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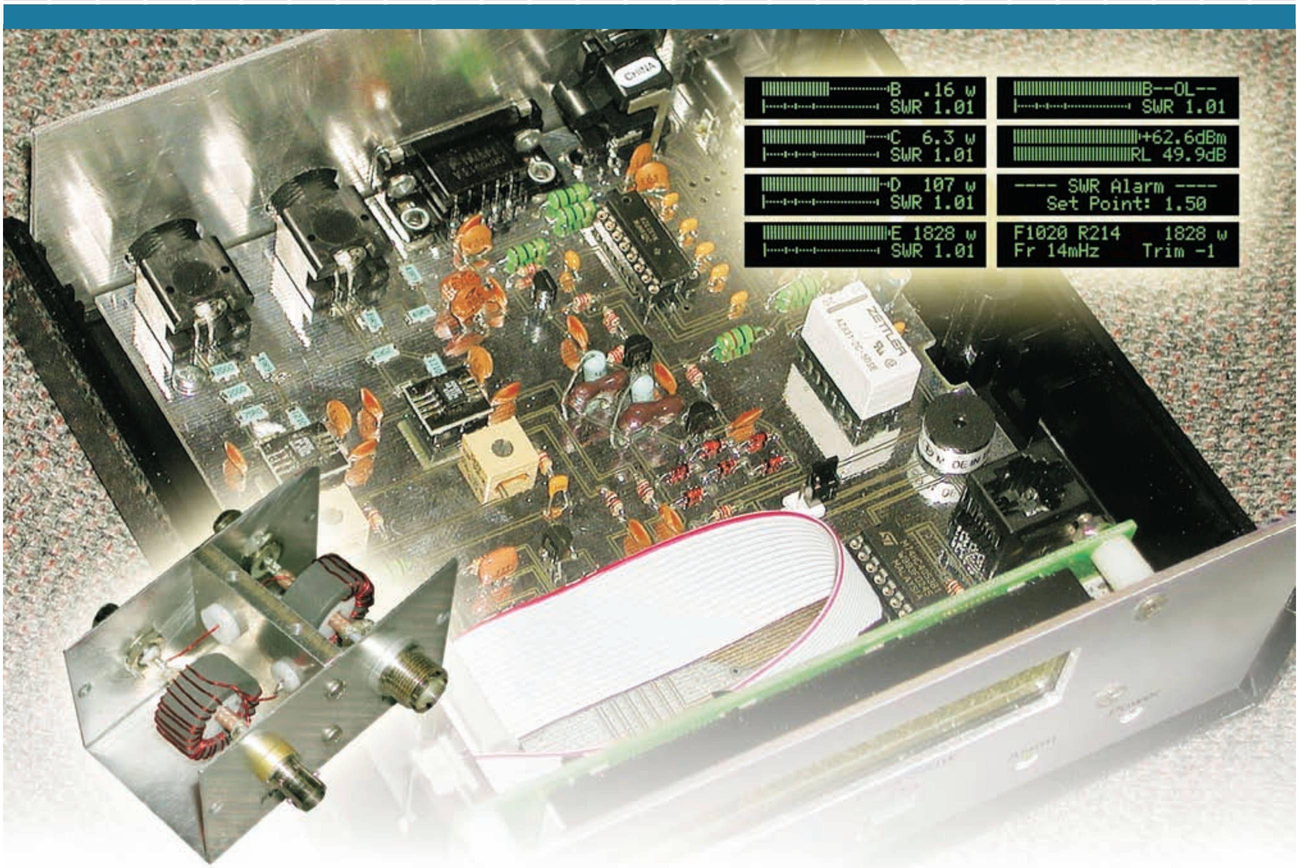

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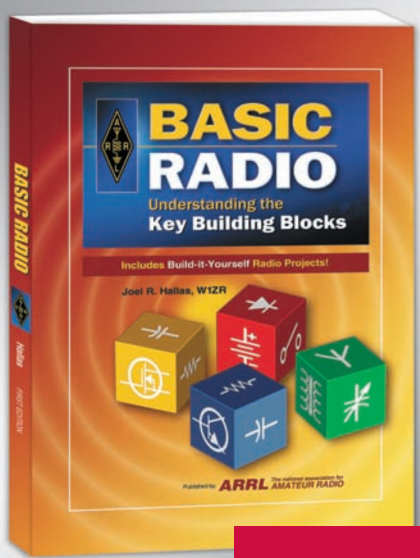

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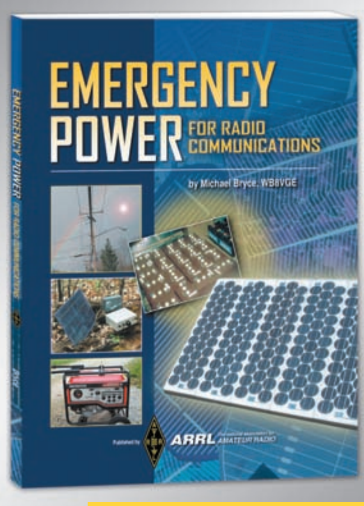

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About the Cover

Not your average, run-of-the-mill wattmeter. Read all about it, beginning on page 3.

Features

Columns

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Empirical Outlook

Know Your Readers

The original writer is not he who refrains from imitating others, but he who can be imitated by none. [François-René Chateaubriand (1768-1848) in *Le Génie du Christianisme, 1802*]

As a technical writer, you strive to explain some science or technology to readers who know less about it than you do, in a way they can understand. That definition might seem obvious, but it's worth keeping in mind every day, every hour you write. Quite often, you must convey rather complex information in simple terms. At other times, more elaborate explanations become necessary. Knowing when they *are* necessary is crucial to good technical writing. The business of straightforwardly explaining technical concepts involves trade-offs—some distinct and some indistinct—among clarity, readability and detail. It's tempting to state that those trade-offs need never occur; at the keyboard, though, you soon discover it's not easy every time.

Part of what readers understand about your topic before they start reading is unknown to you, but some prerequisite knowledge is assumed. If you were writing an installation and operation manual for a VCR, for example, you'd assume that readers know that the thing puts motion pictures and sound on a television receiver. You might also assume that readers expect to connect it to the receiver and to a source of power, although not precisely how to do that. What you cannot assume is their awareness that *bad mistakes during installation can electrocute them*.

Given that you know more about your topic than do your readers, you're a teacher. Some pupils may wonder why a video cable must be shielded and some may already know why. A good teacher tries to impart everything necessary for basic understanding to a class and also accommodates more advanced students by giving some insight into what lies beyond the basics.

In This Issue

Larry Phipps, N8LP, describes his high-performance digital wattmeter. Larry gives a thorough presentation of operating principles and performance, from the directional coupler and detectors to the display.

Randy Evans, KJ6PO, enlightens us about a *PSpice* program that you can get without money. Investing time to learn how to use it may pay substantial dividends. It's nice to know what you're doing at every point in the design process. At our invitation, H. Paul Shuch, N6TX, put some words together about "Quantifying SETI." Paul discusses certain technical issues related to the goal of communicating with extraterrestrial intelligence.

Our friends from "down under," Phil Harman, VK6APH, and Steve Ireland, VK6VZ, show us what they've done with the venerable Drake R4C receiver to "harmanize" it. Check it out.

Dick Hanson, K5AND, produces another great amplifier: the "8877 Lite." At 20 pounds, it will let you stay QRO while on the go.

In *Antenna Options*, Contributing Editor L.B. Cebik, W4RNL, addresses a critical antenna parameter: gain. In *Tech Notes*, I explain the universally accepted way of combining measurement uncertainties. The ARRL is contemplating its use to declare the accuracy of product-review test results. Happy new year.—*Doug Smith, KF6DX*, **kf6dx@arrl.org**

Managing Editor Robert J. Schetgen, KU7G, SK

For eight years, Bob lived at the center of the work storm here at *QEX*. We, his mourning friends and colleagues, dedicate this issue to him, the brilliant editor who contributed so vastly to our success.

A life member, Bob came to ARRL in 1983 as Technical Information Specialist. His other titles included Hints & Kinks Editor, Senior Assistant Technical Editor and *ARRL Handbook* Editor. He edited many other popular ARRL books, such as *QRP Classics*, *Hints & Kinks*, *Vertical Antenna Classics*, and *Packet: Speed, More Speed and Applications*.

Bob became a ham in 1963. He grew up in Southern California, attending Cal Poly University at Pomona. He always greeted you with a smile and a friendly word, even during his illness. He knew his job and performed it well. Bob is survived by his wife Ellen, a daughter and a sister, and a legacy and influence in journalism that will live on. May God bless him.

The LP-100 Wattmeter

A high-performance, microprocessor-controlled digital SWR/wattmeter with wide dynamic range

By Larry Phipps, N8LP

The LP-100 project is really an ongoing design exercise. The main meter chassis has remained fairly unchanged through several iterations of the project, but the directional coupler has undergone some radical changes, with several new designs still undergoing testing and modification. I will describe the first iteration of the project fully in this article. At the conclusion of the article, I will discuss some of the changes to the coupler, and will refer the reader to my Web site (**www.telepostinc.com**) for the latest details.

49100 Pine Hill Dr
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larry@telepostinc.com

Here are the basic specifications/features of the meter…

- Coverage from 0.5 to 54 MHz for rated specs with the specified coupler designs
- Auto-ranging scales covering 0-2, 20, 200 and 2000 W
- Modular design to allow for different couplers
- Scale for dBm measurements, ~ –15.0 dBm to +63.0 dBm in 0.1 dB steps
- Power display resolution of 0.01, 0.1 or 1 W depending on scale.
- Power accuracy better than ±.5 dB to <10 mW, typically 5% after calibration
- Displays actual power delivered to the load (Fwd minus Ref)
- SWR display resolution of 0.01
- Overall SWR accuracy typically within ±0.1 of actual down to 100 mW
- Return Loss (RL) display of 0 to 49.9 dB
- State-of-the-art PLED display with screen saver
- Fast responding logarithmic bargraphs for power, SWR and RL
- Peak-hold numerical power readout with fast and slow time constants
- SWR Alarm system with set points for 1.5, 2.0, 2.5 and 3.0
- Separate calibration screen
- Serial port for external software control

• Virtual Control Panel (VCP) software for *Windows*
• Remote control supported directly or through *TRX-Manager*

Among the most interesting aspects of the design of the LP-100 are the coupler, the logarithmic detectors, the microcontroller and the *Windows* VCP software. The area that received the most attention (and created the most frustration) was the directional coupler. Over 50 designs were built and tested. Each had its good points and bad, but I settled on two for the final versions.

Coupler

Before discussing coupler design, let's define a few terms.

Directivity—the difference (in dB) between forward (Fwd) and reflected (Ref) power when the output port is terminated in the design impedance… 50 Ω in this case.

Insertion Loss—the loss (in dB) between the input and output ports

Coupling Factor—the loss (in dB) between the input port and the forward sample port, or between the output port and the reflected sample port.

Input Return Loss (RL)—the amount of power reflected back into the source (in dB) when all other ports are terminated with accurate 50-Ω terminations.

I used eight criteria to evaluate my coupler designs.

1. 50 Ω Z on all ports
2. cover 160 to 6 m in one design.
3. >30 dB directivity over the entire frequency range
4. Level passband response over the entire range
5. >25 dB input RL presented to the transmitter over the entire range
6. Insertion loss < 0.2 dB
7. 1500 W continuous power handling capability
8. Easy to duplicate

First, let's review how a coupler operates. Probably the grandfather of broadband frequency-independent directional couplers is the design developed by Warren Bruene, W5OLY, of Collins Radio. The best coupler article I've read is his "An Inside Picture of Directional Wattmeters," published in April 1959 *QST*.[1] It is must reading for anyone interested in the subject because it analyzes the Bruene Bridge as well as other coupler designs.

Underlying Bruene's design, as well as that of many other couplers, is a current sampling transformer in series with the transmission line combined with a circuit to sample voltage across the load. The voltage sampler may be a capacitive or resistive divider, or a transformer. The current and voltage samples are combined so as to isolate the forward and reflected components. This is possible because the voltage and current of the forward wave are in-phase, while the voltage and current of the reflected wave are out-of-phase.

Another type of coupler, which at first would seem to be radically different from these, is also related. It is the line-section or loop coupler as pioneered by Bird. In this design, a terminated loop is brought into proximity to the center conductor of a transmission line, and provides a Fwd or Ref sample, depending on which end of the loop is terminated. This is possible because the loop provides two kinds of coupling at the same time. Inductive coupling to the line produces a current sample, and capacitive coupling provides a voltage sample.

My first criterion—50-Ω impedance at all ports—eliminated many published coupler designs, including the Bruene bridge, which must be terminated in high Z loads for the forward and reflected ports. After reviewing the available 50-Ω designs, I settled on a couple to try. One was the Tandem Match,[2] covered in depth in a *QST* article and recent issues of the *ARRL Antenna Book*. Another version uses a binocular ferrite core and is described in a comprehensive article by Michael G. Ellis.[3] These designs, and others such as the Stockton coupler, are descendants of a design using two toroidal transformers patented by Sontheimer and Fredrick in 1969. See Fig 1.

By using identical transformers, wired as shown, all four ports will have the same impedance. The current transformer, T1, has a one-turn primary in series with the main transmission line, while the voltage transformer, T2, has its secondary connected across the transmission line. If the two transformers have the same turns ratio, and the current sample is terminated in a 50 Ω load, then the voltage across the termination will be the same as the voltage from the voltage sampler. If the two voltages are combined, then the phase relationship mentioned above will cause the Fwd components to add and the Ref components to subtract.

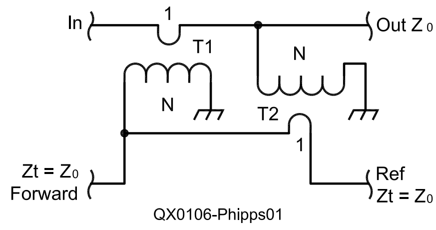


Fig 1—Tandem Match Schematic.

The correct transformer turns ratio, N, is determined by balancing a number of factors, including keeping T2's reactance high to prevent loading of the line, and keeping T1's winding length short to minimize stray capacitance which reduces high frequency response. N also determines the coupling loss, and hence the required Fwd port termination power rating, since the power at this port is proportional to $1/N^2$.

The first version of this design that I tried used a ferrite binocular core, which allowed very simple construction techniques. I discovered that in order to meet my bandwidth objective with available cores, T1 and T2's secondary windings had to be kept under 10 turns. Since the only commercially available binocular core that could possibly handle 1500 W is about 1-inch long, each turn takes nearly three inches of wire. Since the upper frequency response is largely related to the length of the wire in the secondary windings, keeping the coupler functioning at even 28 MHz limited my design to fewer than 10 turns.

But if T1's secondary is only 8 turns, for example, with an input power of 1500 W, the forward coupling port termination must dissipate nearly 25 W. Dissipating 25 W in a small box, while keeping the temperature rise to acceptable levels is not easy.

My second attempt was the Tandem Match coupler. One problem I noted with this design duplicated the experience related in an article by N2PK.[4] He discovered a problem with the input RL of the Tandem Match at 1.8 MHz. He indicated that this was a factor of inadequate X_L at 1.8 MHz, thereby shunting the line at that frequency with a relatively low value inductance.

By replacing the powdered iron cores of the Tandem Match design with higher permeability ferrite cores, this problem is eliminated. I used Fair-Rite # 5961000501 cores (Amidon part # FT-114-61), with an initial permeability (μ_i) of 125, compared with 8 or 10 for the powdered iron cores used in the two Handbook designs. Although the ferrite cores provide excellent RL performance at 1.8 MHz, because ferrite is more easily saturated at high flux levels than is powdered iron material, I had to increase the size of the cores

[1]Notes appear on page 13.

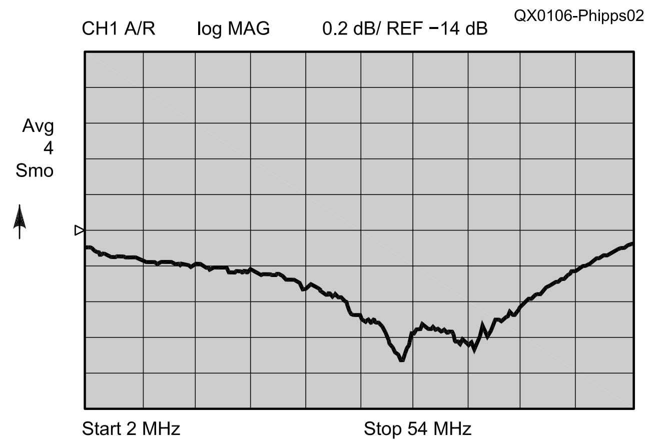


Fig 2—Coupler directivity with a 50 Ω load. Vertical scale is 5 dB/div; mid-scale is –30 dB.

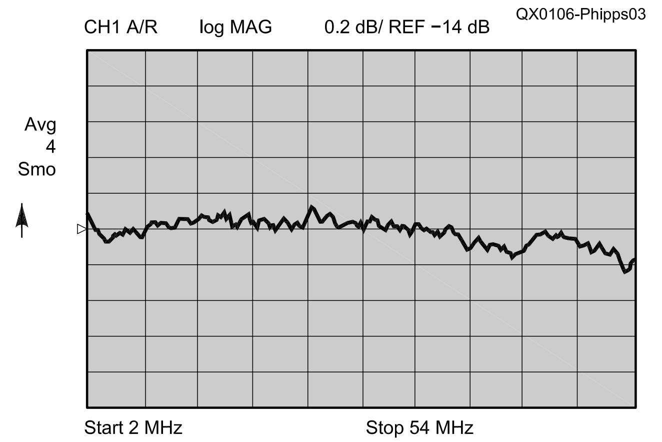


Fig 3—Coupler directivity with a 75 Ω load. Vertical scale is 0.2 dB/div; mid-scale is –14 dB.

considerably. I determined experimentally that the chosen cores were the smallest available cores that would safely handle 1500 W.

The larger cores necessitated the use of longer windings, so I reduced the number of turns from the 31-40 used in the Tandem Match article to maintain good response at 54 MHz. As a compromise between Fwd port termination dissipation, flux density and 54 MHz response, I settled on 26 turns.

Even with fewer turns than the original design, I had to spend time to achieve an acceptable input RL at 54 MHz. Stray capacitance and inductance from mechanical layout was the culprit here. By making the length of the primary windings as short as possible, I was able to achieve close to 30 dB RL at 54 MHz.

