Letters to the Editor

Dual Directional Wattmeters (May/Jun 2006)

Hi Doug,

In my article, I attempted to provide an accurate technical explanation while keeping the mathematics to a minimum. In conversations with readers, I have found that the simplifications left a small degree of uncertainty about subtle details. In particular, it is not perfectly clear how the calculations properly handled phase information. But as I noted in that text, all of the equations are accurate if one solves them using phasors (sometimes referred to as vectors in this context) instead of scalar values.

The only major revelation in my analysis is the proof that these circuits actually measure true power in arbitrary complex loads, in a way that really does not rely on dual-mode wave motion on transmission lines. All the discussion about transmission-line theory is merely an aid to understanding but as I have shown, the circuits will correctly measure power without a transmission line involved. The following is a complete proof of this using proper ac phasor notation. I recognize now that I should have included this in the original article, but better late than never!

Before jumping into a mathematical jungle, let me review the issue quickly. Any passive device, including an antenna, has an impedance $R + jX$. When connected to a sine-wave generator of frequency ω, there will be a transfer of power from the generator to the load. It will generally be accompanied by a "circulating" power whose time average in the load is zero. The goal of power measurement is to measure the non-zero true power even if it is accompanied by any amount of the latter.

Using the notation from the article, particularly from Figures A1, A2 and A3, we assume the wattmeter to be inserted between generator and load, Z_L, resulting in current as shown, with a voltage drop, V. These will now be represented as phasors **V** and **I** [here in boldface — Ed.], which are applied to Equations D1 and D2 as usual, resulting in the phasor "output" voltages v_1 and v_2:

$$\mathbf{v_1} = k_1 \boldsymbol{V} + k_2 \boldsymbol{I} \quad \text{(Eq 1A)}$$

$$\mathbf{v_2} = k_1 \boldsymbol{V} - k_2 \boldsymbol{I} \quad \text{(Eq 1B)}$$

where $k_1 = C_2 / (C_1 + C_2)$ and $k_2 = R_S / 2N$.

Those equations are easily manipulated to a useful form by taking the complex conjugate of both sides of each equation and noting that $XX^* = |X|^2$

$$|\boldsymbol{v_1}|^2 = k_1^2 \boldsymbol{VV^*} + k_2^2 \boldsymbol{II^*} + k_1 k_2(\boldsymbol{VI^*} + \boldsymbol{V^*I}) \quad \text{(Eq 2)}$$

$$|\boldsymbol{v_2}|^2 = k_1^2 \boldsymbol{VV^*} + k_2^2 \boldsymbol{II^*} - k_1 k_2(\boldsymbol{VI^*} + \boldsymbol{V^*I}) \quad \text{(Eq 3)}$$

Note that the voltages $\mathbf{v_1}$ and $\mathbf{v_2}$ are the actual measured voltages, which are then displayed in units of volts-squared on the meters, multiplied by a constant, with units of W/V^2. Such measurements are scalar and necessarily measure the *magnitude* of the voltages. Subtracting these two equations and replacing the amplitudes $\boldsymbol{V}$ and $\boldsymbol{I}$ by their RMS values, we get:

$$\boldsymbol{v_1}^2 - \boldsymbol{v_2}^2 = 2\, k_1 k_2\, (\boldsymbol{VI^*} + \boldsymbol{V^*I}) \quad \text{(Eq 4A)}$$

$$\boldsymbol{VI^*} + \boldsymbol{V^*I} = |\boldsymbol{V}|^* |\boldsymbol{I}| (e^{j\theta} + e^{-j\theta}) = 2VI \cos\theta \quad \text{(Eq 4B)}$$

$$\boldsymbol{v_1}^2 - \boldsymbol{v_2}^2 = 4\, k_1 k_2\, VI \cos\theta \quad \text{(Eq 4C)}$$

where V and I are now RMS values, and θ is the phase angle between V and I.

