Letters to the Editor

RF Power Amplifier Output impedance Revisited (Jan/Feb 2005)

Dear Sir:

In my letter in the Mar/Apr 2005 issue of *QEX*, regarding Bob Craiglow's article, I referenced a simple mathematical proof that an RF power amplifier (RFPA) was conjugately matched with its load when *the load was adjusted to draw maximum power*. In his response in the May/Jun *QEX*, Bob dismissed this proof claiming I did not "…adequately model all the essential features of a nonlinear, non-time-stationary RF power amplifier." The quoted statement is true. I did not model **any** nonlinear features. But this is irrelevant. The proof is independent of any linearity arguments. The only assumptions required are that there is a maximum in the power-output-load resistance relation and that the output current be differentiable by the load resistance. The proof stands, and despite Bob's conclusions, RFPAs are conjugately matched with their load when the load is adjusted to draw a maximum power. See also the numerical calculations below.

Regarding the value of the equivalent source resistance, I now acknowledge that Bob is correct and that the equivalent source resistance for a PA operating with fixed bias is $R_S = \Delta E_p / \Delta I_{p1}$, given by the relation $2\pi r_p / (\theta + \sin\theta)$. The key to me was Bob's remark in private correspondence that E_p should/could be regarded as the terminal voltage of a counter generator connected across the RFPA's output terminals. In this view, E_p can clearly be taken as an independent variable and this eliminates my objection to his procedure in differentiating E_p with respect to I_{p1}. For PAs operating with a fixed conduction angle, the equivalent source resistance is $2\pi r_p / (\theta + \sin\theta)$.

It is necessary to recognize that his equation (Eq B1) is correct in so far as it is a way to calculate the load current. In this equation, θ is a number. It is the operating angle of the RFPA. It is a parameter, not a variable.

The following results of numerical calculations illustrate these points. For these calculations I have taken $\mu E_g = 14.1493$ V, $\mu E_{bco} = -5.0$ V (E_{bco} is the bias beyond cutoff), $r_p = 1.0\ \Omega$ and the operating angle at maximum output as 120°.

Calculated RFPA characteristics, P_{max} at θ = 120°.

θ	R_S	I_{p1}	E_{p1}	R_L	*Power*
126.71	2.0852	2.5017	3.0	1.1992	3.753
123.99	2.0991	2.2627	3.5	1.5468	3.960
120.00	2.1224	1.9550	4.1493	2.1224	4.056
117.58	2.1382	1.7904	4.5	2.5134	4.028
113.75	2.1662	1.5880	5.0	3.2093	3.895

The output power is a maximum when the load resistance equals the source resistance. The RFPA is conjugately matched to its load at the load drawing maximum power.

In the immediate vicinity of the maximum power the RFPA is represented by a Thevenin equivalent circuit with an open circuit potential of 8.298 V and a series resistance of 2.1224 Ω.

However, the source resistance is not constant and the variation becomes appreciable for large changes in the load resistance, although on a plot of E_p against I_{p1} the deviation from linearity is difficult to identify.

Regarding Bob's Appendix C, his conclusion that if a test signal were phase displaced the source impedance would be reactive, leads to an impossible situation. If Bob is thinking (or perhaps if I am thinking that he is thinking) he will measure the impedance with a test signal with α = 0.0, I will need a pure resistance to draw maximum power. But if he is thinking of making the measurement with a signal where α ≠ 0, I need a reactive load to draw maximum power. I note also that the magnitude of the source admittance changes with α, even when purely resistive.

For example at α of 0.0° and 90° the measured impedances, both resistive, are 2.1224 r_p and 5.1151 r_p. Changing the phase of the test signal has changed the "source resistance" by a factor of more than two, with no change in the amplifier! Further, this effect is independent of the magnitude of the test signal. Even an *infinitesimal* test signal with α ≠ 0 causes the source impedance to break into resistive and reactive components and to change in magnitude as above. Bob's conclusions in this area are wrong. —*Bert Weller, WD8KBW,* **sodiumflame @sbcglobal.net**

Dear Doug,

Bert and I now agree that the source impedance is defined as $Zs = -\Delta E_p / \Delta I_p$. We also agree that the output or source resistance is $R_s = 2\pi\, r_{p1} / [\theta + \sin\theta]$ for a PA with fixed dc bias,[1] and $R_s = 2\pi\, r_p / [\theta - \sin\theta]$ if the bias, E_{bco}, is adjusted so as to hold the conduction angle, θ, constant with changes in E_g and E_p.[2] Call the first case *fixed* E_{bco} and the second *fixed* θ. Finally we agree that, for a *fixed* θ design, the load resistance that maximizes the power output for a fixed grid drive voltage, R_{Ld}, equals R_s and therefore gives a conjugate match.