To measure SWR accurately, directivity must exceed 25 dB. I found that it is difficult to achieve reliable measurements when testing a 50 Ω load. This has to do with the vector addition of the Fwd and Ref samples. It is possible to get wildly optimistic readings of directivity at 50 Ω, because while the Fwd power can have a 6 dB maximum range of addition, the reflected power can theoretically be zero under certain conditions. I found it was much better to use a load with an accurately known error. I chose 75 Ω for my tests, and computed directivity from the measured SWR. Table 1 shows the range of SWR accuracy available with various values of coupler directivity, for a 75-Ω load.

With even 20 dB directivity, the SWR error is sizeable. An indicated SWR of 1.5 could actually be anywhere between 1.2 and 1.9. Unfortunately, 20 dB is probably typical for many ham-grade SWR bridges across the HF spectrum, and they're probably much worse at 50 MHz.

The measured directivity of this coupler with 50 and 75 Ω terminations is shown in Figs 2 and 3. These graphs were output from my HP-87510A Vector Network Analyzer (VNA). Especially noteworthy is the 75 Ω directivity, which is within 0.2 dB of the ideal value of 13.98 dB, verifying a directivity of >30 dB.

To obtain the best high-frequency directivity, the toroids needed to be rotated carefully. I am convinced that this was a result of stray coupling: both inductive and capacitive. For this reason, I took care to make the design symmetrical to balance out the effects

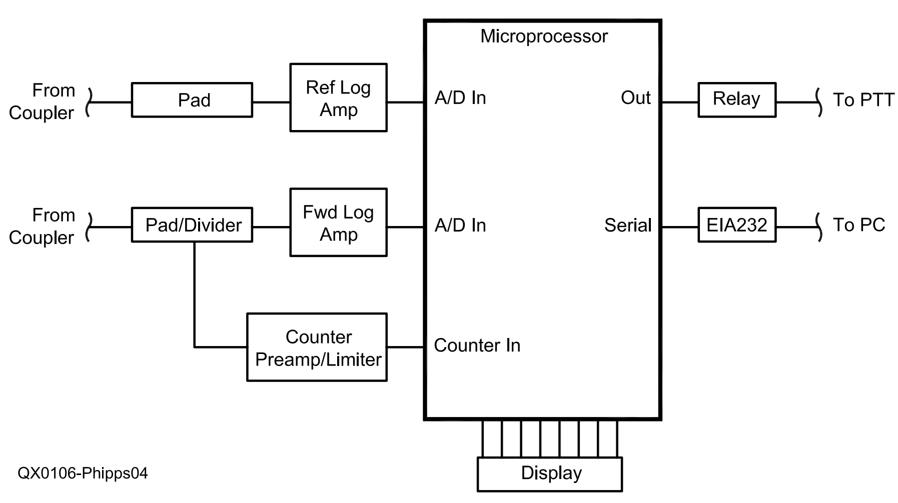


Fig 4—LP-100 block diagram.

Table 1—SWR Measurement Range versus Coupler Directivity for a 75 Ω Load

Directivity	*Minimum SWR*	*Actual SWR*	*Maximum SWR*
10 dB	1.00	1.50	4.22
15 dB	1.04	1.50	2.36
20 dB	1.22	1.50	1.90
25 dB	1.33	1.50	1.70
30 dB	1.40	1.50	1.61
35 dB	1.44	1.50	1.56
40 dB	1.47	1.50	1.53

of this parasitic coupling.

The coupler is built in a separate shielded box, and connected to the LP-100 through a pair of short coaxial cables. Employing a separate coupler eliminates the problems associated with having a small box on the operating table connected to heavy, stiff feedlines. Furthermore, it removes high power levels from the vicinity of the microprocessor and other sensitive circuitry, as well as reducing the possibility of RF noise from the microprocessor being radiated into the station receiver.

Logarithmic Detectors

The genesis for the LP-100 detectors came from the auto-tuner section of a mini-beam project. I sought a fast, efficient tuning algorithm permitting near instantaneous tuning of the mini-beam elements. To make this work, it was especially important to resolve accurately slight differences in SWR at very low power levels, because very little power is available when the antenna is far from resonance.

To compound the problem, a high-directivity coupler—like the one I'm using—can place high demands on the detector when measuring a high-quality load. This is so because the signal

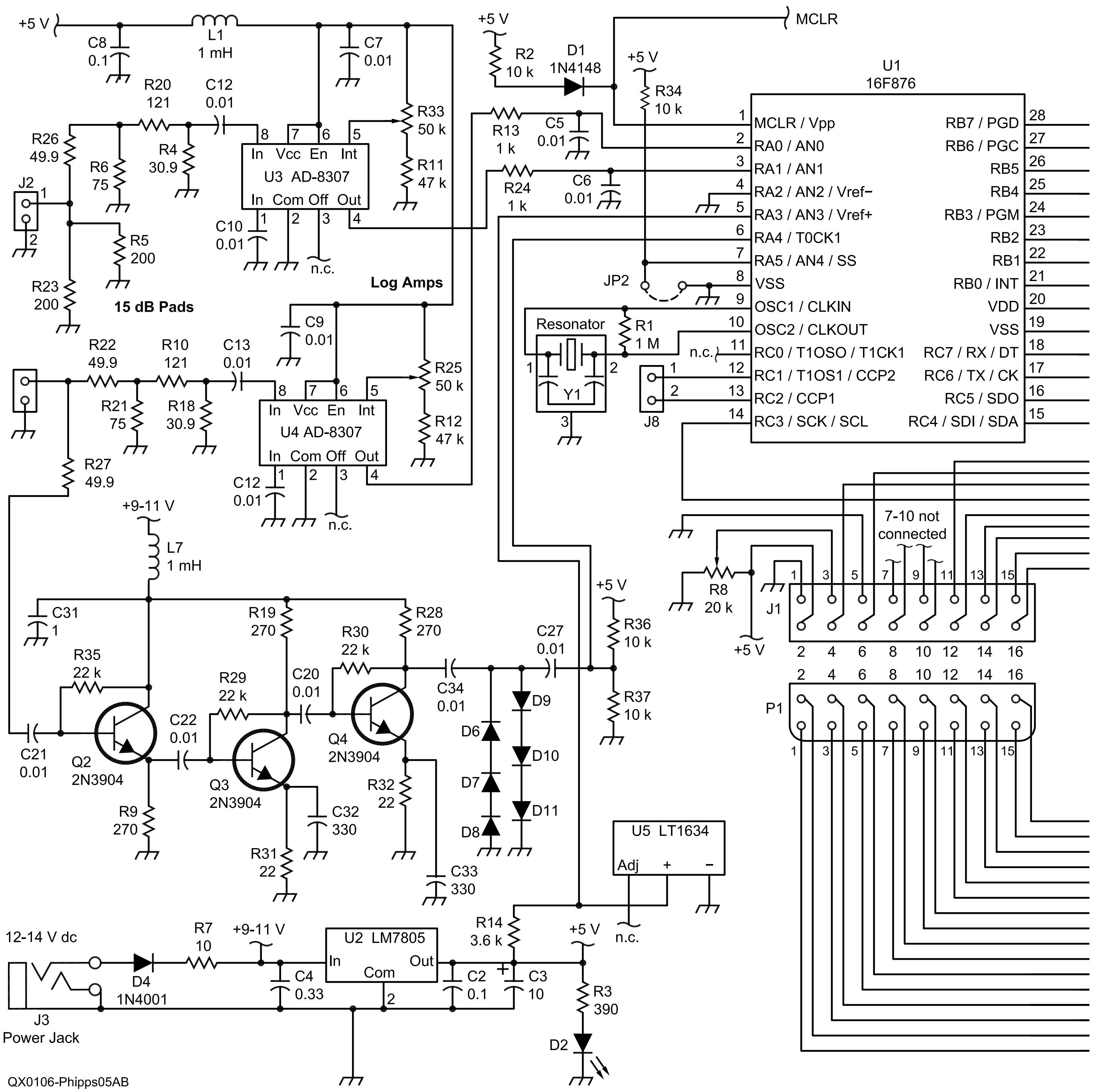
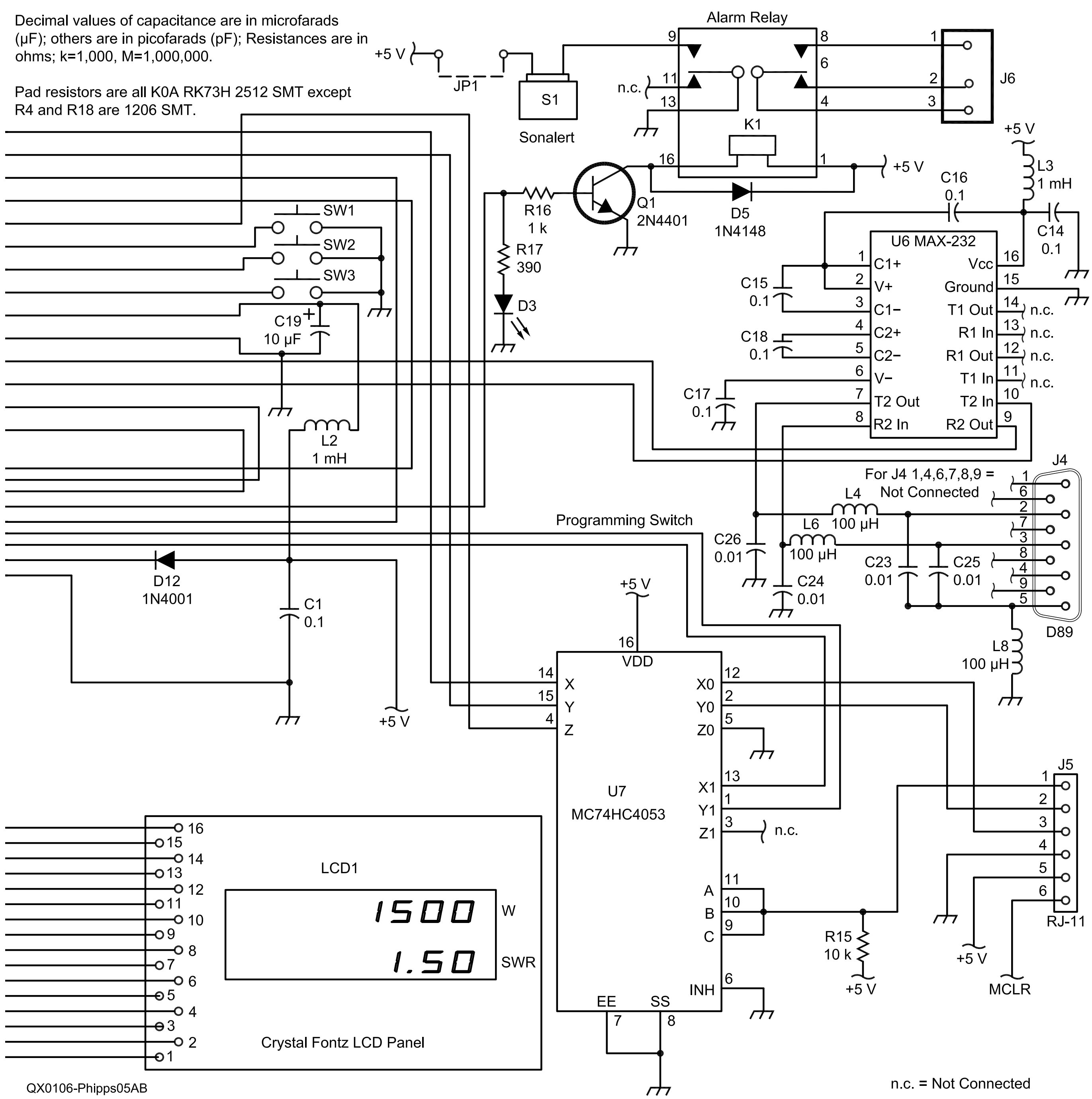


Fig 5—LP-100 schematic diagram.

at the reflected port is very small. For 100 W input, the reflected power detector must provide accurate level data for an input signal of 1 mW or less. I ruled out op-amp corrected diode detectors, even though they have been used effectively in circuits such as the Tandem Match and Lewallen's QRP Wattmeter.[5] To handle a wide range of power with these circuits requires an array of precision resistive dividers to set the ranges. I wanted a meter that would continuously display power over a very wide power range with consistent accuracy at any level in the range. So how might we detect these very small signals while still handling the larger signals generated by 1500 W without overloading?

Enter the logarithmic detector. Based on work I had done in another project, I knew that a log-detector would handle the dynamic range needed for these projects. I chose the AD-8307 logarithmic detector from Analog Devices.[6] This chip was designed for use in the cellular telephone market and is used to convert RF levels into a dc Received Signal Strength Indicator (RSSI) output. The chip can accurately translate an RF input with about 80 dB dynamic range into a dc output corresponding to the



log of the input signal. When connected to a high quality directional coupler, a pair of these chips is capable of providing accurate forward and reflected power detection over a dynamic range of almost 50dB at an SWR of 1.1. In terms of the LP-100, this corresponds to a range of ~20 mW to 2000 W for full SWR accuracy, or less than 1 mW if only forward power accuracy is needed. The chip is capable of operation from dc to over 500 MHz, and provides a linear RSSI output of 0 to 2.5 V, with a conversion factor of 25 mV/dB and a typical accuracy of ±0.25 dB.

For an input power of 5 W, the power available at the Fwd power port, expressed in dBm, is roughly +7 dBm (coupling factor is approximately 30 dB). The signal at the reflected power port, for an SWR of 1.1:1, is ~26 dB lower, or –19 dBm. The AD-8307's rated linear input range is –65 to +10 dBm, so this is no problem.

But since we want to measure power up to about 2000 W (+63 dBm) we see that 30 dB coupling loss results in +33 dBm (2 W) of power at the input of the forward detector, grossly exceeding the AD-8307's maximum permitted input signal level of +20 dBm. To keep the AD-8307's input level within its recommended linear operating range, therefore, we will need a pad between the coupler ports and the log-detectors. My design uses 25-dB total attenuation, split 10 dB at the coupler outputs and 15 dB at the LP-100 inputs. The pads also help isolate the coupler and log-detectors from variances in the connecting lines.

In round numbers, the maximum signal that the AD-8307 would normally see for approximately 1500 W input is:

+62 dBm (1585 W) – 30 dB (the coupler loss) – 25 dB (the pad) = + 7 dBm

With this input power and an SWR of 1.1, the signal at the reflected power port will be –19 dBm. Now, what happens with only 5 W input? Since 5 W is ~25 dB below 1500 W, all the values must be reduced by this amount. The critical port is the reflected port, so let's look at that. The signal available there would be –44 dBm (19 + 25). This is still no problem since the AD-8307 will accurately handle signals as low as –65 dBm. The AD-8307 still has ~20 dB of range left, indicating that power as low as ~50 mW (20 dB below 5 W) can provide accurate results at an SWR of 1.1 or higher. I used an SWR of 1.1 for this illustration since it represents a challenge in terms of testing the limits of the noise floor on the Ref port detector.

LP-100 Microcontroller

The "brain" of the LP-100 is the microcontroller. This circuit is based on the 16F876 PIC chip from MicroChip.[7] It is an 8-bit controller with 8kB of EEPROM program memory (14-bit width), 368 bytes of RAM and 256 bytes of EEPROM working memory. The chip has an internal A/D converter that can accept up to 5 analog inputs, and provides up to 22 I/O pins, which can act as TTL/CMOS compatible inputs or outputs. The LP-100 uses these ports for interfacing the log-detectors, implementing a frequency counter, PLED screen, switches, LEDs and an SWR alarm relay. See Fig 4.

Operationally, the PIC sequentially samples the RSSI outputs of the forward and reflected power log-detectors using two of its A/D inputs. The inputs have 10-bit resolution and use a precision 2.56-V reference. The AD-8307 output is set so that the maximum input power of 2000 W produces a dc output of 2.56 V, corresponding to an A/D output of 1023. Each sample of the A/D converter is therefore 2.56/1024, or 2.5 mV. Since the AD-8307 has an output slope of 25 mV/dB, the A/D readings equal 0.1 dB per step. (The inter-sample interval is sufficiently short that the transmitter output power does not materially change between the forward and reverse samples even when operating SSB.)

The firmware calculates the "true" power in dBm and W then calculates RL and SWR using the forward and reflected power samples. I define "true" power as the net power delivered by the source to the load, or forward power minus reflected power. Thus, the meter always displays the actual power delivered to the load, regardless of SWR. The values are displayed on a 2×20 character PLED panel. I originally used an LCD display, but found the PLED was better looking and has faster response. It is plug-in compatible with the LCD. The various displayed screens are shown in Fig 6.

The first range, labeled "A," is not shown. It is an auto-ranging screen that displays power at any level. The next four screens display power in ranges of 2, 20, 200 and 2000 W, along with SWR. The fifth screen shown in the figure is the overload indication that will appear if one of the fixed-power screens is selected and the power exceeds the limit of that screen. The next screen shown displays power in dBm along with RL in dB, followed by the Alarm Set screen, described later.

The final screen is the calibration screen. The CAL screen shows the raw Fwd and Ref power sample values simultaneously, making it easy to calibrate and match the two log-detectors. It also allows for matching the power readout to a reference meter at 1 MHz intervals from 1-54 MHz. The current frequency and A/D counts are displayed, along with the "trim" value, which is adjusted with the UP and DN buttons. This value is stored in non-volatile memory, indexed to frequency.

The power scales use a logarithmic bar-graph display, which is comprised of custom characters to allow the display of 36 vertical bars. These bars respond instantaneously to changes in power. This was done in part to make it easy to adjust things like an antenna tuner or linear amplifier. Although the response is quick enough to follow SSB voice peaks, the display is smooth due to a combination of the logarithmic scale and over-sampling of the A/D inputs.

The maximum power scale is changed by pressing the SCALE button, which increments the scale once for each button press. After seven presses, you return to the original scale. The current scale selection is stored in the PIC's non-volatile memory, so that if you lose power, the LP-100 will remember your last selection. The first position is an auto-ranging screen, which allows continuous display from 0.01 W to 2000 W. If one of the fixed power screens is used, and the power level exceeds full scale on the bar-graph, the over-limit message is displayed, indicating the operator should increment to a higher scale.

Next to the bar-graph display is a numerical power value. Unlike the bar-graph, the numerical readout is peak reading, so that it is easy to monitor power while operating SSB or CW. The peak-hold algorithm freezes the reading unless a higher value is encountered during the hold period, in which case the higher value is displayed and the hold timer reset. The FAST/SLOW button disables the hold function to allow the numerical readout to follow changes during adjustment of antenna tuners if desired.