That result now shows that the difference of the two meter readings is indeed proportional to the true load power, as expressed by its classical formula in terms of the phase angle θ. The final calibration formula is established by re-writing this using $P = VI \cos\theta$ and a meter-scale calibration factor C (in W/V^2):

$$P = C\,(v_1^2 - v_2^2) / 4\, k_1 k_2 \quad \text{(Eq 5)}$$

All of the circuit constants are absorbed into the constants k_1 and k_2, leaving a true-power value that is independent of the load. Hence, it is proven to be a true wattmeter.

The relevance of Z_0, the calibration impedance, is obviously secondary. It is buried in the circuit constants and its only relevance is that it can be selected to produce a null reading on the reflected power meter for one specific value of load resistance, usually chosen as 50 Ω. However, as the equations show, if this value were changed from 50, the resulting change in k_2 could be compensated by a re-adjustment of either k_1 or the meter calibration resistors (which change the constant C) to maintain overall power calibration.

Final conclusion: DDW operation using the circuits discussed does *not* depend upon transmission-line wave mechanics. These devices function perfectly well for arbitrary two-terminal loads. If there is a transmission line involved, however, those wave mechanics will force the V and I relationships on the line to assume values consistent with the load power and phase. That results in the apparent presence of the standard forward and reverse wave modes, and if the line impedance is Z_0, then the meters will correctly read the amount of power in each mode. Otherwise, the individual meter readings are of little use. *— 73, Eric von Valtier, K8LV;* **evonvaltie@aol.com**

Uniform Current Loop Radiators (May/Jun 2006)

Dear Doug,

Bob, NP4B, has initiated a new era of uniform current loops! I have followed that discussion with a growing interest, because this kind of antenna would solve my problems in monitoring NCDXF beacons.

After the comments about the topic in *QEX*, and kind explanations of the theory from Bob, I decided to simulate a single wire loop with lumped capacitors at the corners. This concept, also using four short supporting spiders, would create a sturdy mechanical construction with minimal wind load. Raising the height to a half wavelength puts the maximum elevation gain to more or less the optimal ionospheric propagation angle and at the same time reduces the man-made ground-wave noise to the minimum.

To my positive astonishment, the simulation result was better than I ever could have imagined. Perhaps Bob, with a better theoretical backgound, may add some further comment. In honor of him, Robert Zimmerman, and of course not forgetting myself, OH2RZ, I named the antenna the 2RZ-LOOP.

I am very thankful to Bob for his great idea and help that inspired me to get a better monitoring antenna than the usual quarter-wave vertical. Now I'm waiting to find time to put the antenna up before the harsh Finland winter sets in.
— 73, Ahti Aintila, OH2RZ; **ahti@attocon.com**

CW Shaping in Software (May/Jun 2006)

Hi Doug,

I am giving some thought to a question raised by a good friend of mine. His proposition is that a high standard of close-in phase noise in a receiver is more hype than of practical significance. This train of thought returned me to the age-old question of how much bandwidth is needed for CW.

I have read many papers on this topic including yours [**www.doug-smith.net/cwbandwidth1.htm**] and the article in *QEX*. It is clear that only 100 Hz or so are needed and hence I want to claim that close-in phase noise is indeed a primary requirement — especially true for CW. I have not seen any discussion of close-in phase noise objectives and would appreciate your thoughts on the topic.
— 73, Ron Skelton; **ron-skelton@charter.net**

Hi Ron,

Well, I think we have two things to consider: 1) how much interference you create because of your TX phase noise, and 2) how much QRM you suffer because of your RX phase noise.

1. Although you didn't mention TX, I do because in a transceiver, there's generally a close correlation between TX and RX phase noise levels. Obviously, the more PM and AM noise you transmit, the more you could QRM someone. And, at some relatively low frequency offset from your carrier, phase-noise power density will surpass keying sideband power density — but that's at a very low level — when the keying is done correctly.