However, for a *fixed* E_{bco} design, Bert maintains that the load that maximizes the power output is still equal to the source resistance and therefore still gives a conjugate match. However, I maintain that maximum output always occurs at $R_{Ld} = 2\pi r_p / [\theta - \sin\theta]$, which does not give a conjugate match for this case. The derivation of R_{Ld} in Appendix B gives the load for maximum power output as $R_{Ld} = 2\pi\, r_p / [\theta - \sin\theta]$, which depends only on the fixed parameters r_p and θ and is therefore applicable to both *fixed* E_{bco}, and *fixed* θ designs.

Below is a table, identical to Bert's *fixed* E_{bco} table, except that it investigates the region around a nominal load impedance of $R_{Ld} = 5.115\ \Omega$ instead of the region around to $R_S = 2.122\ \Omega$. The bias voltage required to adjust θ to 120° with $\mu E_g = 14.149$ V is now $\mu E_{bco} = -3.537$ V. The maximum power output for $R_L = R_{Ld}$ is 4.893 W instead of 4.056 W for $R_L = R_S$, as shown in Bert's table. This shows that the power output is 20% greater for $R_L = R_{Ld}$ than for a conjugate match ($R_L = R_S$).

Calculated RF PA characteristics.

θ	R_S	I_p	E_p	R_L	P_{out}
126.7	2.08	1.77	6.26	3.54	5.54
124.0	2.10	1.60	6.62	4.13	5.29
120.0	2.12	1.38	7.07	5.11	4.89
117.6	2.14	1.27	7.32	5.78	4.63
113.7	2.17	1.10	7.68	6.96	4.23

Bert's second criticism is that an RFPA's source impedance, $Z_S = -\Delta E_p / \Delta I_p$, cannot possibly be reactive. Certainly, ΔE_p is small and so is ΔI_p but this does not imply that their ratio is small. The instantaneous plate

[1]See Eq A1 with $E_{bco} = [E_c + E_b / m]$.

[2]See Eq A3 where the fixed bias term, $mE_{bco} = [mE_c + E_b]$, is replaced by Eq A2.

current, i_p, flows only during the conduction interval of $-\theta / 2 \leq \phi \leq \theta / 2$. Therefore, ΔI_p will tend to be somewhat in phase with E_p and, if ΔE_p is not in-phase with E_p, $-DE_p / DI_p$ may have a reactive component.

The output admittance is $Y_S = G_S + jB_S = -\Delta I_p / \Delta E_t$ where $\Delta E_t = -\Delta E_p \cos(\phi + \alpha)$, ΔE_p is the magnitude of the test signal, and α is its phase relative to the ac grid voltage. For a *fixed* E_{bco} design, the instantaneous plate current is given by Eq A1. When the small test voltage is added, the instantaneous current becomes $i_p = [\mu E_{bco} + (\mu E_g - E_p) \cos f - \Delta E_p \cos(\phi + \alpha)] / r_p$. Therefore the change in the instantaneous current is $\Delta i_p = -[\Delta E_p \cos(\phi + \alpha)] / r_p$. The corresponding change in the fundamental ac component of current, that is in-phase with the test voltage, is

$$\Delta I_{p_I} = 1/\pi \int_{-\theta/2}^{\theta/2} \Delta i_p \cdot \cos(\phi + \alpha) \cdot d\phi$$

so that

$$\Delta I_{p_I} = -\Delta E_p[(\theta + \cos(2\alpha)\sin(\theta)] / [2\pi \cdot r_p]$$

Therefore the conductance is $G_S = -\Delta I_{pI} / \Delta E_p = [\theta + \cos(2a) \sin(\theta)] / [2\pi\, r_p]$.

Similarly, the quadrature component of ac current, relative to the phase of the test voltage, is:

$$\Delta I_{p_Q} = 1/\pi \int_{-\theta/2}^{\theta/2} \Delta i_p \cdot \sin(\phi + \alpha) \cdot d\phi$$

so that

$$\Delta I_{p_Q} = -\Delta E_p[(\sin(2\alpha)\sin(\theta)] / [2\pi \cdot r_p]$$

Therefore the suseptance is $B_S = -\Delta I_{pQ} / \Delta E_p = [\sin(2\alpha) \sin(\theta)] / [2\pi\, r_p]$.

Most published RFPA analyses are for *fixed* θ designs while implementations are normally *fixed* E_{bco} designs. It appears possible, at least in theory, to design a *linear*, class C, SSB, RFPA with a *fixed* θ design, using a grid bias control circuit based on feedback of the RF output envelope voltage. This appears to offer an opportunity to design a high efficiency, linear, class-C SSB PA.
— *Bob Craiglow,* **craiglowrl@juno.com**

Gentlemen,

I'm glad to see new information in your debate but theories need testing for confirmation. That's what we've lacked for the last several years.