The display resolution of the numerical readout varies with the maximum value of the scale, from 0.01-W resolution on the 2-W scale to 1-W resolution on the 200 and 2000-W scales.

Below the power bar-graph is the SWR bar-graph. Next to the SWR bar-graph is the numerical SWR readout, with a display resolution of 0.01. The

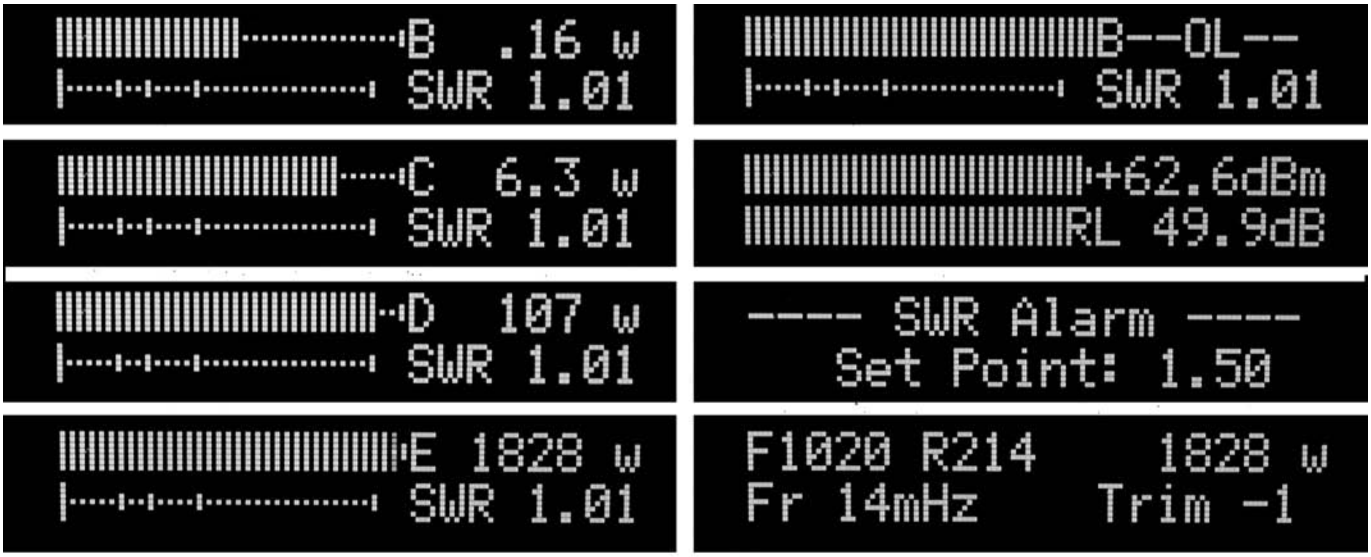


Fig 6—LP-100 Display Screens. (N8LP).

SWR display is the same for all four "W" scales. An interesting idea I borrowed from the on-screen display of one of my TVs, is the use of markers to show the limits of the bar-graph. This makes it easy to see what portion of the full range is being displayed. Additional larger markers are provided on the SWR scale for SWRs of 1.5, 2.0 and 3.0.

The dBm scale displays power in 0.1 dB steps from about –15 dBm (the noise limit of the instrument in a typical ham shack environment) to +63 dBm (2000 W). For this screen, the SWR display is replaced with a RL scale. The RL scale is limited to 49.9 dB maximum, even if RL is higher, to prevent overflow in some of the internal calculations.

The LP-100 also includes an SWR Alarm function. The "Alarm Set" button controls this function. When pressed, the button changes the display to an Alarm Set screen, as seen earlier in Fig 6. Repeatedly pressing the button cycles through the set-point choices... Off, 1.5, 2.0, 2.5 and 3.0. When you have reached the choice you want, stop pressing the button and the screen will return after about one second to the power screen displayed before entering the Alarm Set mode. The set-point choice is stored in EEPROM. If the SWR reading reaches your set point in any mode, the front panel "Alarm" LED will light, the Sonalert will beep and the relay will close.

The Alarm is deactivated by correcting the over-limit condition. There is a delay of about five seconds built into this action, to allow for antenna tuning without constant interruption by the relay. Setting the Alarm Set mode to OFF at any time will also deactivate the alarm, and cancel any current alarm. I included a jumper in the circuit to permanently disable the Sonalert if desired.

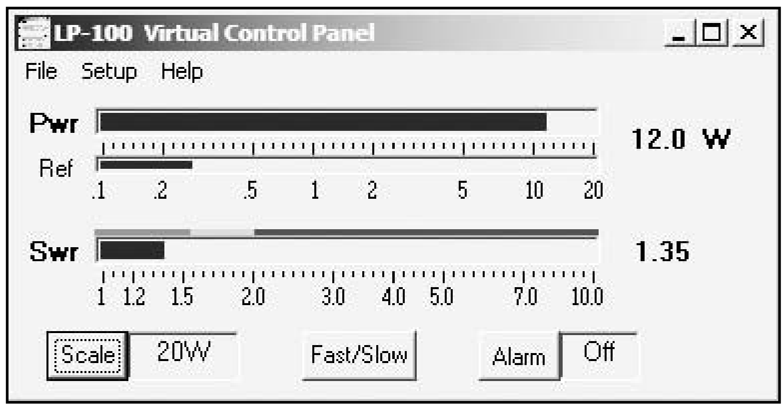


Fig 7—LP-100 Virtual Control Panel.

The relay contacts are brought out to the rear panel to allow disabling the PTT connection to a linear amplifier or transmitter. The relay terminates in a pair of RCA type PTT connections, which should make for easy interfacing to most amplifiers. This is a great way to protect those expensive finals from a catastrophic antenna failure!

I have included a serial port for software control of the LP-100, which terminates on the rear panel with a standard DB9 connector. I have also included a programming port to allow the PIC to be reprogrammed in-circuit. The programming port connects through an RJ-11 jack on the PCB to a Basic Micro[8] programmer. Anyone who doesn't need either of these last two capabilities can simply delete the associated parts from the project.

Software Virtual Control Panel

I have developed Virtual Control Panel (VCP) software to provide computer control of the LP-100, as well as offering extended functionality. The *Windows*-based VCP mirrors the controls and display of the microcontroller box, but adds calibrated scales below the bar graphs (see Fig 7). Further information can be found in the SteppIR VCP Help. A copy of the help file is also available on my Web site. The VCP program is available as a free download.

In addition, I am working on a charting program to automatically create an SWR-versus-frequency graph for a selected frequency range. The chart could be printed and saved for future reference. In order to use this feature, the LP-100-Chart software must be used in conjunction with *TRX-Manager*[9] to control the transmitter.

The LP-100 is also supported directly in *TRX-Manager* allowing remote monitoring of LP-100 basic parameters using *TRX-Manager*'s Remote telnet mode. Details of all of the LP-100 software are available on my Web site (**www.telepostinc.com**). The LP-100 can also be controlled and monitored over any network connection, including the Internet. For details of doing this, see my article in the Oct 2005 *QST* entitled, "Using Networked Equipment for Remote Station Control."[10]

Construction

Most of the parts for the LP-100 are mounted on a single, double-sided PC board, except for those associated with the PLED display and coupler. Fig 8 shows the completed board, with the PLED board mounted in front of it on the front panel. The PLED board is connected to the main board with a 16-pin header and ribbon cable.

A version of the main board with silk-screened component placement, and a solder mask, will be available on my Web site for those wishing to replicate this project. A pre-programmed PIC chip will also be available. Since I have already had numerous requests for a kit of parts for this project, I plan to offer that as well.

The main PC board mounts on four standoffs in a Pac Tec CM6-225 black ABS plastic case. This case is available in plain ABS or with an RF shield coating. The rear-panel PC board-mounted BNCs are secured to the rear panel. All connectors and switches are PC board mounted for ease of assembly and reliability. All parts are through-hole except the resistors in the input pads and dividers. This was necessary from a performance standpoint, but the parts are easier to handle than you might imagine.

K8ZOA did the drilling and milling of the front and rear panels for my prototype. The panels are brushed aluminum, and the labeling is an overlay I produced in Microsoft *Word*, which is printed on adhesive vinyl transparency material.

The prototype coupler is mounted in an aluminum clamshell box with SO-239 connectors for the XMTR and ANT ports, and BNC connectors for the FWD and REF ports. The main connectors could be replaced with N connectors if desired. Figs 9 and 10 show the prototype coupler in detail. I used external 10 dB BNC inline coaxial pads from Mini-Circuits[11] in my prototype. It is important that these two ports always have a load to prevent high voltages from developing on the transformer windings when transmitting. I

Fig 8—Main Board. (N8LP).

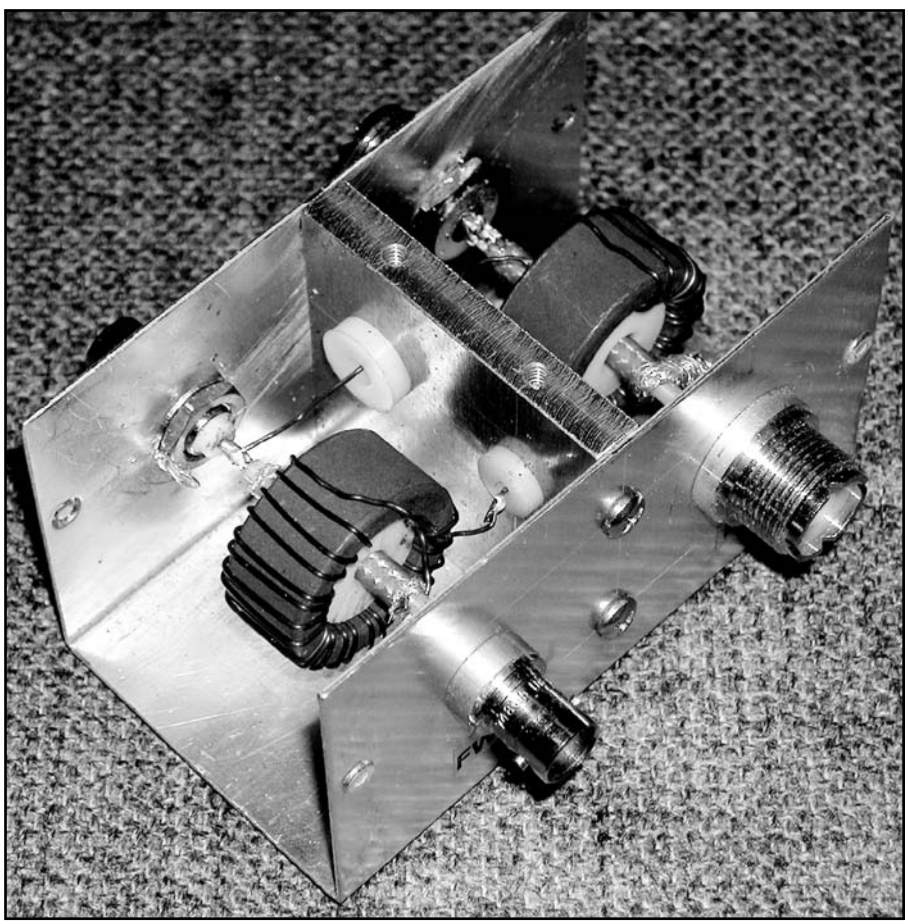


Fig 9—LP-100 Coupler photo. (N8LP)

plan to integrate the pads into the case in the future.

The coupler box dimensions are important. Don't use one that's much smaller than mine, which might result in stray coupling degrading high-frequency directivity. Also, be sure to include the shield partition between sections to improve isolation of the transformers. I used grommets in the shield to connect the transformer secondaries between sections.

The primary of T1, which goes between the XMTR and ANT connectors, is a short piece of RG-142 Teflon coax. The shield is grounded on one end, and left open at the other. This provides electrostatic shielding, and reduces the effects of capacitive coupling. (If you ground the shields at both ends, it will create a shorted turn and render the coupler useless.) The primary of T2 is connected between the FWD and REF connectors.

Both secondaries are 26 turns of #26 (AWG) enameled copper wire, spaced evenly over about 2/3 of the cores. The cores are supported by short lengths of plastic rod, drilled down the center. I used Teflon, but any decent low-loss plastic should work. I have seen small sink plungers with plastic handles of the right diameter, which could be cut and would probably work okay.

A support package with complete assembly details, schematics, parts list, mechanical drawings, panel templates and graphics files for the front and rear panels, is available on my Web site if you want to duplicate this project.

The package should be downloaded by anyone interested in replicating the project. The package contains a wealth of information that will make the project easier to build.

Performance

Tables 2 and 3 show the measured performance of the LP-100 prototype at various frequencies and loads. In general, power accuracy is dependent upon the reference against which the LP-100 is calibrated. (Calibration is necessary because of the manufacturing variation of the log amps.) After calibration, the meter will match the reference exactly at the calibration level and should be within ±0.25 dB at any power level. Table 2 shows a comparison of the LP-100 to Bird 4410A and Alpha 4510 power meters. The source is my Elecraft K2.) These are exceptionally good meters, but even at this level, it is apparent that there are some differences. The LP-100 matches the Bird exactly because it was calibrated against it. A spot check of the meters at 100 W, without further calibration, showed a ±3% variation between the three instruments.

SWR accuracy was calculated to be better than 5% based on device specs. The SWR readings in Table 3 are compared with those from my HP-87510A VNA. The table shows the displayed SWR for nominal load values of 50, 75 and 25 Ω, using precision terminations. Readings were taken at +20 dBm (100 mW) for all load values, using my HP-8640B signal generator.

SWR results below 1.1:1 would require 0.5 W to reach full accuracy. This is due to the reflected power signal being below the noise floor of the log-detector as discussed in the coupler description above.

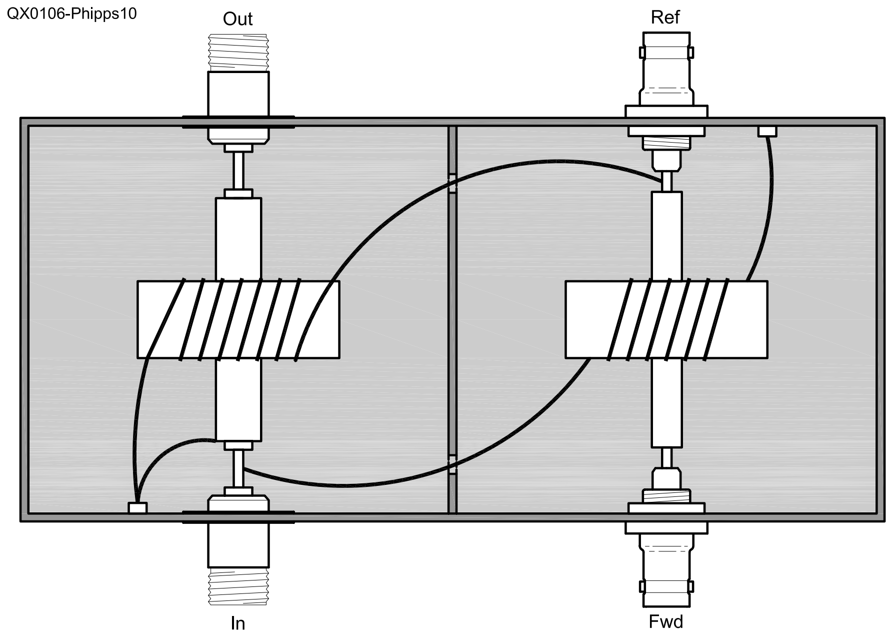


Fig 10—LP-100 Coupler drawing.

Table 2—Power measurements of the LP-100

LP-100 measurements (watts) compared with those of Alpha 4510 and Bird 4410A wattmeters

f (MHz)	*LP-100*	*Alpha 4510*	*Bird 4410*
1.80	7.10	7.20	7.10
3.50	9.50	9.45	9.50
7.00	9.60	9.45	9.60
14.00	8.40	8.45	8.40
21.00	7.80	7.70	7.80
28.00	6.70	7.00	6.70

Table 3—SWR Accuracy of the LP-100

LP-100 measurements (SWR) compared with those of an HP-87510A VNA

Frequency (MHz)	*75 Ω LP-100*	*75 Ω VNA*	*25 Ω LP-100*	*25 Ω VNA*	*50 Ω LP-100*	*50 Ω 0.5 W +*	*50 Ω VNA*
1.80	1.55	1.55	1.95	2.02	1.06	1.03	1.02
3.50	1.55	1.55	1.97	1.98	1.06	1.03	1.02
7.00	1.56	1.53	1.99	1.94	1.06	1.02	1.02
14.00	1.56	1.48	1.99	1.90	1.07	1.02	1.02
28.00	1.57	1.47	2.03	1.95	1.08	1.05	1.04
56.00	1.54	1.48	2.14	2.20	1.10	—	1.09

Table 5—Parts List

Part ID—Description	*Vendor*	*Part #*
Main Chassis		
Caps—Keycaps	Digi-Key	401-1221-ND
Case—CM6-225	Mouser	616-63075-510-000
C1—1 µF	Digi-Key	P4675-ND
C3, 4, 19—10 µF	Jameco	29891CX
C5-7, 9-13 —0.01 µF	Jameco	15229CL
C4, 8, 14-18—0.1 µF	Jameco	25523CL
D1, 5—1N4148	Digi-Key	1N4148FS-ND
D2—LED	Digi-Key	67-1236-ND
D3—LED	Digi-Key	67-1235-ND
D4—1N4001	Digi-Key	1N4001DICT-ND
JP1-2—Header 2 Pin	Digi-Key	A1911-ND
Jumper	Digi-Key	A26242-ND
J1—16-pin ribbon connector	Digi-Key	A26271-ND
J3—dc Jack	Digi-Key	CP-202B-ND
J2, 7—BNC	Digi-Key	A24495-ND
J4—DB9 D-SUB	Digi-Key	A2100-ND
J5—RJ-11	Digi-Key	A9073-ND
J6—Dual RCA Jack	Digi-Key	CP-1433-ND
L1-3—1 mH choke	Jameco	208240CA
L4, 6, 8—100 µH	Jameco	208194CC
PLED1—PLED Display	CRYSTALFONTZ	CFAP-2002A-Y-ECS
P1—16-pin ribbon cable assembly	Digi-Key	M3AAA-1606J-ND
Q1—2N4401	Digi-Key	2N4401-ND
Q2, 3, 4—2N3904	Jameco	38359CC
R1—1 MΩ,-1/8 W	Digi-Key	P1.0MBACT-ND
R2, 4, 15—10 k 1/8 W	Digi-Key	P10KBACT-ND
R3, 17—390 1/2 W	Digi-Key	P330W-1BK-ND
R6, 9—50 kΩ pot	Digi-Key	3386P-503-ND
R7—330 Ω	Digi-Key	P330BACT-ND
R8—20 kΩ pot	Digi-Key	3386P-203-ND
R11, 12—47 k Ω 1/8 W	Digi-Key	P47KBACT-ND
R13, 14, 16—1kW 1/8 W	Digi-Key	P1.0KBACT-ND
R24, 25—52.3 Ω ±1% 1/4 W	Mouser	271-52.3
R10, 20—121 Ω ±1% 1/4 W	Mouser	271-121
R22, 26, 27—49.9 Ω ±1% 1/4 W	Mouser	271-49.9
RL1—Omron G5V-2-H1-5	Digi-Key	Z108-ND
S1—CEM-1212C piezo transducer	Digi-Key	102-1123-ND
SW1-3—Tactile Switches	Digi-Key	EG1822-ND
U1—16F876	Digi-Key	PIC16F876-20/SP-ND
U2—LM7805T	Jameco	51262CX
U3-4—AD8307	Analog Devices	AD8307AN
U5—LT1634		LT1634-CCZ-2.5
U6—MAX232CPE	Jameco	24811CX
U7—74HC4053	Jameco	272111CX
Y1—20 MHz resonator	Digi-Key	X909-ND
Coupler		
J8, 9—Chassis BNCs	Jameco	159484CL
J10, 11—Chassis SO-239s	Mouser	530-CP-AD206
FT1, 2—Ferrite Cores	Amidon	FT-114A-61
Case—Bud CU-3003A minbox	Digi-Key	377-1090-ND

Table 3 compares the LP-100 with my HP 87510A VNA. The table shows the displayed SWR for nominal load values of 50 Ω, 75 Ω and 25-Ω, using precision terminations. Readings were taken at +20 dBm (100 mW) for all load values, using my HP-8640B signal generator.