2. While you may have the lowest phase-noise RX in the universe, the QRM you get from others may depend on the phase noise they're transmitting. In other words, there's little point in having RX phase noise that's more than about 6 dB better than that of the transmitter of the guy whose signal is giving you grief. Make sense? So if you've got the lowest RX phase noise possible, then you at least know it's not your fault. — *73, Doug Smith, KF6DX,* QEX *Editor;* **kf6dx@arrl.org**

Doug,

Thanks for the response. Does the same argument apply to TX/RX IMD performance? — *73, Ron Skelton;* **ron-skelton@charter.net**

Hi Ron,

Good! Yes, I think it does. The low-order IMD performance of some receivers isn't as good as that of many transmitters. I'd be willing to bet that today's solid-state transmitters dominate in high-order IMD, so high-order receiver IMD might not be an issue, but who knows? — *73, Doug, KF6DX;* **kf6dx@arrl.org**

An Innovative 2-kW Linear Tube Amplifier (Jul/Aug 2006)

Dear Doug,

There are some minor errors that I did not detect when reviewing my article. Thanks to the readers who pointed out some of them by e-mail. The errors are:

1) On page 21, Figure 4 — Power supply schematic:

a) Fuse F2 is not connected to the R11/R12 node as shown but only to the R14/C24 node.

b) Pins 2 and 6 of CN5 are tied to pin 3 of CN7, not to ground.

c) The ground connections in this page and in Figure 5 at page 22 are not the same as the rest of the circuit. This ground is not isolated from mains ac supply and must be floating referenced to amplifier ground.

2) On page 23, Figure 7 — Control schematic: Contact A4 of switch SW1:1 must remain unconnected.

3) On page 24, the first paragraph incorrectly states: "…presents an open circuit to odd harmonics, a short circuit to even harmonics." The correct statement is just the opposite: short circuit to odd harmonics and open circuit to even harmonics.

It is also informative to tell that components Q1-5, Q7, Q11, Q17, D1, D3, D5, U12 and U13 are all mounted to the main heat sink or bolted to the aluminum frame.

I liked very much the final presentation and so did the friends to whom I sent the magazine! I had received several e-mails from readers who also liked the article — many thanks to them. — *Regards, Saulo Quaggio, PY2KO;* **saulo@auttran.com.br**

Effective Directivity for Shortwave Reception by DSP (July/August 2006)

Doug,

Carl Luetzelschwab, K9LA, e-mailed me about my comment on circular polarization: "I have found no information as to how this is for DX signals." Carl told me that Bob Brown, NM7M has written about this topic. He pointed me to Bob's article, "Power Coupling on 160 Meters," *Communications Quarterly*, Summer 1999, pp 95-101. I would like to forward this information because it is a very good and understandable article. — *Jan Simons, PAØSIM*; **pa0sim@amsat.org**

Octave **for System Modeling (Jul/Aug 2006)**

Dear Doug,

Chuck Hutton and Ben Bennett, N7IVM, were kind enough to point out a bug in my code that is included in my article. I've been editing the code using a text editor and then running the resulting file from the command line of either a DOS window or the Linux command line and it runs fine that way. When the code is executed from within *Octave*, though, it fails.

The reason for the failure was pointed out by Chuck and is also found on page 111 of the *GNU Octave Manual* by John W. Eaton. The manual points out that to differentiate between script and function files, a script file (which is what we are using here) must not begin with the keyword "function." I had thought that the presence of a line in the file that must be interpreted by the operating system to locate the executable (the first line in the file that begins with "#!") would meet that requirement, but such is not the case.

