. Operator B remembered the six numbers of the frequency, but he swapped the three digits representing kHz with those for MHz (ie MHZ.KHZ became KHZ.MHZ)! While B couldn't find A at that frequency, he tuned 5 kHz lower and there was the second harmonic of A! On what frequency was A operating? Hints: the frequency is an integer number of kHz and there are three digits in both the MHZ and KHZ values, none of which are 0. I'll let you twist in the wind a while on this one — the solution will be in next month's column!

Solutions

Puzzle 1: You know three things — I will be the same for all three resistors; $P = I^2R$ so $R_2 = 2/I^2$; and the sum of the voltage drops across all three resistors equals 10.5 V:

$$2I + I\left(\frac{2}{I^2}\right) + 2 = 10.5\ V$$

Solving for I gives:

$I^2 - 4.25I + 1 = 0$

Quick — head to the Internet and look up the quadratic equation! This gives not one, but two values for I: 4 A or 0.25 A. That means there are two sets of solutions: I = 4 A, R_2 = 0.125 Ω, and R_3 = 2 V / 4 A = 0.5 Ω or I = 0.25 A, R_2 = 32 Ω, and R_3 = 8 Ω.

Puzzle 2: The secret lies in using double-pole switches and wiring one pole of each switch in parallel, using that pole to control the filament transformer, as shown in Figure 3. This solution was devised by Willard Waite, W8DGQ, who noted that a switch in series with each side of the high-voltage transformer would be safer because a single-switch fault would not turn on the high voltage.

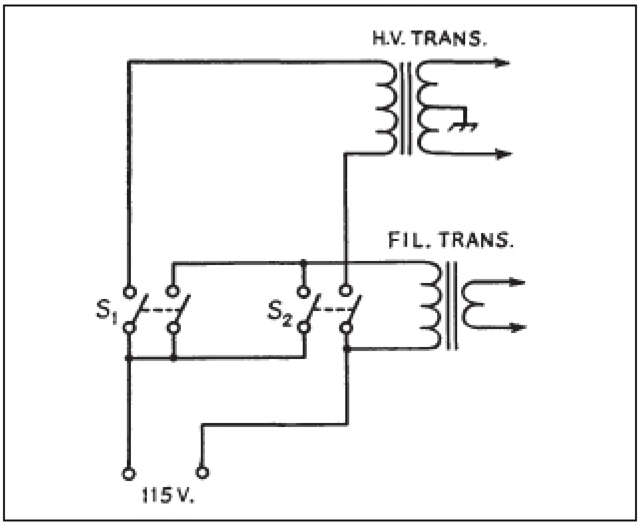

Figure 3 — Using a pair of double-pole switches ensures that filaments are powered up first and powered down last.

Puzzle 3: Figure 4 shows how to solve this problem in a sequence of mental steps without so many physical steps.

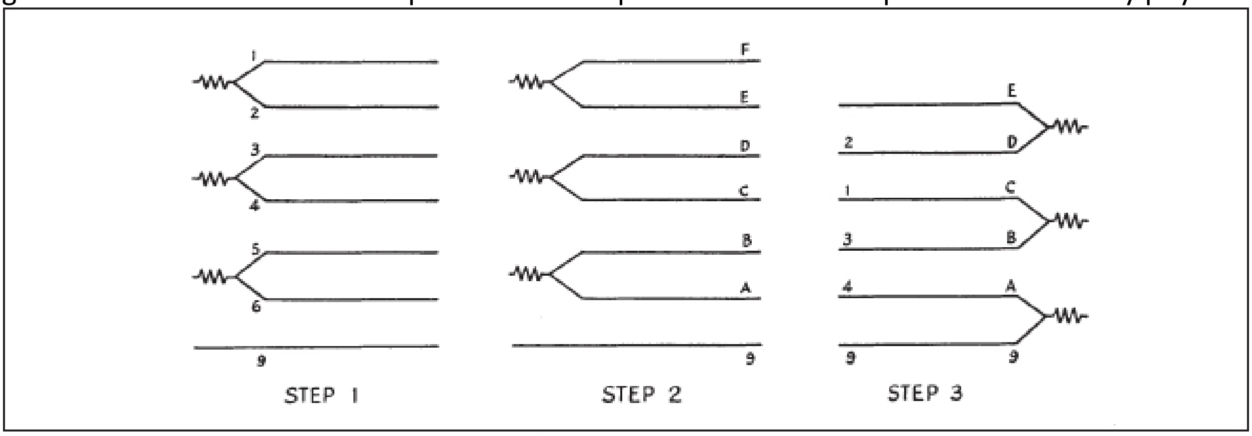

Figure 4 — By pairing wires and using the odd wire as a starting point, any number of wires can be identified with only one round trip between the ends.

Step 1 — At one end of the cable, number each wire consecutively, then connect 1 – 2, 3 – 4, 5 – 6, 7 – 8, leaving 9 open.

Step 2 — Head outside with your continuity tester. Find each of the four paired conductors and label them A&B, C&D, E&F, and G&H. The ninth wire is the one not connected to anything. Connect the ninth wire to A, B to C, D to E, F to G, and leave H open.

Step 3 — Back in the shack, get a pad of paper and disconnect all of the temporary pairs. You already have Wire 9 identified and you know it is connected to Wire A at the other end. Find the wire it's connected to. Let's say that it is Wire

4. You now know that Wire 4 in the shack and Wire A at the other end are the same — write that down! Wire B outside must be Wire 3 inside because that's how you assigned the pairs — write that down, too. Now find the wire that 3 is connected to, and let's say that's Wire 1. That must be Wire C outside, so now you've identified another pair, with Wire 2 corresponding to Wire D outside. Repeat this process and soon you'll have the entire cable "buzzed out!"

What if you have an even number of conductors, such as in common rotator control cable? Because you have no unpaired wire to start off the step-by-step process, you'll have to make an extra round trip to identify and mark off one wire by connecting all but one of the wires together on one end, then looking for the open wire on the other end.

Puzzle 4: Our inventive operator used a cubical cardboard box 3 feet on a side. The diagonal between opposite corners measures 1.73 yards — that's plenty of room for the vertical with a few inches to spare and no excess freight charges!

Puzzle 5: The secret is to think in terms of area and not length. Find the area of the 6-inch roll, subtract the area of the 1-inch core, and divide by the tape thickness. The final answer is 175π, or 549.8 inches. From your quizzical expression, I could tell you knew π would make an appearance!

Experiment #152— Improvisation

Part of what makes ham radio (and hams) fairly unique in the world of telecommunications is its (and, presumably, our) flexibility and adaptability. Power's out? We have batteries. Phone system crashed? We just tune in to the local repeater. Repeater got hit by lightning? We'll switch to a better antenna and go simplex.