Remember, accurately measuring extremely low SWR levels at very low power requires an RF environment with low interference levels. The normal noise floor as indicated on the dBm scale should be between –20 and –10 dBm with no signal applied. This is dependent on a number of factors such as proximity of other gear, or a PC. If it is much higher than that, you may have a strong local broadcast station that is affecting the noise floor. Contact me for a suitable filter design in such a case.

Continuing Development

In an effort to eliminate the parasitic coupling, and provide a repeatable design without the need for tweaking of the transformers, I decided to try a line-section coupler. Instead of the Bird approach of inserting a small loop into a transmission line, I chose to make the line a microstripline sampler using a PC board. The loop was a length of Teflon wire (from the center conductor of RG-142 coax), placed directly on top of a stripline trace. This loop design was settled on after dozens of different configurations were tried by both me and Jack, K8ZOA.

There is one major drawback to line-section couplers, however, and that is that the response varies with frequency at the rate of 6 dB/octave. This meant that I would have to find a method of flattening the response. To level the response, I added a small coil in series with the feed to the log amps. After testing dozens of coils, I settled on a SMD part from CoilCraft. This was necessary to keep the self-resonant frequency of the coil above 54 MHz. Small remaining variations can be easily compensated by calibration.

In an effort to minimize the effects of the connecting lines to the main chassis, I decided to move the log amps to the coupler chassis. I also added a 12-bit A/D converter to the coupler case to increase resolution, as well as a temperature sensor, precision reference chip and RF pickup for the frequency counter.

Fig 11 shows the stripline coupler. The stripline is on the underside for isolation. The cable is an 8-pin DIN cable, which carries power, ground, the SPI bus data to/from the A/D converter and the RF sample for the counter. I am thinking of changing to a VGA video cable, which offers shielded coaxial wires.

This design improves on the already very good SWR accuracy, while eliminating the problems associated with stray coupling in the toroidal design. By moving all RF circuitry to the remote coupler, it should be possible to allow for switching of multiple couplers to allow for VHF and UHF couplers, for instance. I have played a bit with such a design, in fact.

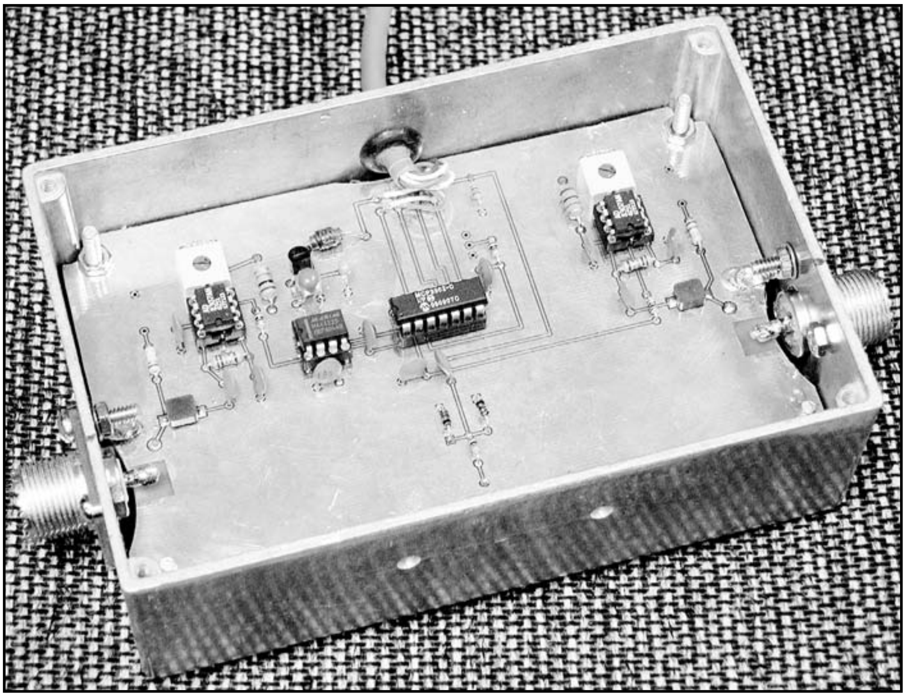


Fig 11—LP-100 stripline coupler. (N8LP)

Because of the need for accurate calibration of the LP-100 before use, I am working on a couple of inexpensive calibrator designs to support the project. For complete details on the stripline coupler design, and for the latest news on the project, check out my website under the LP-100 link. I am also interested in any feedback about feature enhancements that could be implemented in the LP-100.

In closing, I would like to thank my wife, Janet, for her patience and support with all my projects, and to Jack, K8ZOA, for his help and expertise in making this project a success, including design help, machining and testing.

Notes

[1]**http://www.arrl.org/tis/info/pdf/5904024.pdf**

[2]**http://www.arrl.org/members-only/tis/info/pdf/99hb2236.pdf**.

[3]**http://members.tripod.com/michaelgellis/direct.html**.

[4]**http://users.adelphia.net/~n2pk/RLPmtr/RLPv1c.pdf**, also appeared in *QRP Quarterly*, Oct 2002.

[5]**http://www.arrl.org/members-only/tis/info/pdf/9002019.pdf**.

[6]**www.analog.com/UploadedFiles/Data_Sheets/4833689284721AD-8307_b.pdf**.

[7]**http://ww1.microchip.com/downloads/en/DeviceDoc/30292c.pdf**.

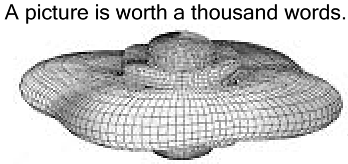
[8]**http://www.basicmicro.com**.

[9]**www.trx-manager.com**.

[10]Oct 2005 *QST,* page 40.

[11]**www.mini-circuits.com**.

Larry Phipps, N8LP, has been licensed since 1965, first as WN8PSD, later as WA9PSD and finally as N8LP, when he upgraded to Amateur Extra. Larry also earned an FCC First Class Radiotelephone license in 1967. He is retired from TelePost Inc, a television post-production facility that he and his wife Janet founded and owned for 16 years. Larry worked for WJBK-TV, Detroit, while attending the University of Michigan College of Engineering. He left school in 1971 to work full-time at WJBK as a broadcast engineer. He then went to NET Television Inc (a subsidiary of WNET-TV, New York), where he was instrumental in launching one of the first computer controlled videotape-editing systems in the country. Larry now devotes time to his house, his wife and ham radio (not necessarily in that order!), all sorely ignored while he and his wife operated the company.



PSpice *for the Masses*

Learn about a circuit simulation program you can download for free.

By Randy Evans, KJ6PO

The purpose of this article is to inform *QEX* readers of a powerful, full featured *PSpice* program that can be downloaded for free from Linear Technology.[1] Unlike many other free design programs available from manufacturers as evaluation versions, this is not a crippled version with limitations. The program, *LTSpice / SwitcherCAD III*, is listed on their Web site as a design program for switching power supplies, which it does exceptionally well. It can also be used to design and analyze any circuit that you would use with *PSpice*. The major limitation is that it lacks the large library of component models available with other large commercial (meaning expensive!) *PSpice* programs. It does come with basic models of all standard components such as resistors, capacitors, inductors, basic NPN/PNP transistor models and some specific types, such as MOSFETs, voltage sources, current sources, etc, and with models of most Linear Technology products. If you want to model an amplifier using a particular transistor such as a 2N5109 or 2N3906, you may have to generate the specific transistor model yourself. That can be done with relative ease if you have the model parameters from the manufacturer. Most transistor/FET circuits can be analyzed using the standard models supplied with *LTSpice*.

The following is a partial list of circuits that I have designed and analyzed using *LTSpice*;

1. Filter designs, including generation of a full set of S-parameters (input/output return loss and frequency response) for low-pass, high-pass, band-pass and notch filters for both

[1]Notes appear on page 20.

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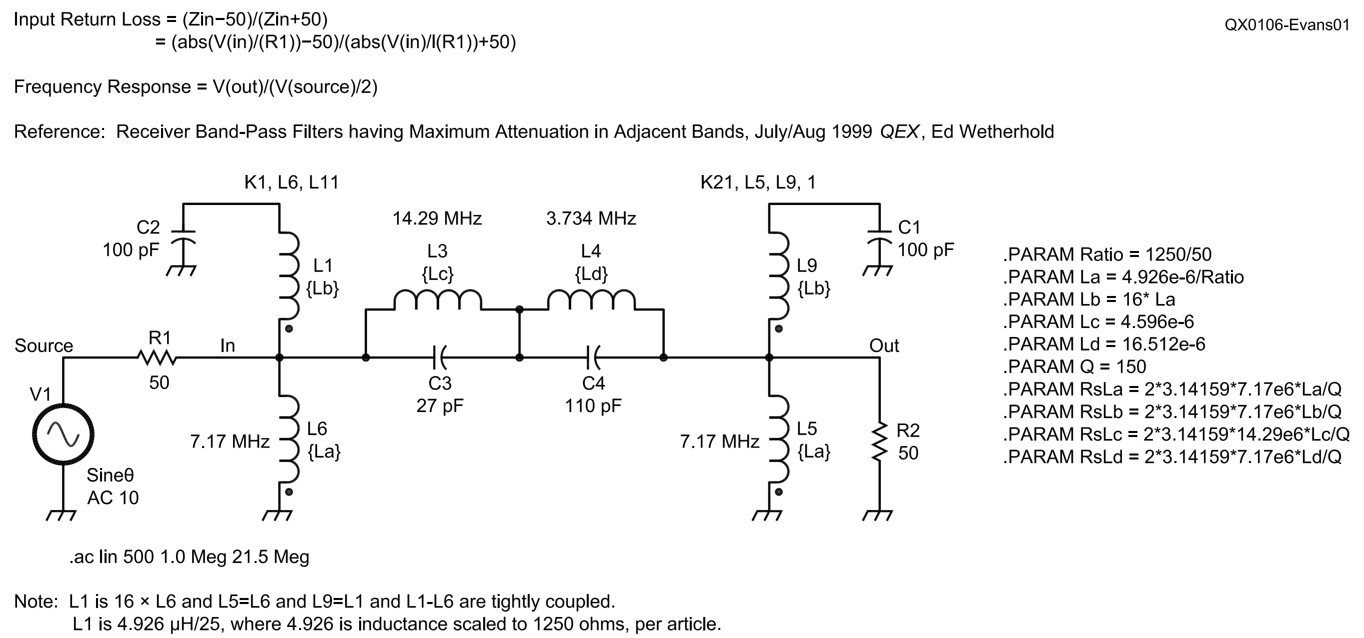


Fig 1—Receiver 40 meter band-pass-filter circuit diagram.

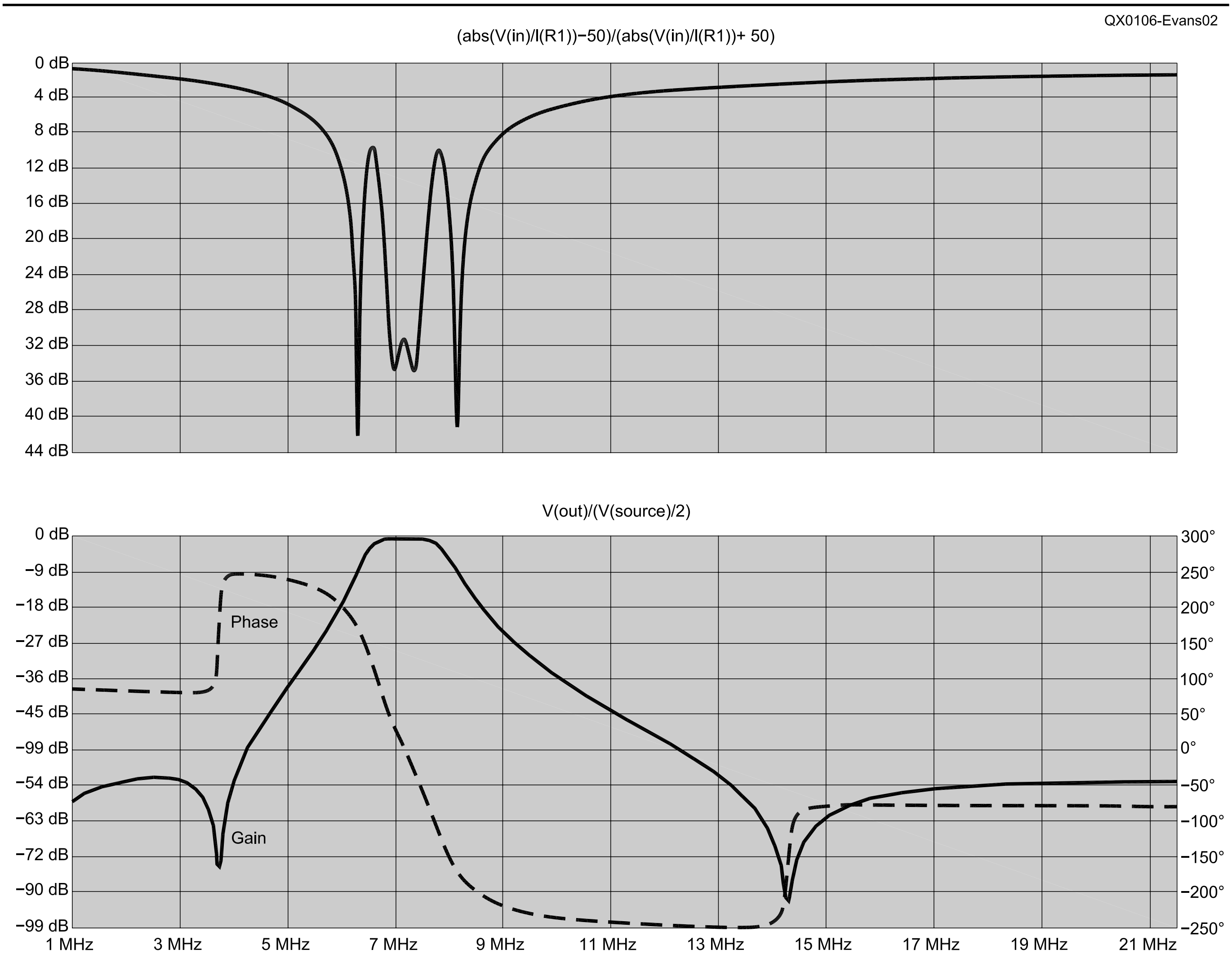


Fig 2—Receiver 40 meter BPF frequency response and input return loss.

Input Parameters
.PARAM Rload = 50 Load Resistance
.PARAM Rs = 800 Source Resistance
.PARAM B = 2e6 Bandwidth
.PARAM Fo = 10e6 Center Frequency
.PARAM QL = 67 Inductor Unloaded Q
.PARAM Cs = 10E-12 Circuit Residual Capacitance

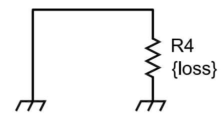
.ac lin 500 8e6 12e6

Waveform Arithemtic: Plot sqrt((v(out)*I(R1))/(v(in)*I(R6))) for frequency response and V(n002)/I(R2) for input impedance vs frequency.

Note: *LT SPICE* only plots in voltage format so sqrt used to convert to power format. R5 and R6 are dummy elements used to calculate the Input power to the circuit since the input and output impedances are different and powers must be used to plot loss verus frequency.

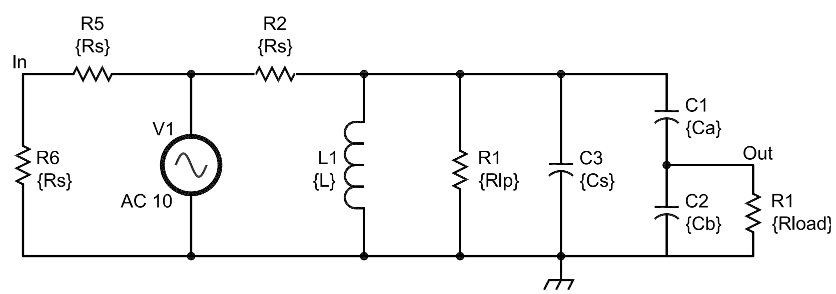
To view component values, go to Control Panel and select Operation tab and check "Generate Expanded Listing", run simulation, go to View menu on task bar, select *SPICE* Error log, and read values on displayed list.
Note: Loss value in dB is shown as value for dummy element R4.