A fix recommended by Chuck and also covered on pages 111 to 112 of the *GNU Octave Manual* is to insert a line that *Octave* will execute but which will produce no effect on the cal-culations or output of the program. One such line is "1;" (a numeral one followed by a semicolon and an endline character). You can insert it immediately before the line beginning with "function" and it will take care of the problem. — *73, Maynard Wright, W6PAP;* **m-wright@eskimo.com**

Hi Maynard,

Thanks for sending the updated file. I have posted the corrected file as part of **7x06_Wright.zip** in the July/August section of the file listings at the **www.arrl.org/qexfiles** Web site. — *Larry Wolfgang, WR1B,* QEX *Managing Editor*; **wr1b@arrl.org**

Magnetic Coupling in Transmission Lines and Transformers (Sep/Oct 2006)

Doug,

In the article by Gerrit Barrere, KJ7KV, the equation on page 30 does not yield approximately 53 + *j* 0 Ω. It yields something quite different.

Further, the equation is the high-frequency approximation for the characteristic impedance of a transmission line. 60 Hz is not a high frequency for RG-58. Assuming 262 nH, 93 pF and 0.046 Ω/m, the characteristic impedance of RG-58 at 60 Hz would be about 810 – *j* 810 Ω. — *Bert Weller, WD8KBW*; **sodiumflame@sbcglobal.net**

Hello Bert,

Sorry, a "typo" sneaked into that equation. The text says 90 nF and 260 µH. It should say 93 pF and 261 nH, which is what the equation should be, too. With those values, Z_0 does come to 53 Ω at high frequencies.

You're right about the Z_0 equation I used not being valid for such low frequencies. I was trying to make the point that transmission line behavior reduces to circuit analysis for electrically short lines, but over-looked the fact that 60 Hz is well below the "break point" of RG-58, where it starts to behave like a fully coupled line. Ott has a good discussion of this (see the article's bibliography). I should have used something like 100 kHz and 50 m of RG-58 for the example. Then the point would have been made with the numbers also correct. — *Gerrit Barrere, KJ7KV*; **gerrit@exality.com**

Dear Bert,

Thanks. For the benefit of readers:

You're right about the 93 pF/m and 261 nH/m. From the ITT book (*Reference Data for Radio Engineers*, 6th Ed, Howard Sams & Co, 1975):

$$Z_0 = \left(\frac{R + j\omega L}{G + j\omega C} \right)^{\frac{1}{2}}$$

where Z_0 is characteristic impedance, L is inductance, C is capacitance, R is resistance and G is the conductance of the dielectric material (polyethylene), all per unit length; and ω is angular frequency ($2\pi f$). Assuming ωL is very small compared with R and that G is small compared with $j\omega C$ allows the simplification:

$$Z_0 = \left(\frac{R}{j\omega C} \right)^{\frac{1}{2}}$$

$$Z_0 = \left(\frac{-jR}{\omega C} \right)^{\frac{1}{2}}$$

Plugging in the values, $Z_0 \approx 810 - j810\ \Omega$ at 60 Hz. The characteristic impedance is high, capacitive and very frequency-sensitive. The Z_0 of RG-58/U doesn't approach $50 + j0\ \Omega$ until you get near 1 MHz or so.
— 73 de Doug, KF6DX

Dear Doug,

I really enjoyed the article by Gerrit Barrere, KJ7KV. He did a great job explaining the relationship between transmission lines and transformers.

I just disagree with one detail: the assertion that the total magnetic flux external to the wire increases greatly as the wire size is reduced. In fact, I believe that total external flux of an infinitely-long straight wire in free space is infinite and so does not depend on wire size.

It's easy to prove. As stated in the article, the external field is proportional to $1/r$, where r is the distance to the center of the wire. If you integrate $1/r$ from R (wire radius) to infinity you get $\log(\infty) - \log(R)$, which is infinity. The consequence is that the inductance per unit length of an infinitely long straight wire in free space is infinite!

When I first realized this many years ago, it made me horribly confused for a while. As a young engineer I had been told that a useful crude rule-of-thumb is that a wire has an inductance of 20 nH/inch (8 nH/cm). That seems like a contradiction. Of course, the explanation is that in real life you never have a wire infinitely far away from all other conductors. A real wire usually connects to something on each end and the circuit is completed, either through a ground plane or a returning wire. Because the inductance is proportional to the logarithm of the distance to the return path, it changes only very slowly as the spacing is varied.