Improvising is built into the amateur DNA, so to speak, right out of the gate in Part 97's Basis and Purpose (§97.1): service built on science and skill (with a good attitude). It's useful to practice being flexible and adaptable to test the old brain cells. I got my chance on Field Day, which could very easily be renamed the "All-Amateur Improv Weekend."

An Interesting Antenna

My challenge was self-inflicted, having volunteered at a club meeting to build an effective HF antenna on the spot during Field Day with no preparation whatsoever. Because the St Charles ARC was setting up next to the parking lot of a home improvement store, I would use only parts from the store. I also accepted the restriction to not use regular wires and insulators and stuff. It had to be different and interesting!

Walking across the parking lot, I was thinking about the type of antenna, not just materials. There were no trees on site, just lots of open space, so horizontal antennas were pretty much out of the question. It was going to have to be a ground-plane vertical. Once upon a time, a beer-can vertical would have been the obvious choice, but I don't think we'd have been invited back and besides, thin aluminum sheet doesn't solder well at all!

[1]

I figured impromptu skyhooks for the low bands were impractical due to size and budget. Planning for an antenna big enough to impress without being impractical, I settled on a 20 or 15 meter vertical between 11 and 17 feet long. If it's in that range, an impedance matcher would probably be able to handle it — that was my thinking.

Cages Are All the Rage

Entering the garden department, inspiration struck in the form of a customer with a cart full of galvanized tomato cages. Only a ham would immediately think, "Gee, those look a lot like biconical dipoles!" I guess I've been at this too long…

Three-foot by one-foot cages (see Figure 1) turned out to be cheap and easy, welded and galvanized just like Rohn towers, but a little lighter. With their large outside diameter (compared to an individual wire), I would get more electrical length out of them, as well. I'd need somewhere between three and six cages for the bands I was targeting.

Figure 1 — The raw materials for the improvised 20 meter "Tomatopole" vertical consisted of several tomato plant cages made of galvanized wire, held together with split bolts; a bamboo stick for support; PVC pipe to keep it all upright, and a pair of 16-foot steel tape measures for radials.

With eight cages in the cart (after Field Day, they were going in the garden!), I imagined my creation stretching into the sky. How was I was going to hold it off the ground? Again, aimless wandering in the garden department turned to be the key — I found a 5-foot bamboo stick, sturdy, and lightweight.

My attention turned next to attaching the cages together. Soldering, splicing, or wrapping was out, so I considered clamping. Lately, I've grown enamored of using *split-bolt* connectors to hold anything together from ground wires to clotheslines — cheap, weather-tolerant, sturdy, and like the tomato cages, reusable at home. It would be easy to use little clamps to hold the cages together at their large circle ends, and bigger ones to clamp all of the stick-in-the-ground wires together in a sort of bow-tie configuration.

The antenna design was rapidly "taking shape," so to speak. A few hose clamps later, another swing by the garden department for some "tomato twine," then over to the tool aisle for a pair of tape measure radials as inspired by a recent *QST* article.
[2]

It was showtime!

All went well — bolts bolted and clamps clamped. Soon, six cages were attached to each other and to the bamboo stick, ready for the coax pigtails. But my estimate of the final assembly's rigidity was, shall we say, a bit optimistic. "Large fat noodle" more or less describes the problem. I quickly discarded the notion of using a *lot* of guying twine in favor of adding a central strengthening element. Back to the store.

By happy chance, the plumbing department was right next to the entrance, with racks of hard-to-miss 10-foot lengths of spotless white pipe. Cheaper than a 12-foot fir closet pole, too! Another bag of hose clamps and I went back to work with the pipe slipping over the bamboo to secure the base.

This time the antenna held itself together long enough to reach vertical and have the guy twine attached to rebar stakes. Tape measures were extended to their full and majestic 16 feet of safety yellow. Success! Oops, the top section decided to turn into a weather vane and bent over 90 degrees. No problem, it's an inverted L! Plus, it was visually striking, satisfying that oh-so-important aesthetic requirement — visitors (and there were lots) loved it! I have christened it the "Tomatopole" and provided a full parts list in Table 1.

Table 1
Tomatopole Parts List

Qty	Part	Price each	Total
2	¾-1¾" hose clamp	0.85	1.70
5	½-1⅟₁₆" hose clamp	0.78	3.90
10 ft	1" PVC pipe		1.69
6	Tomato cage, 3' x 1'	1.29	7.74
1	5' bamboo stick		1.44
6	#8 split bolt	1.14	6.84
4	#6 split bolt	1.24	4.96
2	16' tape measures	3.96	7.92
1 ball	Tomato twine		3.44
		Total	$39.63

The coax was attached and SWR was swept. In between transmissions by the other stations, a simply astounding 1.3:1 was observed at 14.195 MHz! What a stroke of (undeserved) luck! And yes, with 100 W, it held a CQ frequency and made contacts all across the US, even attracting a caller from Europe.

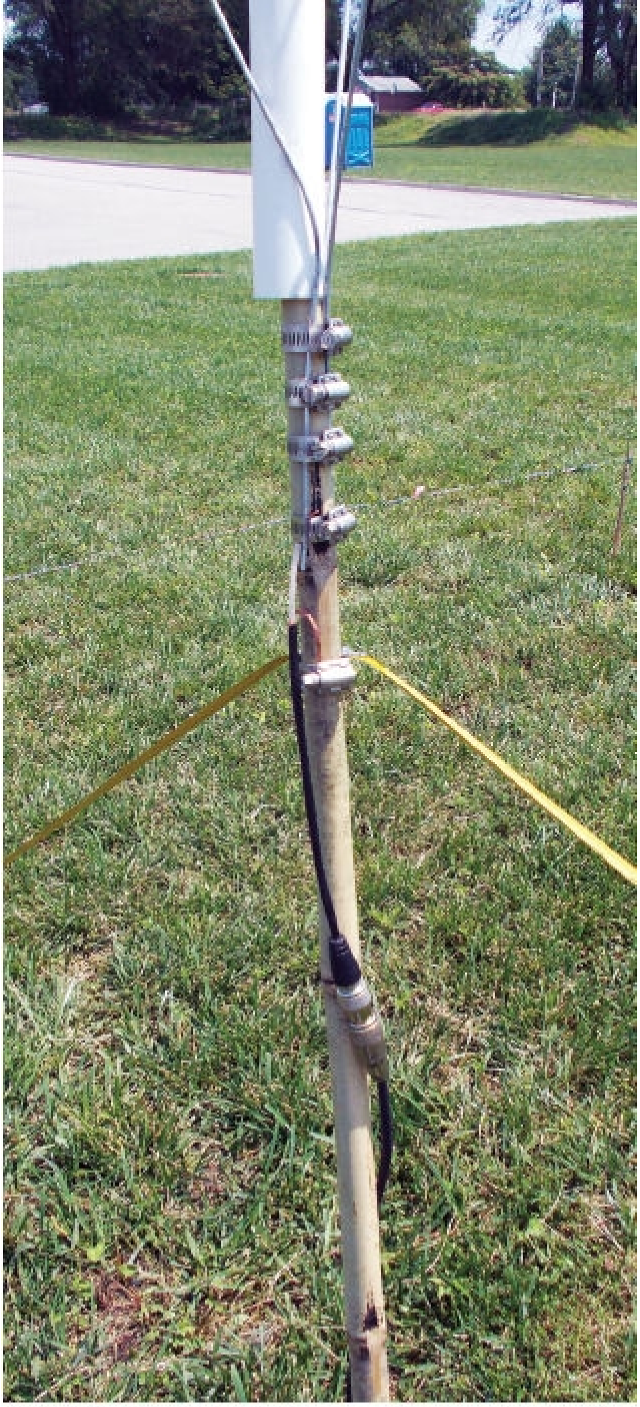

Figure 2 — Close-up of the feed point. The coax center conductor and shield are held in place with hose clamps.