Dummy element used to display Loss value in SPICE Error Log



Reference: *QEX*, Mar/Apr 2004, "Tapped Capacitor Matching Design", Randy Evans KJ6PO

Calculated Parameters
.PARAM Qo = {Fo/B}
.PARAM XL = {Rs/(2*Qo)}
.PARAM w = {2*pi*Fo}
.PARAM Re1 = Rs
.PARAM L = {XL/(2*pi*Fo)}
.PARAM RLp = {QL*XL}
.PARAM Xce1 = XL
.PARAM Ce1 = {1/(L*w**2)}
.PARAM Ce2 = {Ce1 − Cs}
.PARAM Re2 = {Re1*RLp/(RLp − Re1)}
.PARAM QpA = {Re2/(1/(w*Ce2))}
.PARAM Re3 = {Re2/(1 + QpA**2}
.PARAM Ce3 = {Ce2*(QpA**2 + 1)/QpA**2}
.PARAM Cb = {(1/w)*sqrt((Rload − Re3)/((Rload**2)Re3))}
.PARAM QpB = {Rload/(1/(w*Cb))}
.PARAM Ce4 = {Cb*(QpB**2 + 1)/QpB**2}
.PARAM Ca = {Ce4*Ce3/(Ce4 − Ce3)}
.PARAM Loss = {10*log10(Rs/Re2)}



R5 and R6 are dummy elements added to facilitate plotting input/ output power ratios. The elements do not affect the circuit operation.

QX0106-Evans03

Fig 3—Tapped capacitor matching circuit schematic.

active and passive filters.
2. Matching circuits, including the generation of S-parameters.
3. Amplifiers, including the generation of S-parameters and IMD product analysis and dc biasing conditions.
4. Oscillators, including phase shift oscillators, Clapp and Colpitts oscillators.

LTSpice has a help menu in the program and a Users Manual that you can download, as well as numerous switching converter examples, which help immensely in using the program. The documentation assumes the user is familiar with *PSpice*. For those readers not familiar with *PSpice*, a good reference book is suggested. Personally, I have found several good reference books that can be used to understand *PSpice*. I am sure that there are many other good references available as well. Just as importantly, there is a Users Group on Yahoo. Any user can register and access the group

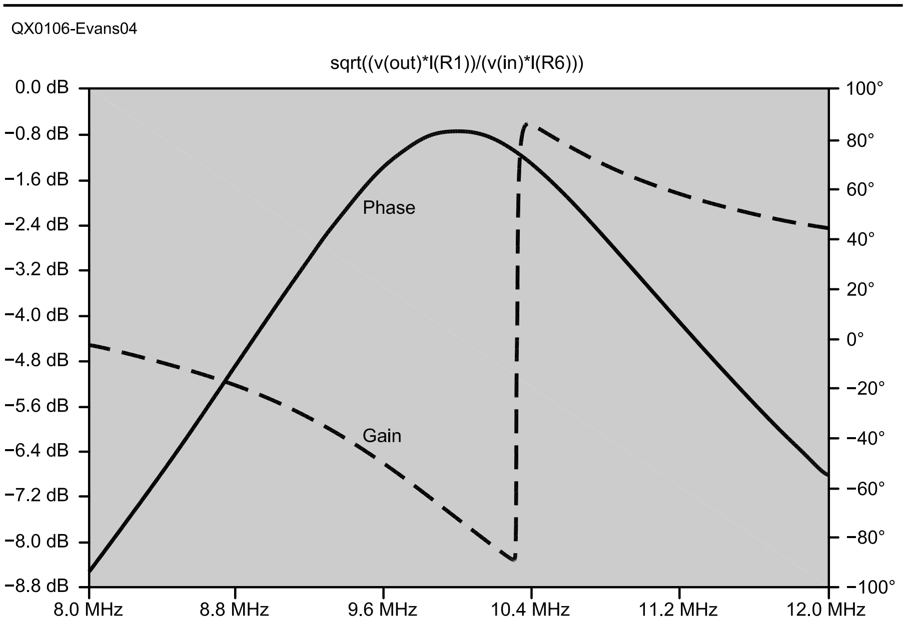


Fig 4—Frequency response of the tapped capacitor matching circuit.

for specific help on a multitude of topics as well as to post questions that very knowledgeable members of the group will answer.[2]

This article is not meant to be a tutorial on the use of *LTSpice* but it will help to familiarize a user about how to analyze basic circuits such as those discussed. First, I would like to present some key features of the *LTSpice* program.

1. Circuits may be input using the built-in general purpose schematic capture capability of the program. There is no need to use the cryptic text descriptions of a circuit so common with simple SPICE programs. Note: the program may be used as a general purpose schematic capture program independent of the *PSpice* capabilities if desired.
2. Both analog and digital circuits may be analyzed using *LTSpice*.
3. Users may do dc analysis, transient analysis, ac analysis (including FFTs for IMD analysis), dc transfer function, and dc operating point analysis on any circuit.
4. Users may generate their own schematic symbols and *PSpice* models as required for those components not in the library.

With that said, let me present some basic circuits for design and analysis. The circuits were chosen to show a variety of techniques to circuit analysis using *PSpice*.

Band-Pass Filter Analysis

The first circuit analyzed is from a *QEX* article by Ed Wetherhold, W3NQN, for receiver band-pass filters.[3] The schematic of the 20-meter BPF circuit is shown in Fig 1. Note that the values for L1, L6, L5, and L9 are shown in braces as {Lb} and {La}. The values for L3 and L4 are shown in braces as {Lc} and {Ld}. This is so that the values for the inductors can be calculated or defined using the *PSpice* directive "PARAM." In the upper right hand section of the schematic are PARAM statements for RATIO, La, Lb, Lc, Ld and Q. This allows the user to specify calculated or defined values for components in one place rather than defining them separately for each component. This can be very powerful for some applications, as will be described later. The PARAM Ratio 1250/50 defines the impedance ratio in the circuit as described in the article and the PARAM statement La = 4.926E–6/ratio defines the inductance in the single-turn winding, again as described in the article. The PARAM statement Lb = 16*La defines the inductance of the remaining four windings relative to the single winding, since inductance increases as the square of the turns ratio.

In addition, a "SPICE" directive is used to define that inductors L6 and L1 are tightly coupled with no leakage inductance by the statement "K1 L6 L1 1." This tight coupling is appropriate for toroid transformers as used in the article, but would not be appropriate for RF coil transformers. Note that the phasing dots for the inductors must be correct. The two band traps L3-C3 and L4-C4 are shown as described in the article. The Q for each inductor is described globally using a "PARAM" directive and the resulting series equivalent resistance is then calculated for each inductor using "PARAM" directives. The model for each inductor then defines the inductance value and equivalent series resistance by the PARAM statements.

For purposes of analysis, the source generator is shown as a sine-wave voltage source in series with a 50 Ω resistor, R1, and the load impedance R2 is shown as a 50 Ω load.

The resultant plots are shown in Fig 2 for both the frequency response and the input return loss. Note that the plot for frequency response uses the equation "V(out)/(V(source)/2)." If Vout alone were plotted, you would see over 6 dB of loss in the frequency response, since *LTSpice* would plot Vsource versus Vout as the default. Vsource is twice the voltage of Vin (assuming a perfect 50 Ω input impedance) since half the voltage is dropped across Rin. Vin varies greatly due to the changing input impedance of the filter input versus frequency, however, so erroneous results would be obtained if Vout were plotted versus Vin.

The input impedance is plotted as "(abs(V(in)/I(R1))–50)/(abs(V(in)/I(R1))+50)." This is from the classic equation for return loss = (Zin –Zo)/(Zin +Zo). The "abs" terms are used to obtain the magnitude of the input impedance, Zin. Zin, of course is obtained from the values of the magnitude of the input voltage divided by the input current versus frequency. The circuit simulation results were not experimentally validated by the author but it does agree with the measured results published in the referenced article by Ed Wetherhold.

Tapped Capacitor Matching Circuit Design and Analysis

The tapped capacitor matching circuit is derived from my previous *QEX* article.[4] The general circuit diagram is shown in Fig 3.

This circuit contains several unique properties necessary to do a circuit design and analysis for this circuit. First is the use of dummy elements R5 and R6 to facilitate plotting input/output power ratios. Since this is a matching circuit, the input and output impedances are going to be different by definition. *LTSpice*, however, only calculates voltages or currents, not power. Since I wanted to plot the power frequency response and *LTSpice* only plots voltages or currents, I had to plot the input/output power ratio using the equation "sqrt((v(out)*I(R1))/(v(in)*I(R6)))" to

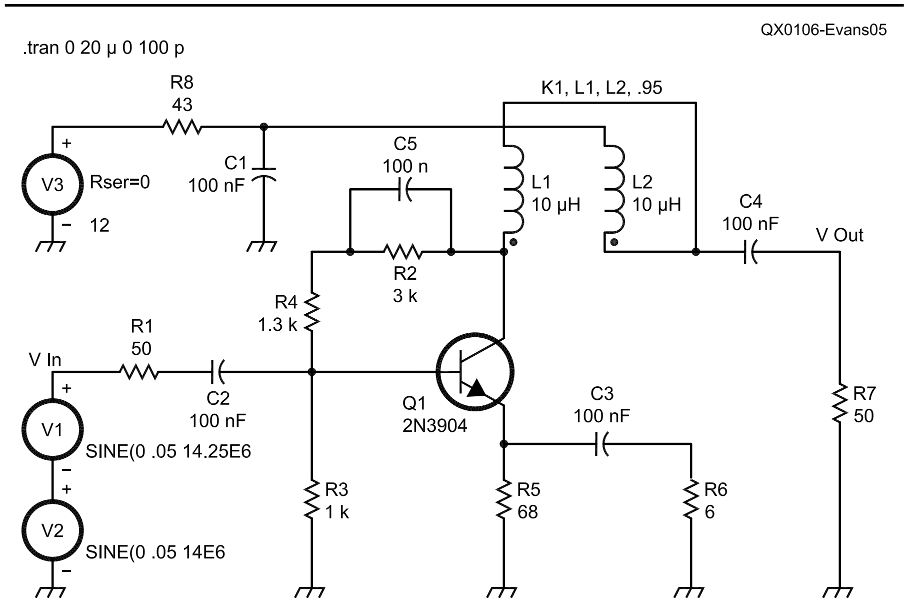


Fig 5—RF amplifier circuit.

convert the power ratio into a voltage ratio. I hope that made sense.

The input parameters for designing a tapped capacitor matching circuit are defined using "PARAM" statements in the Input Parameters section on the circuit diagram shown in Fig 3. The input parameters define the input impedance, the output impedance, the bandwidth and center frequency of the matching circuit, the inductor unloaded Q, and circuit residual capacitance. The circuit parameters are then calculated when the user hits the "RUN" icon on the menu bar. At this point the program plots the frequency response of the matching circuit. Unfortunately, *LTSpice* does not have provisions for displaying the calculated component values directly so in order to view the component values, go to the *LTSpice* Control Panel and select the "Operation" tab and check "Generate Expanded Listing" prior to running the simulation. Then run the simulation, go to the "View" menu on the task bar, select "SPICE Error Log," and read the values on the displayed list.

Note that the dummy element R4 is used to display the center frequency loss value in the SPICE Error Log. It has nothing to do with the circuit itself but it is the only way to get *LTSpice* to display the calculated Loss value in the Error Log. Of course the loss can be seen from the frequency response plot as well.

Fig 4 shows the plotted frequency response of the matching circuit. The plot, as described above, takes into account the different impedances of the circuit input and output. The circuit was experimentally validated as explained in the referenced article, and the simulation agrees almost perfectly with the actual circuit measurements.

Amplifier IMD Analysis

This circuit is a modification of an RF amplifier circuit I received from Wes Hayward, W7ZOI, that he had analyzed on another version of *PSpice*. A variant of his circuit is shown in Fig 5 and uses feedback around a 2N3904 transistor to improve its linearity. In this case, the circuit will be used to demonstrate how to do an IMD analysis of an RF amplifier. Note that the input to the RF amplifier consists of two 50 mVpeak sine-wave voltage sources, one at 14.0 MHz and the other a 14.25 MHz, in series with a 50 Ω resistor. Because half of the input voltage will be dropped across the 50 Ω series resistor (for a 50 Ω input impedance), the actual circuit input voltage for each sine wave is 25 mVpeak (17.8 mVrms/tone or –22.0 dBm/tone). The output is into a 50 Ω load resistor. The RF transformer is modeled as two 10 µH inductors tightly—but not perfectly—coupled using the *PSpice* statement "K1 L1 L2 .95."

When the "RUN" button is selected on the task bar, a time domain simulation is started, showing the steady state condition. (I made sure that the option "Start external dc supply voltages at 0 V" was not selected since I am not interested in the start-up transient of the amplifier for this IMD analysis.) Once the time domain plot is finished, as shown in Fig 6A, select the plot pane and then go to "VIEW" and select "FFT."

Because the IMD analysis uses two sine-wave input signals 250 kHz apart (14.00 and 14.25 MHz), and we want to have an exact integer number of cycles in the time domain waveform

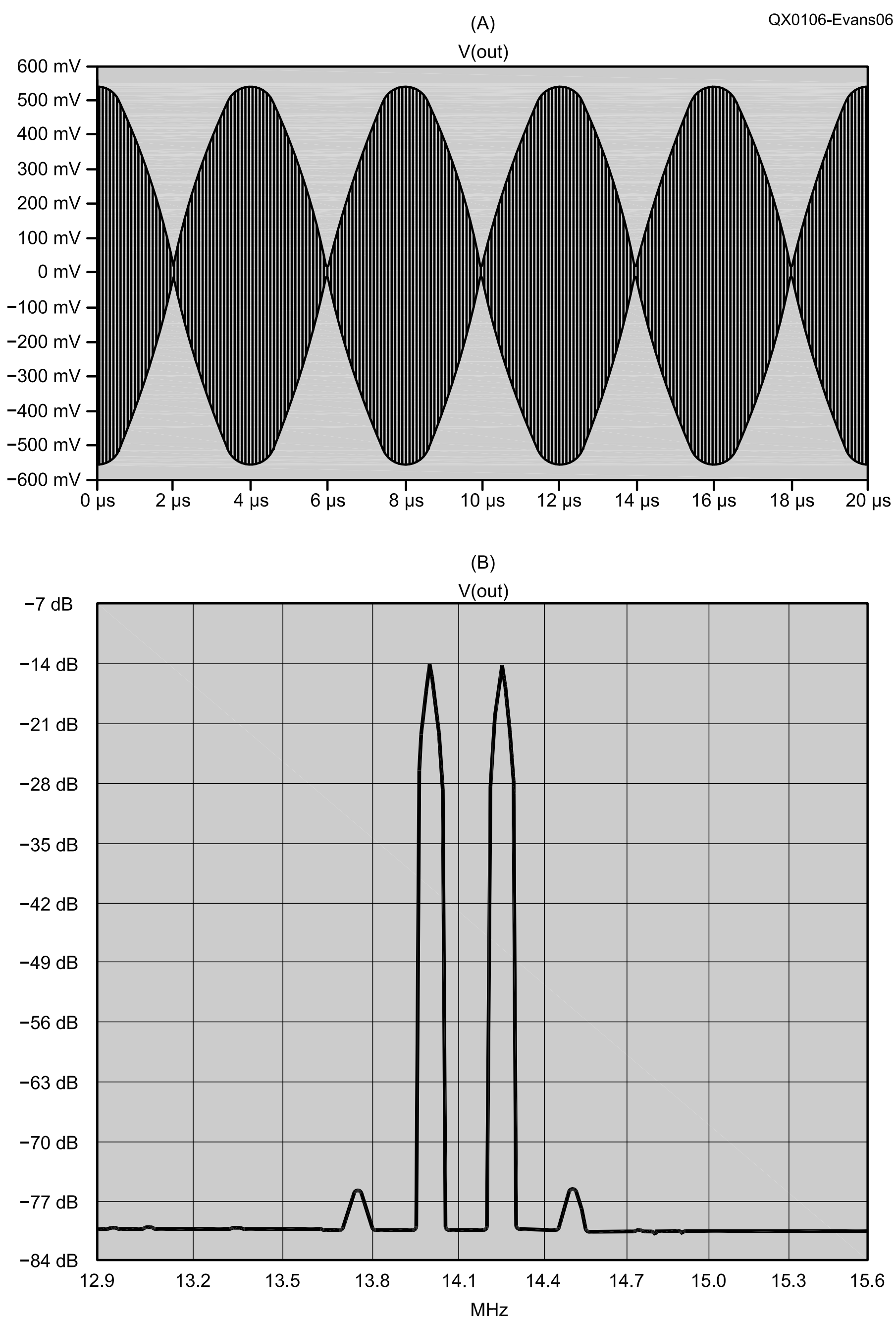


Fig 6—Output time simulation and expanded FFT of RF amplifier output.

to prevent sampling artifacts, I used a multiple of 4 µs for the time duration of the time domain simulation. In this case, I used a 20 µs window with a time step of 100 ps.

I used an FFT of 16,384 points and specified a time range of 4 to 20 µs with 1 point for binomial smoothing and no windowing (not needed with an integer number of cycles). You can play around with different settings to see the effects. After the FFT plot is run I right-clicked on the FFT plot and put a box around the two input signals. This expanded this area to generate the plot shown in Fig 6B. Note that the FFT shows the rms voltage spectral components in dB relative to a 1 Vrms level. The two output signals are at 14.0 and 14.25 MHz, each shown at –14.3 dB below 1 Vrms (192.5 mVrms, or Pout of –1.3 dBm) and the two third-order intermod products at 13.75 MHz and 14.5 MHz are at approximately –78.6 dB below 1 Vrms (117.5 µVrms or –65.6 dBm) for an intermodulation dynamic range (IMDR) of 64.3 dB. This equates to a gain of 20.7 dB (17.8 mVrms/tone input and 193 mVrms/tone output) and a third order output intercept IP3out of +30.9 dBm using the equation IP3out = IMDR/2 + Pout.[5]

I then built the circuit exactly as shown in the schematic using a 2N3904 transistor and measured the amplifier gain and third order intercept. The measured gain was 20.7 dB, exactly as simulated, and the third order intercept was measured at 30.4 dBm, or 0.5 dB lower than predicted. This represents a very powerful tool for accurately predicting the IMD of an RF amplifier design.

Colpitts Oscillator

The circuit shown in Fig 7 is a Colpitts oscillator using a U309 JFET. This circuit is designed to oscillate at about 5 MHz and has an output of 14 Vpk-pk as shown in the transient analysis plot in Fig 8. In this case the SPICE directive ".options method = trap" is invoked. This is a simulator option to set the integration method to "Trapezoidal" for better convergence, which I have found generally works better than the alternative "Gear" method for oscillator circuits.