Certain assertions previously advanced on these pages remain unchallenged and unconfirmed. They include:

1. Resistive load-variation testing cannot confirm a conjugate match because a magnitude-only measurement at the load measures only the magnitude of an amplifier's source impedance. That type of test can refute a conjugate match, however. (W. Bruene, W5OLY, "Letters to the Editor," *QEX*, Jan/Feb 2001, pp 59-61.)

2. Resistance in an amplifier's source impedance implies the conversion of energy from one form to another. Absent radiative effects, only an ohmic resistance that converts electromagnetic energy to heat can explain that resistance. The question is: Is a measurement of an amplifier's s22 (output reflection coefficient) equivalent to a measurement of its source impedance? Or is it a measurement of the conjugate of its optimal load impedance? (D. Smith, KF6DX, "Resistance — The Real Story," *QEX*, Jul/Aug 2004, pp 51-53.)

3. The establishment of the ratio of voltage to current at a source is an indication of the load impedance and not necessarily of the source impedance. Example: A 3-W night light might be plugged into a 120-V ac wall outlet. That ratio V / I is much more than that of a 3-kW amateur amplifier drawing its full current. Does that imply anything about source impedance?

I'd say it's time to lay it on the line — *Doug Smith, KF6DX,* QEX *Editor,* **kf6dx@arrl.org**.

Quadrature Phase Concepts (Nov/Dec 2005)

Hi Doug,

I felt uneasy about the conclusion that quadrature techniques cannot eliminate one sideband in a receiver. After all, the equivalent of Figure 2 is in fact frequently used to eliminate the unwanted sideband in direct-conversion receivers.

It seems that there is an error in how the 45° phase shift in the signal path (not the oscillator path) is taken into account. When a sine wave of frequency f, given by $\cos(2\pi ft)$, undergoes a phase shift, the result can be denoted by $\cos(2\pi ft + c)$, for some c. A positive phase shift corresponds to advancing time t, so a positive phase shift is represented by positive c if $f > 0$, and negative c if $f < 0$ — and conversely for a negative phase shift. KC7FHP, however, unconditionally represents a positive phase shift by a positive c (first line of second column on page 21), even though the frequency here (in $F - O$) may be either positive or negative.
— *73, Pieter-Tjerk, PA3FWM,* **pa3fwm@amsat.org**

Hi Gary and Doug,

In my opinion, the conclusions made in this article are wrong. Hundreds of amateur and professional realizations successfully use the phasing method to make SSB without passband filtering.

Here is the mistake, according to me:

[Eq 5] and [Eq 6] are actually only correct if $F - O > 0$, and Gary implicitly thinks that they remain the same when $F - O$ is negative. If $F - O < 0$, the equations in fact take another form. So when Gary says (p 21): "Assuming $F - O$ is positive or $F - O$ is negative, it doesn't matter…" It actually matters, because the equations are not the same in the two cases. This is why the demonstrations that follow are incorrect.

The fact that the equations are not the same for positive and negative frequencies comes from the Hilbert Transform properties, in which sign of frequency is of crucial importance. For example :

HT[$\cos(2\pi F T)$] = $\sin(2\pi F T)$ if $F > 0$, and $-\sin(2\pi F T)$ if $F < 0$
HT[$\sin(2\pi F T)$] = $-\cos(2\pi F T)$ if $F > 0$, and $\cos(2\pi F T)$ if $F < 0$

Note that in Figure 2, p 21, if Gary had made a difference instead of a sum at the end of the treatments, he would have obtained *zero*, and would have achieved what he wanted: eliminating part of the frequency spectrum! He speaks of "subtraction" in the modulator (p 20), but unfortunately not in the demodulator!

Thank you for this article. It was a good opportunity for everybody to dive deeply into Hilbert Transform theory!
— *73, Christophe Bourguignat, F4DAN, 15, rue Poirier de Narçay, 75014, Paris, France*

Source Coding and Digital Voice for PSK31 (Nov/Dec 2005)

Hi,

I just read your interesting article on using PSK31 to transmit speech. In your system, unlike pure text-to-speech systems, you really do have access to the original accent, phrasing, and intonation information. You also have access to the timing info. In fact, the "time warping" goes to great length to remove this info, so it must be detecting it. Therefore, it doesn't seem unreasonable to encode some of this information with the signal, at some increase in bandwidth. One idea would be to have a special message that says "the next word is accented" or "the pitch curve slope is such-and-such" or "the playback rate slope is such-and-such." To determine how feasible this would be for the extreme

case of PSK31, a "thought experiment" such as you did would be required. Yet another idea for increasing fidelity would be to make a pitch range and spectral characterization of the source voice and transmit that as a header; this could be used to pick a synthesis voice that partially matches the original. At the very least it could be used to select between female and male voice synthesis.
— Best regards, David A. Jaffe, K6DAJ, **David.Jaffe@analog.com**

Hi David,

Thanks for your note. You're right that additional information could be transmitted to take care of the things you mention. I'm glad to see you're thinking about that. As you did, I also thought about transmitting information that would allow the synthesized voice to match the original. Of course, many methods exist for it that are well-documented.