Figure 3 — Not bad for a little rummaging in the hardware store's garden department! The vertical resonated at 14.195 MHz with an SWR of 1.3:1 and made plenty of contacts. The Fritos are optional.

Three-Tool Hams

This column isn't a construction template. Don't you *dare* make a copy of this antenna! Go out and make one up yourself — improvise! The ability to combine radio know-how (the science), organizational discipline (the service), and operating ability (the skill) is what sets Amateur Radio apart from other unlicensed services and makes us valuable as volunteers (and occupiers of prime wireless real estate). If our stuff breaks or isn't available, we can make new stuff because we understand how it works. Why not stress improvisation in your exercises and drills to make every member of your team a "three-tool ham?"

Puzzled Out

The answer to last month's frequency-finding fazer lies in the properties of numbers and a clue about carrying. [2]

First, break the fundamental frequency value into a pair of three-digit numbers: XXX and YYY (for MHz and kHz). That can be shortened to *XY* where *X* = MHz and *Y* = kHz. The three-digit format stays the same when XXX and YYY are swapped to become *YX*, so *X* and *Y* can have values from 0 to 999. We know the second harmonic's frequency is 2*X* (MHz) + 2*Y* (kHz). It is also *YX* – 5.

The only thing we don't know is whether multiplying *Y* by 2 to get the second harmonic causes a 1 MHz carry from the kHz to the MHz. If Y is 0 to 499 there will be no carry and $2X = Y$ (after swapping to reverse the MHz and kHz). If Y is 500 to 999, there will be a carry and $2X + 1 = Y$. How can we tell and solve the problem?

If there is no carry, $2X = Y$ and $2Y = X - 5$. If there is a carry, we have to account for the extra megahertz by changing the equations to $2X + 1 = Y$ and $2Y - 1000 = X - 5$. Solving the first pair of equations results in a value for X of –5/3, clearly not correct. The second pair, however, results in a value of X of 331, so Y must be $2X + 1 = 663$, and — *voila!* — the original fundamental frequency given for the schedule was 331.663 MHz!

Notes

1

Czerwinski, W. Pete, W2JTJ, "Budget 7-Mc. Vertical Antenna," Nov 1955, *QST*, pp 26 – 27.

2

Thibodeaux, Glen, KF5FNP, "A Tape Measure Vertical Antenna," Aug 2014, *QST*, p 33.

[3]M. Gardner, "Mathematical Games," *Scientific American*, May 1959, pp 164 – 174.

Experiment #153 — Learn by Fixing

The late, great Jim Williams, well known for his writing and design skills (he was a senior technical resource at Linear Technology, among other companies), was largely self-taught, like many "Hands-On Radio" readers. One method of learning that he promoted over and over again was to repair equipment. In the chapter "The Importance of Fixing" in *The Art and Science of Analog Circuit Design*, Jim wrote, "…fixing things is excellent exercise for doing design work. A sort of bicycle with training wheels that prevents you from getting into too much trouble." [1]

Jim's point is that the equipment worked at one time but then something changed. Your job is to find that something, which is easier than getting a design to work from scratch.

Along the way to fixing the equipment, you have to actually figure out how the circuit was supposed to work, maybe even drawing up a schematic if you don't have one. Unlike troubleshooting something you just built or designed, you can be pretty sure the design worked. To fix the circuit or equipment, you have to understand that design and look for ways in which the broken equipment deviates from how the design was *intended* to work. A late-night fix-it or build-it session is some of the best schooling you can get! Also from Jim, "…3:00 A.M., Tek. 547 [an oscilloscope], pizza, breadboard. That's Education."

The Fixing State of Mind

This won't be a step-by-step process because troubleshooting and repair is too complex and varied to follow a fixed procedure. If there is a common element to all fixing, though, it is to have an open mind, avoid assumptions, and remember that nature bats last. After you have a lot of experience, you will develop an intuition about certain types of problems, but when you are learning the ways and means of repair, resist the urge to jump around in the equipment thinking, "It must be…!" It usually isn't and you'll get frustrated. Slow down, keep notes, and ask "Why?" a lot.

The other key state of mind is to approach problems from a system standpoint: antenna system, power system, RF signal system, etc. Very few items in our shack are truly independent of other devices and wiring. Look at all the interconnected pieces and remember them as you contemplate what can be causing the problem. The system approach was introduced in Experiment #148, "Proof of Performance." [2]

We'll take a look at troubleshooting from the power perspective to reinforce the idea.

If Power Ain't Happy…

Then ain't nobody happy! Any time I start working on a repair, the very first thing I check (and often recheck) is that the power source *and all connections to it* are in good working order. Remember, think about the *system*, as in Figure 1. All of the elements in that system work together and create symptoms together.