As an example, a piece of bare 22 AWG hook-up wire (0.0253 inch diameter) spaced 0.32 inches (0.8 cm) above a ground plane actually does have an inductance of 20 nH/inch. However, if you move the wire twice as far away, the inductance only increases to 23.5 nH/inch. If you switch to wire that is twice as thick (16 AWG), the inductance at 0.32 inch spacing drops to 16.4 nH/inch. So, as long as you realize that the "20-nH/inch rule" is only a very crude approximation, it can be useful when deciding if component lead length is likely to upset the operation of your circuit.
— Alan Bloom, N1AL, ARRL TA, 1578 Los Alamos Rd, Santa Rosa, CA, 95409; **n1al@arrl.net**

Hello Alan,

Thank you for the feedback! You are correct for the ideal case of the isolated wire. This is one demonstration of why this condition does not exist. For an actual case, however, the total field is proportional to log(big number) – log(R), which is finite and does increase rapidly as R approaches zero. The point is that the ***H*** field increases greatly near the wire as the radius is decreased. I should have put it this way instead of "…the total field…" It is helpful to know how the field near the wire varies with wire size to understand the mechanism of coupling.
— Gerrit, KJ7KV

Dear Gentlemen,

Thanks for an excellent article! I always appreciate a new viewpoint on what is often a confusing subject. I especially appreciated your adding mutual inductance to the models.

I must respectfully disagree with you that a common-mode choke is not a current balun. The common-mode choke, to the extent that it impedes common-mode currents and passes differential-mode currents does indeed force a current balance into the load.

Think of a load that is imbalanced — that is, grounded somewhere other than at the center of the load. The Ruthroff balanced impedance converter is a voltage balun because it attempts to force the voltages in the two parts of the load to be equal (with respect to ground) at the expense of a current imbalance. That is a voltage balun.

Your common-mode choke, when presented with an unbalanced load, will force the currents in the two parts to be equal at the expense of an unbalanced voltage (with respect to ground). That is a current balun.

It should be fairly obvious that you can have one or the other, but that with an unbalanced load you cannot force both the voltage and current to balance simultaneously. We really don't care in the current balun that no load gives an output node at ground potential — it's the current balance that counts, and that is still balanced: zero in both legs!
— Respectfully, Glenn Dixon, AC7ZN; **dixong@ieee.org**

Hello Glenn,

Thank you for the note. You make a good point about the symmetrical nature of the two baluns, that voltage balance is not to be expected from a current balun and vice-versa. I agree that the article doesn't make this distinction very well. The point is that currents are not forced to be equal in this circuit; the degree of current balance depends on the relative impedances between the windings and the rest of the circuit. In that sense, the common-mode choke is only part of an overall filter, and dependent on the rest of the circuit. The voltage balun forces its balance regardless of the rest of the circuit, however.

Thanks for the clarification.
— Gerrit, KJ7KV

An L-Q Meter (Mar/Apr 2006)

Hi Doug,

A sharp-eyed reader, Paul Kiciak, N2PK, has discovered an error in the source code which is in the ARRL files as part of my article. There is a line in the code which says:

Za.r = 1.0 / Rinstrument;

It should read

Za.r = 1.0 /Ri;

I cannot explain how this error appeared since my own current copy of the source code doesn't have this error. I do apologize.
— 73, Jim Koehler, VA7DIJ, VE5FP; **jark@shaw.ca**

Hi Jim,

Thanks for the note, and for the corrected program file, which we have posted at the Web site. Readers can find the corrected file in the March/April section of the file listings at **www.arrl.org/qexfiles.** Look for **3x06Koehler.zip**.
— Larry, WR1B; **wr1b@arrl.org**