Other neat things you can do with digital voice include: adding header information to CQs that indicates the interests of the caller, establishing store-and-forward operation phonegrams, which would be neat for soldiers overseas), interspersing image data and so forth. The ARRL regulation-by-BW proposal has something to say about all that, I think. Other ideas can be found in the reports of the ARRL Digital Voice Working Group, under Committee Reports from the last five years in the Announce territory of **www.arrl.org**.

As I've stated before, standards are firmly established only by their popularity, and not just by committees. It remains, therefore, for someone to get interested enough to start coding. I wish I had the time for it.
— 73, Doug, KF6DX

Quantifying SETI (Jan/Feb 2006)

One reader reports that he was referred to the following Web pages by author H. Paul Shuch, N6TX, to learn how to build an amateur SETI station and more: **www.setileague.org/editor/magic.htm** and **www.setileague.org/articles/minimeta.htm**. We're happy to pass those URLs along. *— Doug, KF6DX*

Pspice for the Masses (Jan/Feb 2006)

Hi Doug,

Just got the new issue — as usual, interesting. Though of no great consequence, the curve identifications in Figure 4 are reversed. Most readers will hopefully recognize this.
— 73, Bob Hicks, W5TX, **w5tx@w5tx.com**

Tech Notes (Jan/Feb 2006)

We want to clarify that in last issue's Tech Notes, Eq 5 is properly called the standard deviation of the mean. The standard deviation of the mean is an estimate of the standard uncertainty to be associated with the mean, or average, of a set of measurements or calculations. *— Ed.* □□

In the next issue of

QEX

Our Italian friends Paolo Antoniazzi, IW2ACD, and Marco Arecco, IK2WAQ, return with a study of helical feeds with disk reflectors for Phase-3E parabolic antennas at 2.4 GHz. They also look at Phase-5A, 10-GHz possibilities for future Earth-Mars communications. Extensive computer modeling is balanced by the authors' experience in microwave antenna construction. Don't miss this one! □□

Upcoming Conferences

40th Anniversary CSVHFS Conference, 27 – 29 July, 2006
Bloomington, MN
(Across from the Mall of America)
Call for Papers

The Central States VHF Society is soliciting papers, presentations, and Poster displays for the 40th Annual CSVHFS Conference to be held in Bloomington, Minnesota (across from the Mall of America) on 27 – 29 July, 2006. Papers, presentations, and Posters on all aspects of weak-signal VHF and higher-frequency Amateur Radio operation are requested. You do not need to attend the conference, nor present your paper, to have it published in the *Proceedings*. Posters will be displayed during the two days of the Conference.

Topics of interest include (but are not limited to):

- Antennas, including modeling/design, arrays, and control.
- Construction of equipment, such as transmitters, receivers, and transverters.
- RF amplifiers (power amps) including single-band and multi-band vaccum tube and solid-state preamplifiers (low noise).
- Propagation, including ducting, sporadic E, and meteor scatter, etc.
- Test Equipment, including homebrew, using, and making measurements.
- Regulatory topics.
- Operating, including contesting, roving, and DXpeditions.
- EME.
- Digital signal processing (DSP).
- Software-defined radio (SDR).
- Digitial modes such as WSJT, JT65, etc.

Generally, topics not related to weak signal VHF, such as FM repeates and packet radio are not accepted for presentation or publication. There are always exceptions, however.

Please contact either the Technical Program Chairman, Jon Platt, WØZQ, or the *Proceedings* Chairman, Donn Baker, WA2VOI/Ø at the the e-mail addresses below.

The deadline for submissions for the *Proceedings* is Monday, **1 May 2006**. For Presentations to be delivered at the conference the deadline is Monday, **3 July 2006**. The deadline for Posters to be displayed at the conference is Thursday, **27 July 2006**. (Bring your poster with you!)

Further information is available at the CSVHFS Web site (**www.csvhfs.org**). See "The 2006 Conference," and "Guidance for Proceedings Authors," "Guidance for Presenters," and "Guidance for Table-top/Poster Displays."

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