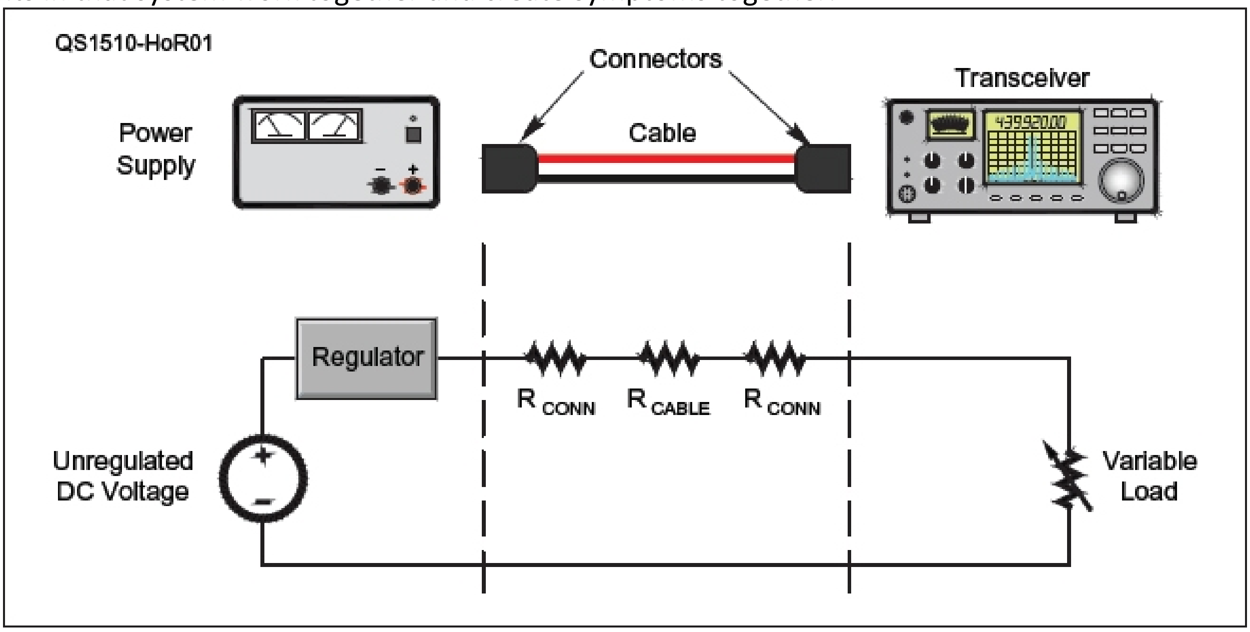

Figure 1 —Troubleshooting the combination of a power supply and a variable load like a transceiver should be approached as a system in which all of the parts work together.

A common linear power supply's output voltage can look fine at a light load (even if it has bad pass transistors) yet be wildly out of regulation at higher currents. Test the supply under load to be sure it's okay. It's always a good idea to check for unwanted ac on a power supply output, as well. An output voltage that looks "close" but wanders or varies might be due to an oscillating or "motorboating" regulator. Oscilloscopes may seem like overkill for testing a power supply but they show you things that meters can't.

If the power supply is okay, test the power *all the way* to the electronics that use it. That system includes the connecting cable, the panel connectors, and internal wiring. A common "gotcha" is using a different (good) power cable during testing and being unable to reproduce the problem. If possible, keep the system together when troubleshooting.

Note any changes and check those items later. For example, you might not be able to remove power wiring from a vehicle but be aware that you have changed the system and test the vehicle power wiring separately.

Most “12 volt” radios start misbehaving at a power supply voltage of 12 V and lower. The typical radio might draw peak currents of around 20 A. How much series resistance (R_{CONN} + R_{CABLE} + R_{CONN} in Figure 1) does it take to drop 13.8 V to 12 V?

R_{SERIES} = (13.8 – 12) / 20 = 0.09 Ω

That’s not a lot of resistance and a good reason to be sure connectors, particularly crimp-on terminals, are secure and clean. A radio may draw enough current through a bad connector to misbehave on SSB peaks, which can act like RF feedback or excessive drive, even though it performs just fine in receive and at low power levels.

It’s not uncommon for power connectors on radios and other equipment to get loose, dirty, or overheated. Remember that caution about assumptions! Trace the power all the way through the input connector into the internal circuitry under load.

[3]

DC Operating Point

Let’s say you are happy with the power supply and all of the connections. Everything looks good for power throughout your system — it’s time to open up the equipment and go to work. After you’ve familiarized yourself with the system inside the box, break the system down into modules or sub-assemblies and start trying to understand what they do. Further, you might also get to the point where you suspect a particular circuit is not working properly. Maybe it has no output or distorted output.

For an analog circuit, the next step is to verify the dc operating point. Take a look at Figure 2, the schematic of a simple common-emitter (CE) amplifier, subject of the very first “Hands-On Radio” column more than a dozen years ago. You might think that discrete component transistor circuits are out of vogue, but they are extremely inexpensive (the circuit shown probably costs less than 50 cents in production volumes) and are still common. Even without having a theory of operation, what can you tell just from taking some dc measurements?

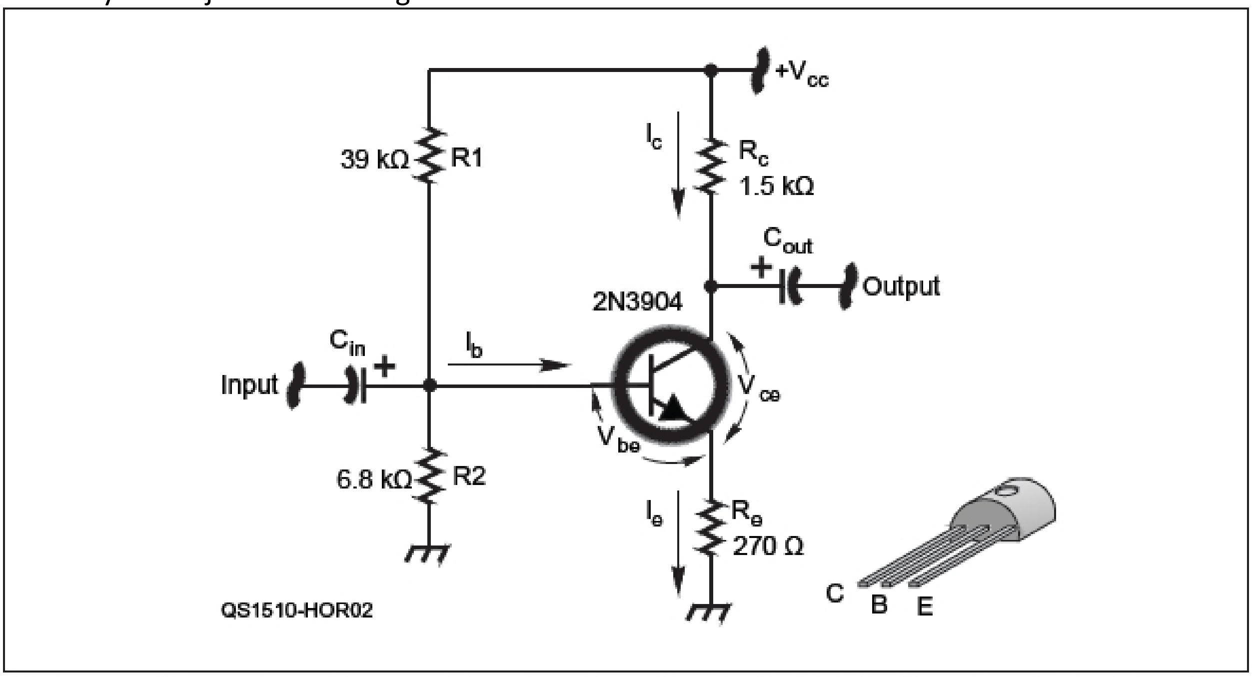

Figure 2 — Common-emitter amplifier circuit from Hands-On Radio’s Experiment #1. From the design inputs and equations, V_{CC} = 12 V, I_C = 4 mA, V_{CEQ} = 5 V, and voltage gain, A_V = –5.