Crystal Filter Design and Analysis

The last circuit presented is based on a design using the software program XLAD included with the book *Experimental Methods in RF Design* published by ARRL. The crystal parameters were measured using techniques in the author's *QEX* article, "Ladder Crystal-Filter Design".[6] A five-pole Butterworth ladder filter with a 250 Hz bandwidth is designed per XLAD using crystals with a series resonance frequency of 5.998051 MHz as described in the article, on page 43. The crystal parameters measured were a series resistance of 12.5 Ω, a crystal parallel capacitance of 4.46 pF, an unloaded Q of 111000, a crystal motional capacitance of 20.89 femtofarads (20.89×10^{-15} F), and a crystal motional inductance of 33.7 mH. The end terminations to be used are 400 Ω.

The resulting circuit is shown in Fig 9. The circuit's frequency response is shown in Fig 10.

Summary

I hope this article has given an inkling of what *LTSpice* can do. The article has only scratched the surface of the program capabilities. Since it has no limitations on circuit size, the user could simulate quite large and complex circuits if needed. For those readers interested in pursuing *LTSpice*, I encourage them to join the Yahoo Users Group to learn many different approaches to circuit design and analysis documented in the files and messages in the User Group.

I would particularly like to thank Wes Hayward, W7ZOI, for help with the RF amplifier IMD analysis and for assistance in using the crystal filter design program XLAD.

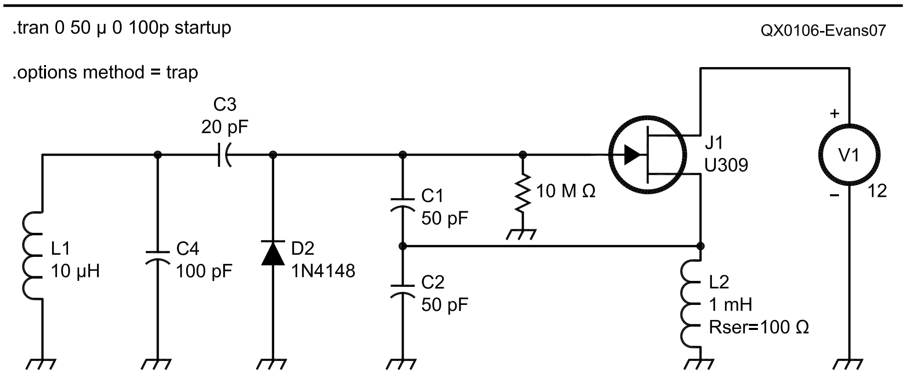


Fig 7—Colpitts oscillator circuit.

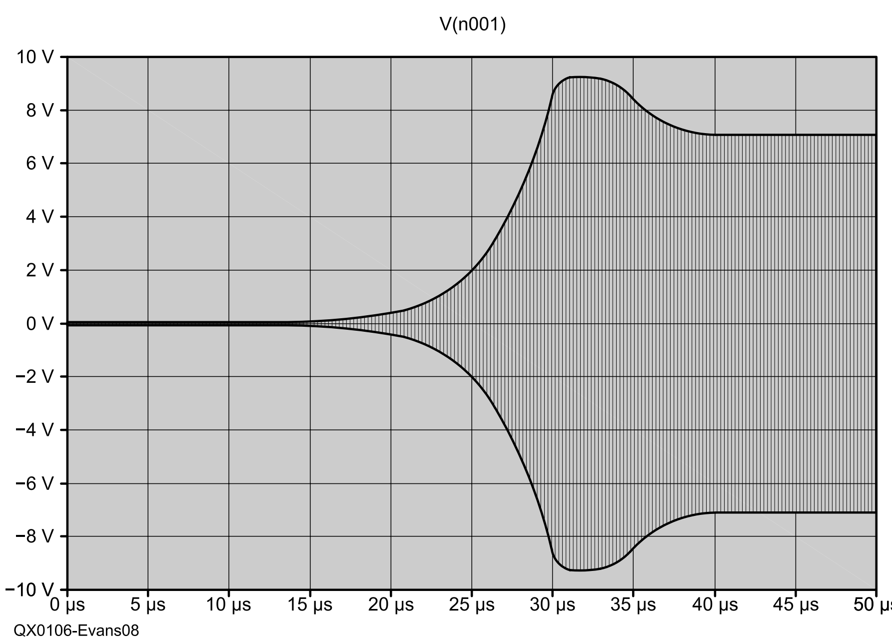


Fig 8—Colpitts oscillator output.

The design files for the circuits analyzed with *LTSpice* in the article are available in the *QEX* file section.[7]

Notes

[1]**www.linear.com/software/**.

[2]**http://groups.yahoo.com/group/LTSpice-SwitcherCAD/**.

[3]E. Wetherhold, W3NQN, "Receiver Band-Pass Filters Having Maximum Attenuation in Adjacent Bands," *QEX*, Jul/Aug 1999, pp 27-33.

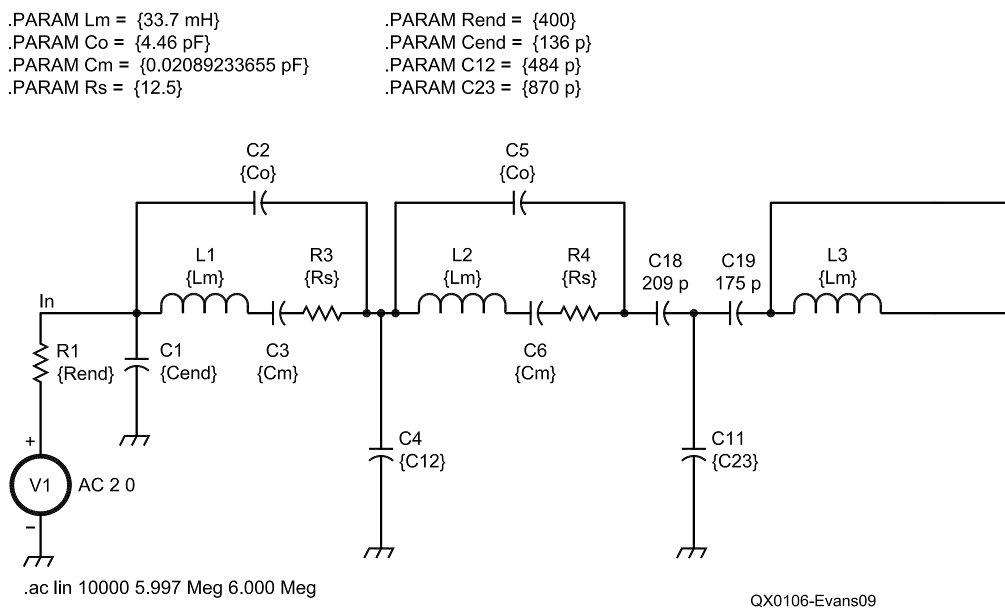
[4]R.Evans, KJ6PO, "Tapped-Capacitor Matching Design," *QEX*, Mar/Apr 2004, pp 46-52.

[5]W. Hayward, W7ZOI, R. Campbell, KK7B, B. Larkin, W7PUA, *Experimental Methods in RF Design*, ARRL, Equation 2.38, p 2.22.

[6]R. Evans, KJ6PO, "Crystal Parameter Measurement and Ladder Crystal-Filter Design," *QEX*, Sep/Oct 2003, pp 38-43.

[7]You can download the *PSpice* circuit files analyzed in this article from the ARRL Web at **www.arrl.org/qexfiles/**. Look for 1X06Evans.zip.

Randy Evans was first licensed as K3VFE in 1960 after a year with a Novice license. After his license expired while in Vietnam, he dropped out of Amateur Radio until his son got interested in Amateur Radio. So, in 1998 he became licensed again as KJ6PO. He holds a BSEE degree (1967) from Lehigh University in Bethlehem, PA and an MBA from the University of Utah in Salt Lake City, UT. He has been employed in the communications engineering field for almost 38 years, working from the telephone to HF/VHF and with satellite communications at UHF, SHF and EHF bands. He has also worked with terrestrial and tropo microwave communications. Most recently he has worked with optical communications systems at 40 Gbps rates, with up to 3.2 terrabits per second on a single fiber. He holds four patents in the communications area. His Amateur Radio interests are mainly in the experimentation areas, especially in advanced radio development using DSP techniques.



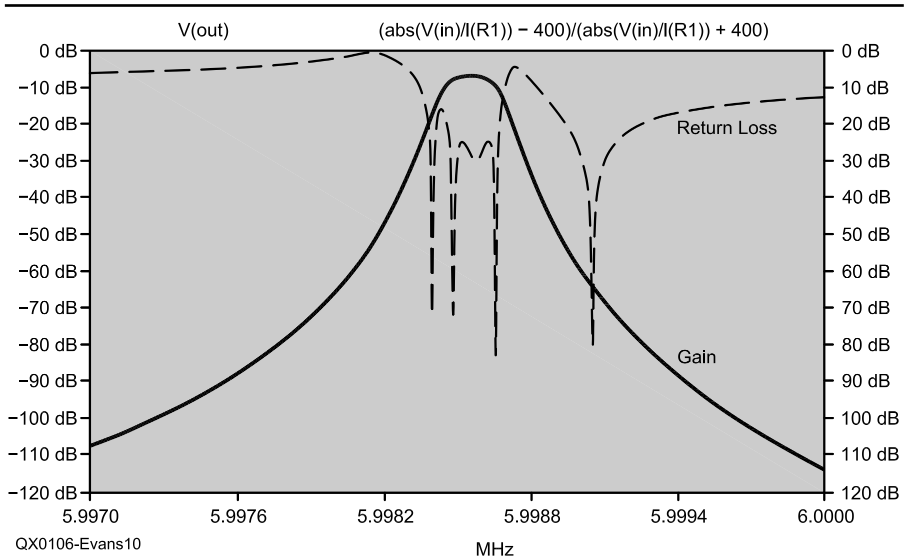


Fig 10—Four-pole crystal-filter frequency response.

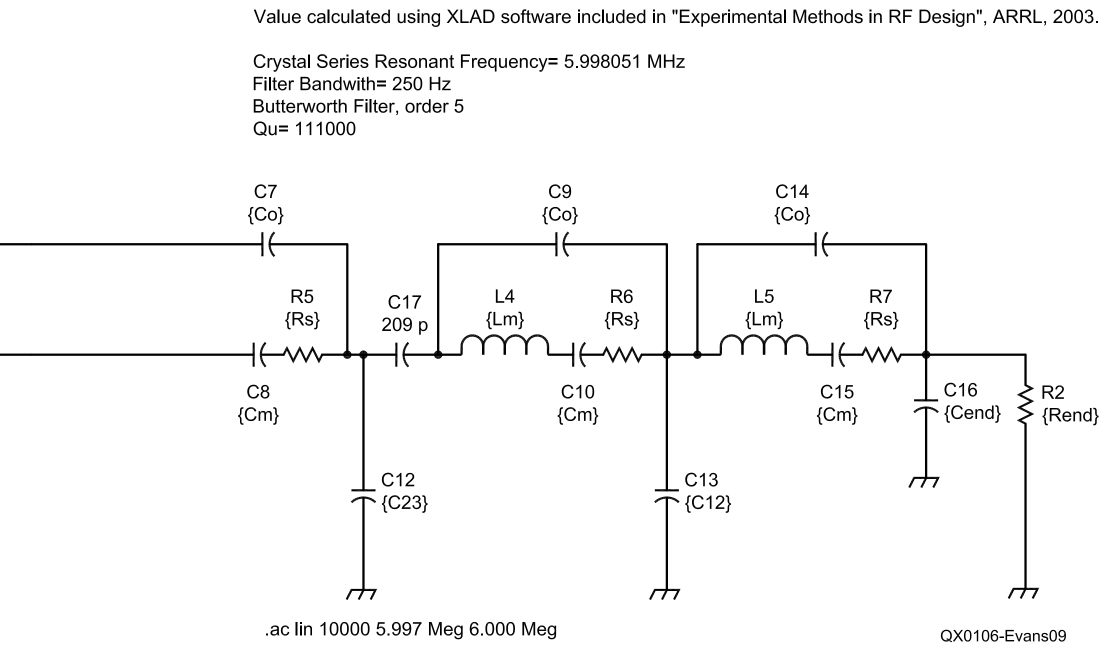


Fig 9—Five-pole 250 Hz bandwidth Butterworth ladder crystal filter.

Upcoming Conferences

10th Annual Southeastern VHF Society Conference, Call for Papers

The Southeastern VHF Society is calling for the submission of papers and presentations for the upcoming 10th Annual Southeastern VHF Society Conference to be held in Greenville, South Carolina on April 28 and 29, 2006. Papers and presentations are solicited on both the technical and operational aspects of VHF, UHF and microwave weak signal Amateur Radio. Some suggested areas of interest are: Transmitters; Receivers; Transverters; RF Power Amplifiers; RF Low Noise Preamplifiers; Antennas; Construction Projects; Test; Equipment and Station Accessories; Station Design and Construction; Contesting; Roving; DXpeditions; EME; Propagation (Sporadic E, Meteor Scatter, Troposphere Ducting, etc); Digital Modes (WSJT, etc); Digital Signal Processing (DSP); Software Defined Radio (SDR), Amateur Satellites, and Amateur Television.

In general, papers and presentations on non weak signal related topics, such as FM repeaters and packet, will not be accepted but exceptions may be made if the topic is related to weak signal. For example, a paper or presentation on the use of APRS to track rovers during contests would be considered.

The deadline for the submission of papers and presentations is **March 3, 2006**. All submissions should be in Microsoft Word (.doc) or alternatively Adobe Acrobat (.pdf) files. Pages are 8½ by 11 inches with a 1-inch margin on the bottom and ¾-inch margin on the other three sides. All text, drawings, photos, etc, must be black and white only (no color). Please indicate when you submit your paper or presentation if you plan to attend the conference and present there or if you are submitting just for publication. Papers and presentations will be published in bound proceedings by the ARRL. Send all questions, comments and submissions to the technical program chair, Jim Worsham, W4KXY, at **w4kxy@bellsouth.net**. □□

Quantifying SETI

A state of the search report.

By Dr H. Paul Shuch, N6TX
Executive Director, The SETI League, Inc

It should come as no surprise to readers of *QEX* that the scientific Search for Extra-Terrestrial Intelligence (SETI, sometimes called the search for the ultimate DX) was initiated by, and is still largely conducted by, radio amateurs. The world's hams instinctively understand the nature of electromagnetic communications, as well as the nature of the "free space" (or "aether") that fills the void between the stars, and forms a transmission medium for our favored photons. If asked to identify one radio amateur whose contributions to SETI were the most significant, I would not hesitate to name the late Dr Philip Morrison, W8FIS, the father of modern SETI science.[1]

Institute Professor Emeritus of Physics and Astronomy at the Massachusetts Institute of Technology, Phil Morrison was a distinguished theoretical astrophysicist, and a pioneer in the search for extraterrestrial intelligence through radio communication. He authored scores of books, produced countless television documentaries and lectured tirelessly around the world, despite the physical limitations imposed upon him by post-polio syndrome. Phil Morrison passed away quietly at his modest home in Cambridge, Massachusetts, in April, 2005 at the age of 89; but not before inspiring a whole generation of scientists (including me) to ask the difficult questions and then attempt to answer them. It is to Phil Morrison's memory that this article is dedicated, and it is with his most important publication that it deals.

Searching for Interstellar Communications

Although speculation about the existence of life on other worlds is as old as the ancients, modern SETI science traces its origins to a single brief article published just under a half century ago in the prestigious British science journal *Nature*.[2] Co-authored by Phil Morrison, then a professor of physics at Cornell University, and his Cornell colleague Giuseppe Cocconi, "Searching for Interstellar Communications" was the first scientific paper to quantify the challenge of signaling between the stars. In just three pages, with only eight very carefully chosen equations, Morrison and Cocconi

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[1]Notes appear on page 30.

sought to summarize all that was known, or could be known, about interstellar signaling using the best available technology of their day. Its four major section headings would form the syllabus for any contemporary course in radio communications: "The Optimum Channel," "Power Demands of the Source," "Signal Location and Bandwidth," and "Nature of the Signal and Possible Sources." We will revisit those same four subjects, plus a few more, in this paper.

These two scholars were probably the first to recognize a paradigm shift, as Earth was just beginning to develop the kinds of technologies that would take a search for alien emissions out of the realm of science fiction and into the scientific mainstream. More importantly, Cocconi and Morrison went so far as to suggest that SETI was a subject worthy not just of speculation and debate, but of serious observational study. Their concluding words are as valid today as they were in 1959: "We therefore feel that a discriminating search for signals deserves a considerable effort. The probability of success is difficult to estimate, but if we never search, the chance of success is zero."

Fifty Years of Solitude

In the wake of Cocconi's and Morrison's seminal article, we have indeed invested that considerable effort in a much deserved search. The first modern SETI Experiment, Frank Drake's 1960 *Project Ozma*,[3] was, in fact, in the construction stages even as the *Nature* article went to press. That Drake's assumptions and strategies closely paralleled Morrison's and Cocconi's, even though neither group knew anything about the other's work, is an indication of what I like to call the Parenthood Principle: When a great idea is getting ready to be born, it sets out in search of a parent. Sometimes, it finds more than one.

To his credit, Drake chose for his two target stars a couple of Morrison's and Cocconi's favorite candidates. He conducted his search at the very frequency they proposed. To his consternation, Drake failed to detect our cosmic companions. Nobody said this was going to be easy.

For 47 years now, various government agencies, educational institutions, non-profit entities and, yes, even Amateur Radio clubs,[4] have conducted hundreds of searches, from dozens of countries, over literally millions of frequencies, in all conceivable directions across 4π steradians (the SI measure of angular area) of space and time. They have all been precisely as successful as Drake's *Project Ozma*: To date, not *one single* verified emission of intelligent extraterrestrial origin has ever been observed and independently confirmed. To SETI's critics, it begins to seem as though even as we do search, the chance of success is zero.

Might it, in fact, be time to reexamine Cocconi's and Morrison's assumptions? In this article, we audaciously deconstruct the most important paper in the history of SETI science to update its methodology, bringing it in line with 21st-century technology.

The Optimum Channel

Have you ever tried to work a DXpedition that didn't announce its frequencies in advance? "We'll be on the air next Wednesday," they might have advertised in *QST*, "on some ham band or other." So, you flip the band switch until you find one that is open, tune around until you hear a pileup and start calling. Sometimes, you get lucky.

Only do not expect to get lucky when that rare DX is ETI itself. First off, how do you know which bands are open? There is not likely to be a pileup lighting your way. And you do not even know for certain whether the DXpedition actually exists, much less whether it made it to that particular island in the interstellar sea.