The first thing to check in this circuit is V_{CC} to be sure it’s the “right” voltage. Assuming V_{CC} is good, take a voltage measurement at each pin of the transistor and calculate V_{BE} and V_{CE} or measure them directly pin-to-pin. (Review the original experiment if you’re not familiar with the CE circuit.) In this amplifier circuit, V_{BE} should be around 0.7 V or the transistor is not being biased correctly for some reason. V_{CE} should be somewhere between 0.1 – 0.2 V (saturation) and V_{CC} (cutoff). If those *Q-point* or *quiescent point* voltages aren’t right, that amplifier won’t operate properly.

Carry the same “simple tests” mentality to digital circuits. Are the power voltages right? Are all inputs and outputs at the right voltages for HIGH and LOW states? Are there pulse and data signals where they are supposed to be? Check the very basic stuff first, like the ENABLE and RESET pins. Slow and methodical, you’ll eventually find one clue and then another.

Practice, Practice, Practice

Try this with a friend: wire up a circuit on a breadboard and get it working. Then change something — a resistor value, power supply voltage, substitute a known bad component, even miswire or disconnect a component. Then ask your friend to find the problem with simple measurements. Start with something that works and then change one little thing. Once you’ve gotten the hang of it, move up to simple accessories and gadgets.

When trying to fix real-world circuits, you also have a powerful tool that will never touch the physical circuit but which can provide valuable insights — the circuit simulator. Recreate the circuit in a simulator like *LTSpice* (see Experiments #83 – 86) and have it compute a dc operating point for you at every point in the circuit. You can then look for significant discrepancies between the simulation voltages and what your meter tells you.

The SDT

No discussion of troubleshooting would be complete without mentioning the "SDT" (aka – Some [Darned] Thing). Sherlock Holmes is quoted as saying, "When you have eliminated the impossible, whatever remains, however improbable, must be the truth." Remember that nearly all of our equipment operates as part of a system, sometimes a very complex system. As such, it's quite common for an unexpected problem to cause vexing and inscrutable symptoms elsewhere, such as a rotary switch contact that has slipped out of alignment or a control signal set improperly or a loose connector shell or anything, really. The SDT often appears right after you think, "Gee, that's funny…"

One last quote from Jim: "Fast judgments, glitzy explanations, and specious arguments cannot be costumed as 'creative' activity or true understanding of the problem. After each ego-inspired lunge or jumped conclusion, you confront the uncompromising reality that the [darn] thing still doesn't work…When it's finally over, and the box works, and you know why, then the real work begins. You get to try and fix you."

References

1

J. Williams, *The Art and Science of Analog CIrcuit Design*, Butterworth-Heinemann, 1995.

2

All previous Hands-On Radio experiments are available to ARRL members at **www.arrl.org/hands-on-radio**
.

3

M. Zonnefeld, WØLTL, "Mobile HF Power RFI," Hints & Kinks, *QST*, Oct 2015, p 64.

Experiment #171 — AC Power Distribution

We all start off the same way — the table or desk near a power outlet with a computer and the rig's power supply plugged in. No problem! Then another radio and a wall wart or two make their appearance, so we daisy-chain one power strip into another, and so it goes. This leads to trouble, so let's head it off before it starts, beginning with…

Cheap Power Strips

Inexpensive plastic power strips are sold everywhere. They might do for a lamp or radio, but they are not appropriate for use in your station. A cheap strip has cheap sockets, wiring that's not heavy enough, and a switch that's not rated for load interruption (see the sidebar, "MOV Surge Protectors — Avoid Them," too). A plastic housing isn't going to do you much good in avoiding shock hazards, either.

Poor wiring and socket quality means that your all-important safety ground connections are poor, too. The leakage current from all of those power supplies and appliances creates a small voltage as it flows through the ground conductor. That voltage can show up as a buzz in low-level audio circuits, such as voice keyers and digital mode signal connections to computers and radios. If there is a lightning strike, even just nearby, the flimsy connections can't equalize the voltage between the enclosures, possibly allowing equipment damage.

What makes a good power strip, then? For starters, you need to be able to open up the strip and take a look. Make sure the wiring (including the power cord) is at least #14 AWG for a 15 A-rated strip. (20 A strips should have #12 AWG wiring and cords.) The connections should be short and direct. Outlets should grip the plugs securely and be firmly mounted. The strip's enclosure should be metal and connected to the power cord's green wire. (Make sure there is no paint between the green wire and the enclosure.) What you are looking for is often listed as "industrial" quality, and I have to warn you, these aren't cheap! Instead of $3.99, expect to spend at least $30, and more for the longer strips.

Making Your Own Power Strip

I have a better idea. Save money and make a better one! Figure 1 shows a 10-outlet box with all the parts available from the local hardware store. Here's how to build it, as explained to me in an e-mail from Jim Brown, K9YC:

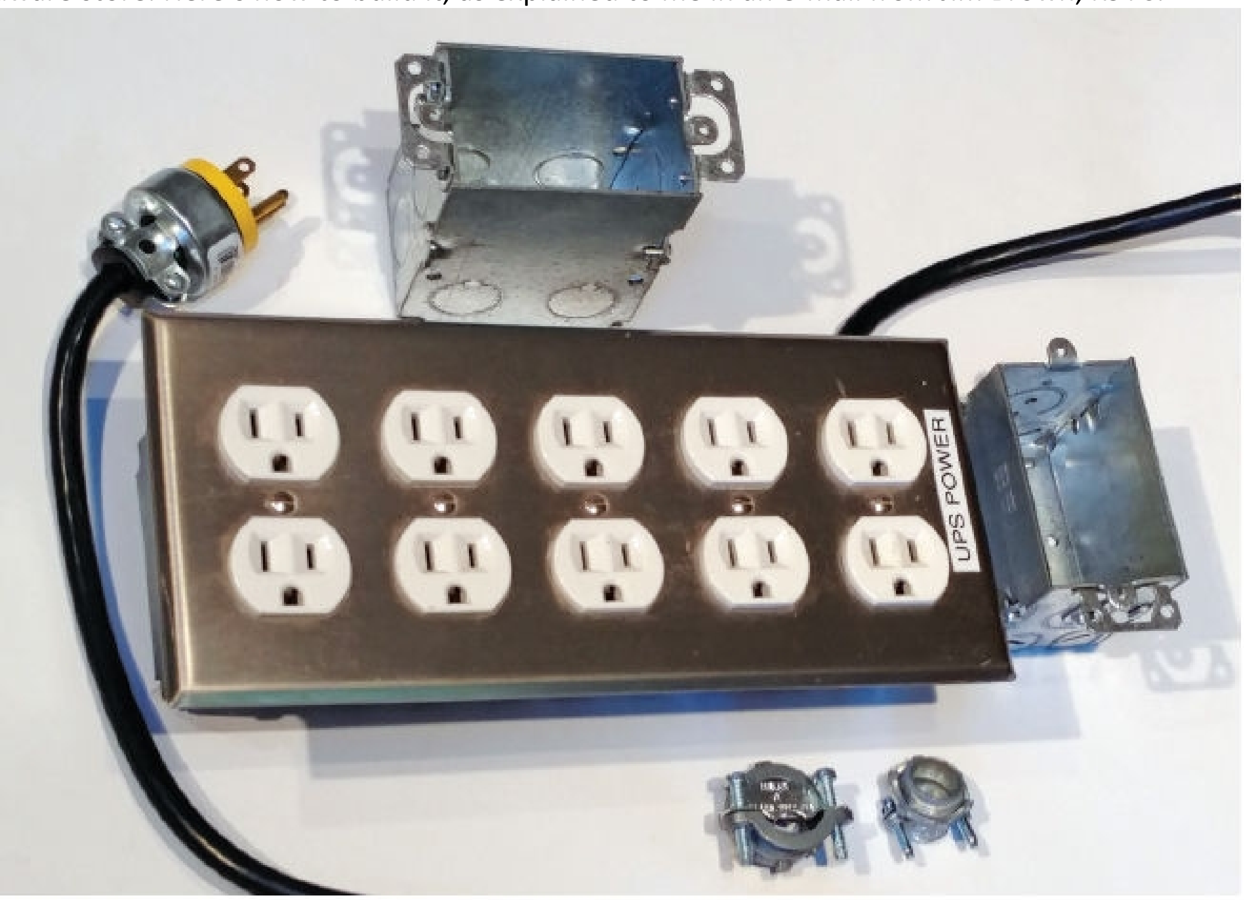

Figure 1 — This multiple-outlet power strip can be constructed from commonly available electrical supplies. A light switch or GFCI breaker can be added, as well. [Photo courtesy of Jim Brown, K9YC]

The [outlet] boxes are [gangable] "back boxes" for switches and outlets with removable side plates and mounting ears. Some also come with interior strain-relief brackets. The boxes come in at least two depths. This box was built from the shallower boxes, shown at right. The deeper box, shown above the completed strip, provides more space for things like chokes or filters.

To build the strip, remove each box's mounting ears, remove both side plates from interior boxes, and only one side plate from the end boxes. Use screws that held side plates to gang the boxes together [connect them side-by-side]. [Boxes are available with four or more outlet positions, as well — *Ed.*]

Before mounting outlets in the box, wire them together, feed the power cable into the box using one of the strain

reliefs at one of the "knock-outs" and wire it to one of the duplex outlets. Tighten the strain relief, then mount the outlets, centering them as carefully as possible. It may be necessary to loosen one or more mounting screws to fit the cover plate.
[1]

MOV Surge Protectors — Avoid Them

Many inexpensive power strips offer surge protection from internal MOVs (metal oxide varistors). Avoid them! As each transient is clamped, the dissipated energy slowly lowers their resistance. Eventually, they start generating some heat (look for an overheated spot near the switch on a power strip) and can start a fire. Check your power strips — if there is a MOV inside, clip its leads and remove it. Use a dedicated surge protector instead.

Although a regular non-metallic (Romex) cable clamp will secure a heavy flexible cord, use of a strain-relief fitting is recommended. Jim also added a label showing what type of power (UPS) it is connected to.

Along with the outlets, there's no reason not to include a light switch in one position to turn all of the outlets on and off. Or maybe make one outlet a ground-fault circuit interrupter (GFCI) model and use it to protect the whole string. (Note that vintage radios often have high enough leakage current to the chassis that a few of them on the same circuit could cause a GFCI to trip.)

When you have this beauty completed, you won't have to be worried about whether your power strip is up to the job. And it will cost a fraction of a top-quality commercial strip. (For more about ac wiring practices, see *The ARRL Handbook*'s chapter on Safety.
[2]
)

GFCI in a Box

Along with Jim's extra-heavy-duty power strip, another good idea showed up as I was browsing the internet — a GFCI in a box. Shown in Figure 2, this is a simple thing to make. Buy a metal back box, a GFCI-protected duplex output with cover plate, and a heavy-duty power cord. Wire it up, and you have a very nice addition to your portable station, generator kit, or RV. It's particularly useful if you're going to be using ac outside or anywhere that might be (or get) wet.

Figure 2 — This standalone GFCI-protected power outlet is a good accessory for portable or outdoor stations, including use at Field Day.

Field Day AC Safety

I've been to a lot of Field Days in my 40+ years of ham radio, and when I look back on how we distributed power years ago, it's a wonder we survived the first yank on the generator's starting rope! Luckily, the generators available to us today at rock-bottom prices per watt are much safer than in the past. We need to do *our* part to make sure we use them safely, too.

If you're going to use a generator, set it up properly and make sure its voltage is within spec at a reasonable load. (There is an overvoltage protector circuit in *The ARRL Handbook*'s "Power Sources" chapter.) Because the instructions for many generators leave a lot to be desired regarding grounding, download the two-page OSHA "Grounding Requirements for Portable Generators."[3]

It echoes the National Electrical Code's requirements. You may be surprised to learn that a ground rod isn't required, but using one that is connected to the generator frame and the ground contact of power outlets is not a bad idea.

Generators are often placed a good distance from the Field Day stations. It's common for one generator to supply more than one station, which themselves are separated for noise and RF isolation. In such cases, use a ground rod at each station *and* at the generator. (Hint — ground rods make a good place to anchor an extension cord in case someone trips over it.) Make sure the ground conductor of each extension cord is intact and wired correctly. Connect the metal enclosure of each piece of equipment together and then to the ground conductor of the power system. Maybe this would be a good place for the GFCI in a box? CQ Field Day!

Notes

1 Based on material from **k9yc.com/GroundingAndAudio.pdf**.

2 Available from your ARRL dealer, or from the ARRL Store, ARRL Item no. 0628. Telephone toll-free in the US 888-277-5289, or 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

3 Grounding Requirements for Portable Generators: **www.osha.gov/OshDoc/data_Hurricane_Facts/grounding_port_generator.pdf**

Parts List

HANDS-ON RADIO PARTS LIST (Thanks, John AF4WM and Steve AD7KR)
Updated through Experiment 179
Quantities assume all parts available for re-use
Optional parts not included
Parts not included for experiments 130 due to custom item modifications