You can improve the odds by choosing to monitor a portion of the electromagnetic spectrum where the band *might* be open. Morrison and Cocconi chose a different approach: identifying, and then excluding from the search space, those bands that were known *not* to be open to signals from beyond Earth: "Radio frequencies below ~1 Mc./s., and all frequencies higher than molecular absorption lines near 30,000 Mc./s.,…are suspect of absorption in planetary atmospheres." So, we can eliminate the very low- and very high-frequency ends of the electromagnetic spectrum. That still leaves a lot of band to scan!

Even if we guess right as to where the signal might appear on the dial, it still needs to override the background noise if we are to detect it on Earth. In their paper, Morrison and Cocconi quantified the most likely source of cosmic interference: the emission spectrum of the galactic continuum. Its known characteristics, which have not changed in the years since their article was published, allowed them to compute a frequency range with a minimum of spurious background. That broad interstellar communications band is centered around 10 GHz.

Okay, so we will search the microwave spectrum. But where, exactly? That broad noise minimum centered on 10 GHz is still more than a decade wide. That is to say, anywhere from 1 GHz to 30 GHz is fair game, if background noise is our primary consideration. Quoting from the original SETI article again, "A long spectrum search for a weak signal of unknown frequency is difficult." We need to narrow the search area. Earth's first SETIzens hoped there'd be a cosmic band plan, with a well publicized inter-species calling frequency. In "Searching for Interstellar Communications," one such frequency was proposed: "…just in the most favoured radio region there lives a unique, objective standard of frequency, which would be known to every observer in the universe: the outstanding radio emission line at 1,420 Mc./s. (λ = 21 cm.) of neutral hydrogen."

First observed from Earth in 1951, the hydrogen emission line[5] is indeed a cosmic calibrator. Hydrogen is, after all, the most abundant element in interstellar space. There's something like one hydrogen atom per cubic centimeter filling the black void between the stars. And hydrogen atoms chirp occasionally at the precisely known frequency cited above. Whereas the weak chirp of a lone cricket in an otherwise empty field might well go unnoticed, add the chirps of millions of its neighbors and the field resounds with a strong and healthy chorus. Point your antenna at the empty depths above, tune your microwave receiver to the hydrogen line, and the resulting audio is rather like unsquelching an FM handheld transceiver on an unused channel. Here, reasoned Morrison and Cocconi, is nature's crystal calibrator, marked out in hydrogen chalk for all to see.

Why would ETI choose to transmit on the hydrogen emission line? Morrison and Cocconi again: "It is reasonable to expect that sensitive receivers for this frequency will be made at an early stage of the development of radio astronomy. That would be the expectation of the operators of the assumed source, and the present state of terrestrial instruments indeed justifies the expectation. Therefore, we think it most promising to search in the neighbourhood of 1,420 Mc./s."

Note that they wrote "in the neighbourhood of" the hydrogen line. Of course, you wouldn't expect to hear ETI calling precisely on hydrogen's emission frequency. The hydrogen noise would drown out the signal. But tune around the band in that vicinity, they proposed. If hydrogen noise is present, ETI can't be far away.

That logic held so well in 1959 that, for his *Project Ozma* search, Frank Drake had already independently decided to tune his down converter-equipped Hallicrafters across a band in hydrogen's general vicinity. It seemed like a good choice then. It still does today. If you must select but a single frequency on which to conduct a search for intelligently generated signals from an alien species, then 1420 MHz is as good a guess as any.

In the years A.D. (after Drake), SETI scientists have proposed numerous other such "magic frequencies." Several still seem like fair game and in fact, spot-frequency searches still go on. But in recent decades, a technological breakthrough has occurred which may well negate this kind of reasoning: the development of real-time multi-channel spectrum analyzers[6] (MCSAs). Whereas Drake's receiver, like most ham receivers, could only tune one narrow slice of spectrum at a time, today it is possible to monitor millions of channels at once. And that breakthrough has begun to change the way the SETI game is played.

What if instead of laboriously monitoring frequencies one by one, we could concoct the ultimate panadapter, capable of viewing the entire microwave spectrum, say, 1-100 GHz, simultaneously, in real time? The analytical tool that makes MCSAs possible is the fast Fourier transform (FFT), along with the powerful and affordable microcomputers on which today it can be run. Limited only by computer power, which, as Gordon E. Moore[7] reminds us, keeps doubling every year or two, we can now apply complex digital signal processing (DSP) techniques to the challenge of monitoring ever more channels across an increasingly wide spectrum, quickly approaching that lofty goal.

Today, even modest amateur SETI stations routinely monitor tens of thousands of channels, spread across tens to hundreds of kHz of the microwave spectrum. Our professional counterparts, with their presumed greater budgets and related resources, have expanded their searches to many tens of MHz at a time, divided into millions of DSP bins. (Our ambitious goal of monitoring tens of GHz of frequency span in real time still eludes us, but achieving that objective is just a matter of time.) So, guessing right about ETI's calling frequency is becoming ever less important, and in time the significance of Cocconi's and Morrison's channel recommendation may fade into obscurity. Still, if I had to choose just one frequency on which to conduct SETI….

Power Demands of the Source

From listening for signals buried in QRM and QRN, we hams know that successful communication is achieved not just by maximizing signal amplitude, but rather by maximizing signal-to-noise ratio (SNR). To determine the signal amplitude required for interstellar communication, Cocconi and Morrison first computed the amplitude of the galactic background around the 21-cm hydrogen line. Their calculations, which have weathered the test of time, quantified the interference level across two-thirds of the sky. The authors noted, "Near the plane of the galaxy there is a background up to forty times higher." Fortunately, as viewed from Earth, there are promising target stars in all directions, so it was deemed possible to minimize QRN: "It is thus economical to examine first those nearby stars which are in directions far from the galactic plane."

"Searching for Interstellar Communications" introduced an equation for assessing the transmitter power required for overcoming the cosmic background radiation, assuming transmit and receive antennas of known and equal size. In 1959, the largest parabolic reflector on Earth was the Jodrell Bank 80-meter reflector.[8] Given two such antennas separated by 10 light years, Cocconi and Morrison computed a required transmitter power "...which would tax our present technical possibilities." They then cited a planned Naval Research Laboratory antenna of 200 meters diameter, noting that between two such dishes, "The power needed is a factor of 40 lower, which would fall within even our limited capabilities." This was true even with the very crude microwave receivers of the day, and even lacking the DSP capabilities that we now enjoy.

Just a decade later, with the completion of the 305-meter Arecibo radio telescope,[9] interstellar communication over a 10-LY path, using technology no more advanced than Earth already possessed, became entirely feasible, validating Cocconi's and Morrison's claim, "We can then hope to see a beam toward us from any suitable star within some tens of light years."

But that was then. What of now? Arecibo is still our largest radio telescope, although significantly larger capture areas are contemplated through presently planned arrays of thousands of modest antennas. But thanks to the solid-state revolution, spurred by the needs of our terrestrial telecommunications infrastructure, receiver noise temperatures have decreased from thousands to mere tens of kelvins. A combination of improved frequency stability and advanced DSP techniques has reduced our channel bandwidths—and with them, our corresponding receiver noise—from tens of kHz to mere tenths of Hz. Today, we have available microwave power amplifiers putting out megawatts of RF. Considering the very best technology extant on Earth, I compute that two Arecibos can now communicate with each other not merely over tens of LY, but rather over tens of *thousands* of LY.

Different SETI scientists come up with different solutions to the very same equations, depending upon their underlying assumptions. Frank Drake himself (onetime Director of the Arecibo Observatory, and thus as knowledgeable about its capabilities as anyone) once computed the communications range between a pair of Arecibos. He determined that they could complete a QSO from anywhere within the Milky Way galaxy,[10] a result which supposes a range on the order of 100,000 LY. My own solution suggests a more modest 30,000 LY range, under the very best of circumstances.[11] Expressed in terms of signal strength, that's about a ten dB discrepancy. Drake responded to my result by stating, "All the parameters used in the Arecibo numerical example are plausible. The point was to show that if one tried hard, one could detect an Arecibo anywhere in the Galaxy."[12] Without belaboring the precise computations, suffice it to say that Drake and I agree on the big picture: Recent advances have it made it entirely possible, using technology no more advanced than that which was possessed on Earth by the end of the 20th Century, to communicate between the stars, over very vast distances indeed.

And yet, despite these advances, and despite the best efforts of some very talented scientists and technologists, SETI success still eludes us. Might there be other factors that we have overlooked?

Signal Location and Bandwidth

In Morrison's and Cocconi's paper, great attention was paid to the expected Doppler shift—and the corresponding difference between transmitted and received frequency—of hydrogen-line signals emanating from planets orbiting their stars. While such considerations were significant in the case of magic frequencies and single-channel receivers, our previously mentioned development of multi-channel spectrum analyzers tends to make the issues moot. But there is another reason to concern ourselves with Doppler shift, and it is not the absolute frequency change, but rather the *rate* of frequency change over time, that is significant.

Consider a deliberately beamed beacon, sent Earthward from a planet orbiting a distant star. It is evident and quantifiable that there will be a change in frequency over time as that signal arrives at Earth's vicinity. That frequency change is dominated by the rotation of the originating planet, but also contains components corresponding to the planet's orbit around its star, as well as its local sun's motion relative to the galactic center. One would hope that a technologically advanced civilization wishing to make its presence known would make the task of detection as easy as possible for us mere adolescents. One of the ways they could do so is to chirp their transmitter's frequency to compensate for the relative motions of their planet and star, with respect to the Galactic Center of Rest. The mathematics of such drift compensation is relatively straightforward.

However, even with Doppler compensation at the transmit end of the path, the frequency of the signal received on Earth will still change over time, because of the Doppler components imposed by the relative motions of our own planet, and our own star. These too are easily computed, and we could in theory chirp our own receiver's local oscillator in compensation for them, resulting in a fixed frequency of reception.

On the other hand, a received signal which varies over time has certain benefits, when one attempts to validate it as being extraterrestrial in origin. Consider that Earth suffers from extreme RF pollution of its own making, from terrestrial and orbital sources of RF. Separating the cosmic wheat from the terrestrial chaff is becoming ever more challenging as we continue to despoil our electromagnetic environment. And the Doppler shift imposed on a received signal by our planet's relative motion is an excellent indicator of its interstellar origin.

Consider an interfering signal emanating from a terrestrial source. That signal was generated on a rotating and revolving planet, but also received on that same rotating and revolving planet. Hence, the relative motion between the points of transmission and reception is zero, and the received signal is stable in frequency over time. In contrast, a signal emanating from a low-Earth-orbiting (LEO) satellite, as received on Earth, exhibits significant Doppler shift, its frequency varying as an S-curve over time. Both cases, that of fixed frequency and that of rapid Doppler, can be readily excluded from further analysis as emissions of human origin.

In between these two extremes, there is the case of a signal with slow and steady Doppler shift, consistent with Earth's motion with respect to the stars. Any signal whose frequency change matches that expected by sidereal motion is a likely candidate for further analysis. Thus, modern SETI experiments attempt to measure the rate of a candidate's frequency change over time to help us in identifying it as being of extraterrestrial origin. This kind of analysis, impractical in SETI's infancy, is light work for today's signal analysis computers.

There are two ways to increase the sensitivity of our receivers when recovering a weak CW source: through decreasing the detector's bandwidth, or through averaging many samples (increasing signal integration time). Once we have solved the Doppler equations for sidereal motion, it is feasible to employ both techniques in parallel. Morrison and Cocconi proposed as much in their 1959 paper: "Of course, the smaller the bandwidth chosen, the weaker the signal which can be detected.... It looks reasonable for a first effort to choose a bandwidth Δf_d normal for 21 cm. practice, but an integration time τ longer than usual. A few settings should cover the frequency range F using an integration time of minutes or hours."

In 1959, IF filtering with LC circuits and hardware integration with RC networks ruled the day, limiting our capabilities with respect to both variables. Today's DSP techniques allow us a wider set of options and permit almost arbitrarily narrow bin widths, as limited only by signal dispersion in the interstellar medium, and equally arbitrary integration times, limited only by the visibility of the source. These flexibilities hold the potential for significantly increasing our receive station's sensitivity, at the expense of adding perhaps more degrees of freedom than we wish to tolerate in the task of SETI signal analysis.

Nature of the Signal and Possible Sources

It's reasonable to expect that any artificially generated signals detected by Earth's SETI projects most likely would have emanated from a radio transmitter on a planet's surface, or on one of a planet's moons. Planets are not particularly easy to detect from Earth. In fact, though we know of 168 extrasolar planets at the time of this writing,[13] it is only within the past decade that we have been able to detect them at all. Detecting planets' moons is even more challenging. However, planets orbit stars that are quite visible, and we have rather strong knowledge about the characteristics of those stars most likely to accommodate potentially habitable planets. So the very first SETI experiments concentrated on identifying likely candidate stars.

"The first effort," wrote Cocconi and Morrison, "should be devoted to examining the closest likely stars. Among the stars within 15 light years, seven have luminosities and lifetimes similar to those of our Sun. Four of these lie in the direction of low background.... There are about a hundred stars of the appropriate luminosity among the stars of known spectral type within some fifty light years. All main-sequence dwarfs between perhaps *G0* and *K9* with visual magnitudes less than about +6 are candidates."

Early SETI concentrated on precisely those stars identified in "Searching for Interstellar Communications." When pursuing his *Project Ozma*, Drake trained a single 85-foot dish on two of those nearby stars specifically identified by Morrison and Cocconi, though at but a single frequency, and for just a few days in April of 1960. Later studies surveyed those two candidates more extensively, along with all hundred of the mentioned promising stars within fifty light years of our Sun.

But our Milky Way galaxy contains some *four hundred billion* stars, and it is just one of perhaps a hundred billion galaxies! That makes the number of candidate star systems truly mind-boggling. The very best lists of candidate "good suns" go out only a few hundred light years, and include only perhaps a few thousand stars. Beyond that distance, the waters are relatively uncharted.

A targeted search of promising Sun-like stars makes sense if extraterrestrial radio-using civilizations are relatively commonplace. That is, if there are many such civilizations, then the average distance between them is small, and ETI's home planet may in fact be orbiting a star now in one of our catalogues. But what if radio-using civilizations are scarce? Then, it stands to reason that the distance between oases is great in the interstellar desert. Under those circumstances, a potential life site may not even appear on our maps, its interesting stars being completely unknown to us. And, should scarcity be the rule, a targeted search of known stars is unlikely to prove productive. Another search strategy is called for.

That other search strategy is the all-sky survey. It differs from the targeted searches of early SETI in that no particular direction on the sky is favored. Rather than pointing at known stars, the sky survey sweeps out broad expanses, eventually sampling the whole sky visible from a given location in hopes of stumbling across an interesting signal. It's tedious and time-consuming work; but if you do not know which star ETI calls home, the best way to stumble across his signals may be to look in all directions.

Here then is where modern SETI diverges from Cocconi's and Morrison's search modality: by introducing a complementary search strategy—the all-sky survey—to fill the gaps left by targeted searches. Since we do not know for certain whether alien plentitude or scarcity holds true, we must conduct both searches until one or the other hits pay dirt.

Targeted searches fall well into the realm of the monster radio telescopes, of which Arecibo is the prime example. If you wish to point at a particular known star, you should do so with the highest-gain, most-directional antenna at your disposal to minimize interference from other potential emit-

The SETI League, Inc. **Link Analysis**

User specifies variables shown in **Bold**

Transmitter:

Frequency =	**1420** MHz;	λ =	21.1 cm		
Transmit Power =	**1.0E+06** W =	60.0	dBW =	90.0	dBm
Eff. Antenna diam. =	**305** meters =	1001 ft			
Illum. Efficiency =	**70** %				
Computed Antenna Gain =		1.4E+07	Ap =	71.6	dBi
Antenna Half Power Beamwidth =		7.1E–04	radian =	4.1E–02	degrees
Effective Isotropic Radiated Pwr =		1.4E+13	W =	161.6	dBm

Path:

Range =	**1.3** parsecs =	4.238	LY =	4.0E+16	m
Free Space Isotropic Path Loss =				367.5	dB
Incident Isotropic Power = EIRP – path loss =				–206.0	dBm

Receiver:

Eff. Antenna diam. =	**3.7** meters = 12.1390833 ft				
Illum. Efficiency =	**60** %				
Computed Antenna Gain =		1.8E+03	Ap =	32.6	dBi
Antenna Half Power Beamwidth =		5.9E–02	radian =	3.4E+00	degrees
Drift Scan Time (zero declination) =		1.3E+01	min =	807.9	sec
Recovered Power =		P inc	+ G ant =	–173.4	dBm
System Noise Temperature =		**50**	K =	–7.6	dB/To
Detector Noise Bandwidth =		**1**	Hz =	0.00	dB/Hz
Receiver Noise Threshold = kTB		= 6.9E–22	J/S =	–181.6	dBm
Integration Time =		**10**	sec =	5.0	dB/cy

SIGNAL TO NOISE RATIO 13.2 dB

Fig 1—Arecibo from Alpha Centauri.

ters. So for years, SETI was thought to require the kinds of facilities that only governments could afford. This is why the modestly funded NASA SETI program of the late 20th Century still consumed $12.5 million a year. That Congress terminated this planned ten-year search, just one year in, is proof that SETI was requiring the kinds of facilities that *not even* governments could afford.

All-sky surveys work with a different trade-off. It is true that large antennas have high gain and are sensitive to weak signals. But if our objective is to cover all 4π steradians of the sky, that sensitivity is buried under the burden of being blind to 99.9999% of the sky at a given time. For all-sky surveys, perhaps it makes more sense to sacrifice sensitivity for sky coverage.

That is where amateur SETI comes in, and with it, a new paradigm: SETI is possible with the kinds of facilities *you and I* can afford. The radio amateur's rather modest antenna, with a sensitivity two or three orders of magnitude below that of Earth's giant dishes, has the advantage of seeing hundreds to thousands of times more sky. A reasonable number of such antennas (on the order of 100, a critical mass achieved by The SETI League perhaps five years ago[14]) can scan the whole sky once per day. A more ambitious number, say 5000 such SETI stations, can monitor the whole sky *all the time, in real time*. This is the philosophy underlying Earth's most ambitious all-sky SETI survey, The SETI League's *Project Argus*, an initiative of radio amateurs around the world, which Phil Morrison enthusiastically supported during his last decade of life. Searching for interstellar communications, Morrison realized, had come to involve search strategies that went beyond merely identifying interesting nearby sun-like stars.

How Near Do We Hear?