1/4 WATT RESISTOR (All values are ohms)	QTY	EXPERIMENT	NOTES
1	1	129	Added in third set of experiments (122-181)
10	2	2,13,15,16,46	
27, non-inductive	1	21	
33	1	143	Added in third set of experiments (122-181)
39	2	13	
47	1	10	
49	2	129	Added in third set of experiments (122-181)
51	2	13,26,71,129,162	Added 1 ea in third set of experiments (122-181)
68	2	45,147	
75	2	13	
100	several	2,13,15,16,24,29,37,46, 48,74,77,99,100,103,161	Special batch requirements for experiment 100
150, non-inductive	1	21	
220	1	48,77	
220, non-inductive, such as carbon comp	2	81	Added in second set of experiments (62-121)
270	1	1,28,45,71,76,155	
300	1	79	Added in second set of experiments (62-121)
330	1	6,28,29,38,147	
390	1	48	
470	7	8,28,29,37,69,77,141	
510	1	9,55	
910	1	71	Added in second set of experiments (62-121)
1k	10	2,3,8,12,28,29,31,33,36,37,39, 40,55,73,77,80,101,109,111 121,125,149,155,161	
1.2k	1	2	
1.5k	1	1, 28,77,93,155	
2.2k	2	4,8,11,14,77,78	Added 1 ea in second set of experiments (62-121)
2.5k	1	129,161	Added in third set of experiments (122-181)
2.7k	1	27	
3.3k	3	93,141,147	Added in second set of experiments (62-121), added 1 ea in third set of experiments (122-181)
3.9k	1	3,6	
4.7k	3	5,8,9,45,54,68,77,93,97, 109,158	Added 1 ea in third set of experiments (122-181)
5.1k	1	6	
6.8k	1	1,28,55,77,147,155	
10k	4	2,3,4,7,17,18,30,31,40,41,46, 54,69,73,99,109,125,141,155	
11k	1	109	Added in second set of experiments (62-121)
15k	1	2	
16k	4	42	
22k	1	2,4,11,41,45	
27k	1	11,99	
39k	1	1,5,28,41,77,93,155	
47k	1	4,31,103,147	
56k	1	41	
62k	1	5	
75k	1	41	
91k	1	5	
100k	2	99,125	Added in second set of experiments (62-121)
120k	1	39	
220k	1	11	
270k	1	38	
330k	1	97	Added in second set of experiments (62-121)
470k	1	31,143	
680k	1	38	
1M	several	43,55,99,125,147,161	
4.7M	1	97	Added in second set of experiments (62-121)

POWER RESISTORS (All values are ohms)

1, 5W	1	123	Added in third set of experiments (122-181)
5, 5W	1	24	Resistors with larger power dissipations can be used
6, 50W	1	158	Added in third set of experiments (122-181)
15, 10W	2	158	Added in third set of experiments (122-181)
50, 10W	2	9(2)	Also OK to combine smaller units in series or parallel
100, 1/2 W	4	129	Components for experiment 129 not included in parts kit
100, 1W	1	10,32	
2.2 to 10, depends on installation preferences	2	129	Components for experiment 129 not included in parts kit

POTENTIOMETER (All values are ohms)

100	1	28,42	
50 to 100, depends on installation preferences	1	129	Components for experiment 129 not included in parts kit
500	1	158	Added in third set of experiments (122-181)
1k	1	3,4,14,155	
5k, panel-mount	1	78	Added in second set of experiments (62-121)
5k, PCB mount, Bourns series 3386F	1	109,141	Added in second set of experiments (62-121)
10k	2	3,8,11,12,14,54,55,68,69, 80,134	Added 1 ea in second set of experiments (62-121)
10k, PCB mount, Bourns series 3386F	1	141	Added in third set of experiments (122-181)
20k	1	9	
50k	1	10,103,174	
100k	1	1,2,38,41,92	
1M	1	17,39,40,109	

CAPACITORS

50 to 200 pF, fixed-value or air-variable	1	21,43	
500 pF variable	1	102	Maximum value not critical
2.7 pF, ceramic	1	43	
68 pF, silver mica	2	71	Added in second set of experiments (62-121)
100 pF, polystyrene or ceramic	2	45,46,97	Added 1 ea in second set of experiments (62-121)
200 pF, polystyrene or silver mica	1	43,46	
220 pF, silver mica	1	71	Added in second set of experiments (62-121)
270 pF, film or ceramic	1	97	Added in second set of experiments (62-121)
390 pF, polystyrene or silver mica	2	46	
470 pF, silver mica	1	79	Added in second set of experiments (62-121)
470 pF, film or ceramic	1	97	Added in second set of experiments (62-121)
560 pF, silver mica	2	156	Added in third set of experiments (122-181)
680 pF, silver mica	1	156	Added in third set of experiments (122-181)
750 pF, silver mica	1	156	Added in third set of experiments (122-181)
1 nF, 50V ceramic	3	17,37,74,93,103,121,147	
1.2 nF (1200 pF), silver mica	1	79	Added in second set of experiments (62-121)
1.5 nF, film or ceramic	1	18	
2.2 nF (2200 pF) silver mica	1	156	Added in third set of experiments (122-181)
2.7 nF, film or ceramic	1	4	
5.6 nF, film or ceramic	1	4	
10 nF, film or ceramic	4	5,25,42,45,46,71,93,109,121, 125,155,161	
16 nF, film or ceramic	2	41	
22 nF, film or ceramic	1	68	Added in second set of experiments (62-121)
33 nF, film or ceramic	2	4	
56 nF, film or ceramic	1	4	
0.1 uF, 50 V ceramic	6	8,12,17,24,26,27,38,39,40,43, 68,69,71,99,121	Added 2 ea in second set of experiments (62-121)
1 uF, 25 V electrolytic	4	7,40,69,93,121,125	
1 uF, 35 V tantalum	1	6,8	
10 uF, 25 V electrolytic	3	1,2,3,4,5,7,17,28,31,68,77, 147	
33 uF, 25 V electrolytic	1	109	Added in second set of experiments (62-121)
47 uF, 15 V tantalum	1	15,16	Changed from 100 uF in original article
470 uF, 35 V electrolytic	1	143	Added in third set of experiments (122-181)
100 uF, 35 V electrolytic	2	10,15,16,30,31	Exp 15/16 input capacitor changed to electrolytic
2200 uF, 35 V electrolytic	1	143	Added in third set of experiments (122-181)
4700 uF, 15V electrolytic	2	10,38	

INDUCTORS & CORES

2 uH	1	43	Handwound, 8 turns, 1-1/2" dia, 1" long
8.2 uH, air-wound (transmit-capable) or molded (receive-only)	1	156	Added in third set of experiments (122-181)
15 uH, air-wound (transmit-capable) or molded (receive-only)	1	156	Added in third set of experiments (122-181)
22 uH	1	46	Should be rated for 50 mA, large chokes OK
100 uH, 1 amp	1	15,16,118	Added in second set of experiments (62-121)
1 mH	1	26,74	
FT37-43 (ferrite) or equivalent	1	45	
FT240-61 (ferrite), FT240 balun kit from Amidon Associates is equiv source	1	48,129	
T50-6 (powdered iron) or equivalent	1	46	Original shopping list had FT50-6
T30-10 (powdered iron) or equivalent	2	71	Added in second set of experiments (62-121)
Ferrite snap-on core, or type #31 or #43 mix	5	103,136,163	Added 4 ea in third set of experiments (122-181)
T-157-2 (powdered iron) or equivalent	1	103	Added in second set of experiments (62-121)

TRANSFORMERS

115 V Pri, 12.6 V Sec,	1	10,33,143	
1000 ohm - 8 ohm audio transformer,	1	33	

TRANSISTORS

2N3904	2	7,134,141,147	
2N3906	1	19,25,40,109,155	
2N4401	1	8	
2N4416	1	43	
2N7000	1	39,40	
2N7002	1	161	Added in third set of experiments (122-181)
J310 or MPF102 JFET	2	99	Added in second set of experiments (62-121)
MJE350	1	155	Added in third set of experiments (122-181)
TIP31	1	9,80	
TIP42	1	38	
IRF510 MOSFET	1	9,12,15,16,161	
SCR, 100 V, 8 A - Digi-Key MCR218-004OS-ND	1	10	Original RadioShack part number no longer available
BS170 MOSFET	1	143	Added in third set of experiments (122-181)

DIODES & RECTIFIERS

1N34	1	76,102,103,134	Added in second set of experiments (62-121)
1N4001	2	71,76,77,109,121,143	Added in second set of experiments (62-121), Added 1 each in third set of experiments (122-181)
1N4007	1	155	Added in third set of experiments (122-181)
1N4148	12	6,7,14,25,26,30,39,40,42,54, 55,76,97,125,134,161	Added 8 ea in second set of experiments (62-121)
1N4732A	1	76	Added in second set of experiments (62-121)
1N4733A	1	6,38	
1N5234B	1	109	Added in second set of experiments (62-121)
1N5817	1	76	Added in second set of experiments (62-121)
1N5819	1	15,16	

INTEGRATED CIRCUITS

78L-05, -08, -12, or -15 voltage regulator	1	25	
4N35 Optocoupler	1	14	
741 op amp	2	3,4,8,17,19,42,54,55,73,134, 155	
555 timer	2	5,69,93	
565 Phase-Locked Loop	1	68	Added in second set of experiments (62-121)
LM317 op amp	1	8,71	
LM311 Comparator	1	11,38	
LM324 quad op amp	1	17	
LM393 dual comparator	1	109,141	Added in second set of experiments (62-121)
CD4013 dual-D-type flip-flop	1	101	Added in second set of experiments (62-121)
CD4027 dual JK-type flip-flop	1	97	Added in second set of experiments (62-121)
CD4028 BCD-to-Decimal Decoder	1	37	
CD4029 Up/Down Counter	1	36	
CD4511 BCD-to-7-Segment Decoder	1	37	
TL082 dual op-amp	1	141	Added in third set of experiments (122-181)
74HC4001 Quad NOR gate	1	97	Added in second set of experiments (62-121)
74HC4040 12-stage ripple counter	1	97	Added in second set of experiments (62-121)
74HC14 hex inverter Schmitt trigger	1	125	Added in third set of experiments (122-)
74HC86 quad XOR Schmitt trigger	1	125	Added in third set of experiments (122-)

LIGHTS & DISPLAYS

Bulb, 12 V	1	10	
LED, Red	5	11,36,39,40,93,101,109,141, 158	
LED, Green	1	39	
7-Seg Display, Common-Cathode, Jameco 17187 or equiv.	1	37	

MISCELLANEOUS

40-meter crystal in HC-8 or HC-16 holder	1	46	7030-7045 kHz recommended
#12 AWG solid, bare wire, use scrap house wiring and remove insulation	6 feet	96,136,175	Added in second set of experiments (62-121)
#14 AWG solid, enameled wire (Incl. in FT240 balun kit from Amidon Associates)	14 feet	48	
#16 stranded, zip cord	3 feet	131	Added in third set of experiments (122-181)
#20 or #22 AWG solid wire	16 feet	21,71,79,102	Total wire length for #46 & #46 & #71(6' beyond anthology list)
#24 to #28 AWG solid wire	3 feet	45,46,47	Total wire length for #46 & #46, use scrap for #47
PL-259 connector	2	116	Added in second set of experiments (62-121)
SO-239 connector	1	21,48,131	
UHF double-male adapter	1	21	
UHF tee connector	1	103,116	Added in second set of experiments (62-121)
Plastic film can or pill bottle	1	22	
SPST or SPDT switch	1	22, 129	Components for experiment 129 not included in parts kit
SPST momentary NO pushbutton, Omron 6mm for the PCB layout	1	109,125	Added in second set of experiments (62-121)
RG-58 coaxial cable	25 feet	22,111,116	
RG-59 coaxial cable	20 feet	81	Added in second set of experiments (62-121)
Wooden spring-loaded clothespins		23	
Scraps of 2x4 wood		23	
Thick cardboard		23	
1/4-inch diameter dowel		23	
Glue		23	
Wood screws		23	
#4 self-tapping pan-head screw	4	131	Added in third set of experiments (122-181)
#6-32 machine screws		23	
#6-32 nuts		23	
#6 flat washers		23	
DPST switch	2	30	
TO-220 heat sink	1	24	
Mounting hardware for TO-220 heatsink	1 set	24	
1.5 V AAA or 9 V battery	1	32,155	
Terminal strip, 3 terminals plus ground	1	71	Added in second set of experiments (62-121)
Barrier strip, 6 terminals	1	158	Added in third set of experiments (122-181)
Crimp-on ring terminals, #10 holes, 16-14 AWG wire	2	116	Added in second set of experiments (62-121)
PC board scrap	1	70,71	Added in second set of experiments (62-121)
5A inline fuse and holder	1	71	Added in second set of experiments (62-121)
Mechanical quadrature encoder, Bourns PEC16 series recommended	1	101	Added in second set of experiments (62-121)
Microammeter, 50 to 250 uA max value	1	102,103	Added in second set of experiments (62-121)
SPST Relay, 12 V coil, Potter & Brumfield T90N1D12-12 for the PCB layout	1	109,143	Added in second set of experiments (62-121)
V8ZA05P 6V MOV	1	121	Added in second set of experiments (62-121)
1.5KE8.2CA 8.2V bipolar TVS	1	121	Added in second set of experiments (62-121)
Communications speaker, 4-32 ohms	1	129	Components for experiment 129 not included in parts kit
Plastic or glass jar or beaker (1 pint or 500 ml)	1	175	Added in third set of experiments (122-181)
Aluminum strip (a few inches long)	1	175	Added in third set of experiments (122-179)
Galvanized nail or bolt (3 or 4 inches long)	1	175	Added in third set of experiments (122-179)
Stainless steel strip or rod (a few inches long)	1	175	Added in third set of experiments (122-179)
1/8" stereo phone plug	1	64	
1/8" mono phone plug-to-stereo phone jack adapter	1	64	
3' male-to-male stereo audio jumper cable	1	64	
1/8" stereo headphone splitter	1	64	