Computed detection ranges esti-

The SETI League, Inc. **Link Analysis**

User specifies variables shown in **Bold**

Transmitter:

Frequency =	**1420** MHz;	λ =	21.1 cm	
Transmit Power =	**1.0E+06** W =	60.0	dBW =	90.0 dBm
Eff. Antenna diam. =	**305** meters =	1001 ft		
Illum. Efficiency =	**70** %			
Computed Antenna Gain =		1.4E+07	Ap =	71.6 dBi
Antenna Half Power Beamwidth =		7.1E–04	radian =	4.1E–02 degrees
Effective Isotropic Radiated Pwr =		1.4E+13	W =	161.6 dBm

Path:

Range =	**6** parsecs =	19.56	LY =	1.9E+17 m
Free Space Isotropic Path Loss =				380.8 dB
Incident Isotropic Power = EIRP – path loss =				–219.2 dBm

Receiver:

Eff. Antenna diam. =	**3.7** meters = 12.1390833 ft			
Illum. Efficiency =	**60** %			
Computed Antenna Gain =		1.8E+03	Ap =	32.6 dBi
Antenna Half Power Beamwidth =		5.9E–02	radian =	3.4E+00 degrees
Drift Scan Time (zero declination) =		1.3E+01	min =	807.9 sec
Recovered Power =		P inc	+ G ant =	–186.7 dBm
System Noise Temperature =		**50**	K =	–7.6 dB/To
Detector Noise Bandwidth =		**1**	Hz =	0.0 dB/Hz
Receiver Noise Threshold = kTB		= 6.9E–22	J/S =	–181.6 dBm
Integration Time =		**10**	sec =	5.0 dB/cy

SIGNAL TO NOISE RATIO 0.0 dB

Fig 2—Maximum Range Calculation for Amateur SETI.

mated at tens of thousands of light years between similarly equipped Arecibos may be all well and good, but how many of us are blessed with an Arecibo in our backyard? A more important question might be, "At what distance can I, with my typical backyard amateur SETI station, expect to detect an alien Arecibo beaming my way?" It is an important question because, as the late SETI pioneer Dr Bernard Oliver wrote in 1995, "If your system wouldn't detect the strongest signal the ETI might radiate, even if it came from the nearest star, then years of listening, or thousands doing it, won't improve the chance of success. To cross the Golden Gate, we need a bridge about 10,000 feet long. Ten thousand bridges...one foot long won't hack it."[15]

So let us run the numbers. My well-documented[16] backyard SETI station is typical of hundreds now operational or under construction around the world. It features a parabolic dish 3.7 meters in diameter, illuminated at 60% efficiency. My system noise temperature (including LNA noise figure, feed line losses, antenna noise temperature, and sky noise looking far from the galactic center) is on the order of 50 kelvins. My DSP software is set up for 1-Hz bin widths and 10 seconds of integration time. Assume a CW beacon from an alien Arecibo, running a 1-megawatt transmitter, beamed our way. My signal analysis spreadsheet[17] (Fig 1) shows, given the 1.3-parsec range to the very nearest star, that we can expect an impressive 12 dB SNR. That's an S-2 from Alpha Centauri, folks!

Perhaps even more interesting is the maximum range of detectability for the system described above. Let's assume that a unity (0-dB) SNR is adequate to identify the DX station. (Many of us routinely claim contacts where the signal was in fact well below the noise threshold; for EME contacts, negative SNRs are almost obligatory!) Note in Fig 2 that our maximum range for 0 dB SNR is on the order of twenty light years. Within that modest range are several dozen Sun-like stars, including Morrison's and Cocconi's most promising candidates, several of which are now known or expected to harbor planetary systems. So contrary to Barney Oliver's cautionary statement, we amateurs appear well able to cross the Golden Gate, even with our humble equipment. Whether there is anyone waiting for us at the other end of the bridge remains to be seen.

Factors Beyond Our Control

We have established that even a modest amateur SETI station can detect emissions from a civilization no more technologically advanced than our own, if it resides within 20 light years or so of Earth, and if it happens to be beaming toward us from the equivalent of our own Arecibo Observatory. But, how likely is ETI to actually direct a beacon our way, even given its existence in the right neighborhood, at the right technological level, in the right timeframe? Here I can only speculate about factors which are, in the words of the cynical Vicomte de Valmont in the De Laclos novel,[18] "completely beyond my control."

Social scientists tell us that only two possibilities motivate all human actions: altruism and self-interest, although some argue that even seemingly altruistic acts are performed with an underlying selfish motive. Can we imagine selfish or altruistic reasons why another civilization would expend considerable resources on the deliberate transmission of electromagnetic signals over interstellar distances? Much has been written about the altruistic case,[19] less about the selfish possibilities.

Successful altruistic civilizations, it has been theorized, harbor an innate desire to share their cultural wealth with those less fortunate. Such civilizations may consider it a cosmic imperative to undertake the transmission of their accumulated knowledge and experience to younger, emerging species. If this theory holds, we stand on the brink of reception of *Encyclopaedia Galactica*, a knowledge base that can transform human existence in ways we cannot begin to imagine. This justification for human SETI endeavors is only warranted if our cosmic companions are disposed to such generosity.

But what of the other possibility, that our galactic neighbors might choose to transmit in our direction, strictly out of self-interest? Of what possible benefit could such a transmission be to civilizations presumed older, wiser, and more capable than we? It's easy to concoct scenarios whereby the very act of reception of interstellar signals is somehow damaging to humanity and advantageous to the transmitting species. Competition rules the jungle, so why not the cosmos? And as Earth is, in essence, a paranoid planet, any such scenario that you can imagine will easily attract a host of followers willing to embrace it. I believe this says far more about the human condition than it does the alien. Further, such speculations have served to inhibit the acceptance and growth of SETI science on Earth as though, somehow, one can believe that turning a deaf ear to the universe can somehow protect us from harm.

There is a third possibility, little discussed in the literature, as to why we might someday find ourselves on the receiving end of an interstellar CQ. We believe that time and space are finite. Civilizations, as far as we understand the laws of nature, can be long-lived but not eternal. Imagine a technologically advanced civilization facing its own inevitable demise. Might it not wish to put its whole history and culture into an electromagnetic time capsule—a modern message in a bottle—in hopes that someone else (maybe us) might pluck it out of the cosmic pond, and simply know that they were? Might not they transmit in the hopes of achieving a degree of immortality? Might not we?

Given the above possibility, I can envision someday receiving a beamed transmission from a civilization long dead. It would seek to inform us about their art, culture, society, history, spirituality, hopes, dreams, and aspirations. Such a transmission could be an unparalleled look into a neighboring civilization's past—and humanity's future.

What Next for Amateur SETI?

The nonprofit, membership-supported SETI League, Inc[20] is a ham club formed in 1994, in the wake of Congressional cancellation of the short-lived NASA SETI program, to keep the search alive. During its first decade, the club's emphasis was on technical education. Our members wrote dozens of articles and papers, and gave scores of presentations[21] to like-minded radio amateurs at such meetings as the annual AMSAT Space Symposium; the Central States, Mid-Atlantic, West Coast, Northeastern, and Southeastern VHF Conferences; Society of Amateur Radio Astronomers meeting; International Space Development Conference; Dayton Hamvention; various ARRL Division and National Conventions; and elsewhere. Our mission in those early years was to demonstrate that credible science could be done by amateurs, with amateur equipment, and that assembling a workable SETI radio telescope was not only feasible, but affordable and rewarding.

During its second decade, the focus of The SETI League is shifting somewhat, into more of a coordination role. The microwave hardware and DSP software now are well-defined, with well over a hundred amateur SETI stations currently on the air in 67 countries on all 7 continents. Our next challenge is to achieve full sky and spectral coverage, with sufficient redundancy to ensure successful independent verification should an interesting candidate signal be detected. Given that The SETI League is an all-volunteer organization with officers who all have day jobs, families, and actual lives, such coordination taxes our limited resources. For-

More on SETI Probabilities

As Paul ably notes, Frank Drake's contributions to SETI are outstanding. In 1961, he and J. Peter Pearman organized the first SETI conference held at the National Radio Astronomy Observatory in Green Bank, West Virginia. At that small gathering, they proposed agenda representing discrete factors in the probability of interstellar communication. Almost jokingly, Drake assembled those factors into what's now known as the Drake equation or the Green Bank equation (see Paul's discussion at **www.setileague.org/editor/quantify.htm**). Although never intended to be used quantitatively, the equation relates to the number of civilizations in our galaxy from which someone would be likely to get that big CQDX.

The result is the product of seven variables, seemingly sorted left-to-right in order of increasing uncertainty:

$$N = R^* f_p n_e f_l f_i f_c L$$

where:

N=number of extraterrestrial civilizations that might expect contact from another

R^*=rate of star formation in our galaxy (stars/year)

f_p=fraction of those stars that have planets

n_e=average number of habitable planets per star that has planets (planets/star)

f_l=fraction of the above that develop life (civilizations/planet)

f_i=fraction of the above that develop intelligent life

f_c=fraction of the above that are able and willing to communicate

L=the average lifetime of such a civilization (years).

The observational uncertainty of the value of R^* is low compared with the uncertainties of the other variables: $R^* \approx 10$ is the accepted value. Drake and his colleagues estimated that about half the stars in our galaxy have planets, so they set $f_p \approx 0.5$; observation provides reasonably strong support. They set $n_e \approx 2$, although now there's plentiful evidence that typical stellar radiation levels in binary systems and those with red dwarfs would dictate a value lower by several orders of magnitude. f_l was set to unity because all our evidence points to the rise of life on Earth very shortly (in cosmic terms) after suitable conditions were met. The uncertainties of the other factors are quite high.

Note that civilization lifetime L could be redefined as the length of time a civilization has emitted recognizable signals, purposefully or not. As Dr Shuch demonstrates, electromagnetic signals from Earth not intended for extraterrestrial communications are unlikely to be detectable even at the nearest star; so willingness, or at least cognizance, is legitimately part of the equation. He emphasizes to *QEX* that the equation is much more useful in its originally intended role as a research tool than as an actual calculation. However, much debate continues over its numerical solution, often with misleading results because the uncertainties of some factors are so high.

Carl Sagan optimistically speculated that all factors but L were relatively high. His pessimism about the value of L had to do with our own tendency toward self-destruction. At present, observation shows that N=1. Think about that observation, but not for too long! Obviously, any assumptions about factors in the equation that produce values of N<<1 expose one or both of the following: 1) large uncertainties, or 2) that human beings are truly unique in the galaxy, by chance or otherwise. Without more data than we have now, the uncertainties of all the factors are uncertain! That's why the Drake equation is not yet a serious statistical tool.

We can keep trying to quantify the uncertainties, but being unique by any means can be a heavy-duty philosophical matter. At the base of modern scientific thinking regarding that is a thing called the anthropic principle. It comes in two flavors: weak and strong.

The weak anthropic principle states that we shouldn't be surprised by what we observe in our corner of the universe, since conditions must have developed in it that allowed us to be here to observe them. As Stephen Hawking wrote in *A Brief History of Time* (Bantam, 1988, ISBN: 0-553-05340-X), "It is a bit like a rich person living in a wealthy neighborhood and not seeing any poverty." In other words, if you measure that Earth is 4.5 billion years old, don't be surprised—that's how long it took for human beings to appear who were able to make the measurement.

The strong anthropic principle is much broader than the weak. It states that intelligent life as we know it couldn't exist under conditions significantly different from our own. In other words, if life as we know it didn't exist, then likely neither would our universe. That implies that a universe like ours *ought* to admit life as we know it at some point. Dr Hawking wrote, "The laws of science, as we know them at present, contain many fundamental numbers, like the size of the electric charge of the electron and the ratio of the masses of the proton and the electron....The remarkable fact is that the values of these numbers seem to have been very finely adjusted to make possible the development of life. For example if the electric charge of the electron had been only slightly different, stars either would have been unable to burn hydrogen and helium, or else they would not have exploded [to redistribute heavier elements].... One can take this either as evidence of a divine purpose... or as support for the strong anthropic principle." I told you it was heavy-duty philosophy!

The often-used estimate of $f_i \approx 0.01$ may be way too optimistic. Our solar system's orbit about the galactic center is evidently nearly circular, at such an orientation that it avoids significant radiation and ejecta from novae for hundreds of millions of years at a time. Other star systems are not so lucky. Systems like ours may be five orders of magnitude rarer than the average, indicating $f_i \approx 10^{-7}$. That might make us feel a bit lonely. On the other hand, taking only our own experience and setting $L \approx 50$ may be too pessimistic. Civilizations might be able to communicate long after they've departed the scene. That likelihood increases with the time they actually lived.

Could ours be the only planet in a galaxy of hundreds of billions of stars supporting this kind of discussion? Some feel another question should be asked: "Should we begin our search for intelligent life right here at home?" But Dr Shuch warns that funding for that research is unlikely because of its low probability of success.—*Doug Smith, KF6DX* tunately, we have at our disposal a communications infrastructure that exceeds the very best that Amateur Radio had to offer in years past: a fast, efficient, global and affordable Internet.

We hams tend to decry the advent of the Internet as somehow undermining the vitality of our hobby. While I tend to agree that Amateur Radio as we knew it in the mid-20th Century is probably a thing of the past, I embrace the digital revolution as a positive force in ham radio's future. Without it, collaborative science across borders, such as is now routinely practiced by the roughly 1500 SETI League amateurs around the world, would have been impossible. So, as the scope of our vocation begins to change in keeping with the new reality, can we begin to set our sights on the Ultimate DX.

In one of his last publications,[22] Phil Morrison wrote, "The key parameter is not simply the range, DX, but rather DX/c; it is time and not space that sets the firmest limits. Indeed, our Milky Way supports a hundred thousand light year range to its outskirts, but the round-trip transit time is longer than the full age of our articulate species. Historical time enters the rules. Human ability to signal across space beyond the hundred-mile horizon, open to the sunlit mirrors of mountaintop surveys or to ship's signal rockets by night, is no older than Marconi. Our SETI Milky Way surveys cannot rely on the guide of symmetry, for we are almost certain to be the juvenile in any dialogue." Given the youth and immaturity of our species, how can it be said that amateurs know less, or have less to contribute, than anyone else?

I am confident that once the first credible signal from beyond is detected and confirmed, countless others will follow. Whether that first detection is made by an amateur, or by someone with lesser skills but greater resources, I will not begin to speculate. But it is inevitable that follow-on observations and subsequent discoveries will surely be made by the world's amateur SETIzens. I hope you'll join us in this global quest for cosmic communications.

Conclusion

In summarizing the rationale for a redoubled amateur SETI effort, I can do no better than to reiterate Morrison's and Cocconi's concluding paragraph, including its final two sentences, cited earlier in this paper:

"The reader may seek to consign these speculations wholly to the domain of science fiction. We submit, rather, that the foregoing line of argument demonstrates that the presence of interstellar signals is entirely consistent with all we now know, and that if signals are present the means of detecting them is now at hand. Few will deny the profound importance, practical and philosophical, which the detection of interstellar communications would have. *We therefore feel that a discriminating search for signals deserves a considerable effort. The probability of success is difficult to estimate, but if we never search, the chance of success is zero.*"

Notes

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[21] H. Paul Shuch, "Introduction to Amateur SETI." *AMSAT Journal* 20(5): 14-17, September/October 1997.

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H. Paul Shuch, N6TX, brings to the SETI enterprise more than 40 years of leadership experience in ham radio, electronic communications and radio astronomy, including having served as Executive Director of the nonprofit SETI League since its inception in 1994. A distinguished engineering professor, Dr Shuch earned his PhD. in Engineering from the University of California, Berkeley, after serving in the US Air Force as a telecommunications systems controller during the Vietnam conflict. Founder and chief engineer of Microcomm, the Silicon Valley start-up credited with developing the world's first commercial home satellite TV receiver, he served as Technical Director and Chairman of the Board of Project OSCAR Inc, builders of the first non-Government communications satellites.

While an engineer with such major aerospace corporations as Itek and Lockheed, Dr Shuch was responsible for the design of electronic countermeasures, satellite remote sensing and submarine launched ballistic missile telemetry systems. He holds patents in the areas of airborne radar and phased array antennas, and taught electronics and avionics at various colleges and universities for a quarter century. Dr Shuch is the author of over 400 publications, of which nearly half deal with radio astronomy, SETI science, and related technologies. He is the publisher of Contact In Context*, the first on-line, indexed and peer reviewed SETI academic journal. Paul is a member of the International Academy of Astronautics, Vice-Chair and Webmaster for its SETI Permanent Study Group; Principal Investigator and Webmaster for the Invitation to ETI initiative; Vice President and Webmaster for the Society of Amateur Radio Astronomers, a life member of the Society of Wild Weasels, AMSAT, and ARRL; a member of the AACS Alumni Association; a Fellow of the British Interplanetary Society and the Radio Club of America; and serves on numerous international committees and boards.*

The Harmanized R4C— A High-Performance Analog HF Receiver

The authors' mods to a vintage Drake rig turned a decent receiver into a competition-grade radio.

By Phil Harman, VK6APH, and Steve Ireland, VK6VZ

The Drake R4C is regarded as perhaps the best amateur band receiver of the 1970s and one of the best of all time when modified according to Rob Sherwood, NCØB, and George Heidelman, K8RRH's classic series of articles in *Ham Radio* magazine.[1]

The replacement of the R4C's 8-kHz-wide first IF filter, which allowed close-in signals through to the second mixer with disastrous results with a 600 Hz 6-pole filter,[2] was the first step in turning the receiver from a good receiver into a great CW contesting and DXing radio. The next was to reduce the gain (by about 20 dB) in front of the often overloaded second mixer and restore this gain after the second IF crystal filters, at the high impedance grid of the third mixer, using a cascode JFET amplifier.

These modifications radically improved the close-in blocking dynamic range of the R4C receiver from 58 dB at 2 kHz to around 85 dB. Coupled with 500 or 250 Hz filters in the second IF, the replacement of the two diode detector (which allowed detected audio to leak back into the final IF stage and modulate it) for an active double balanced mixer and the replacement of the class-A audio amplifier (which one commentator remarked made the R4C sound like a bad transistorized 1960s car radio) the R4C became rather special and *the* radio of choice for contesters and DXers.

[1]Notes appear on page 40.

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Still the Receiver of Choice

During the mid-1990s, one of the authors, VK6VZ, revived his original interest in CW DXing on 1.8 MHz and found that the R4C with Sherwood modifications was still the receiver of choice for many "topbanders." With its 85 dB dynamic range at 2 kHz and extremely low phase noise, it was as good as and mostly better than even the higher performing synthesized transceivers, with their VHF first IFs and (consequently) wide first IF filters, when it came to pulling weak signals

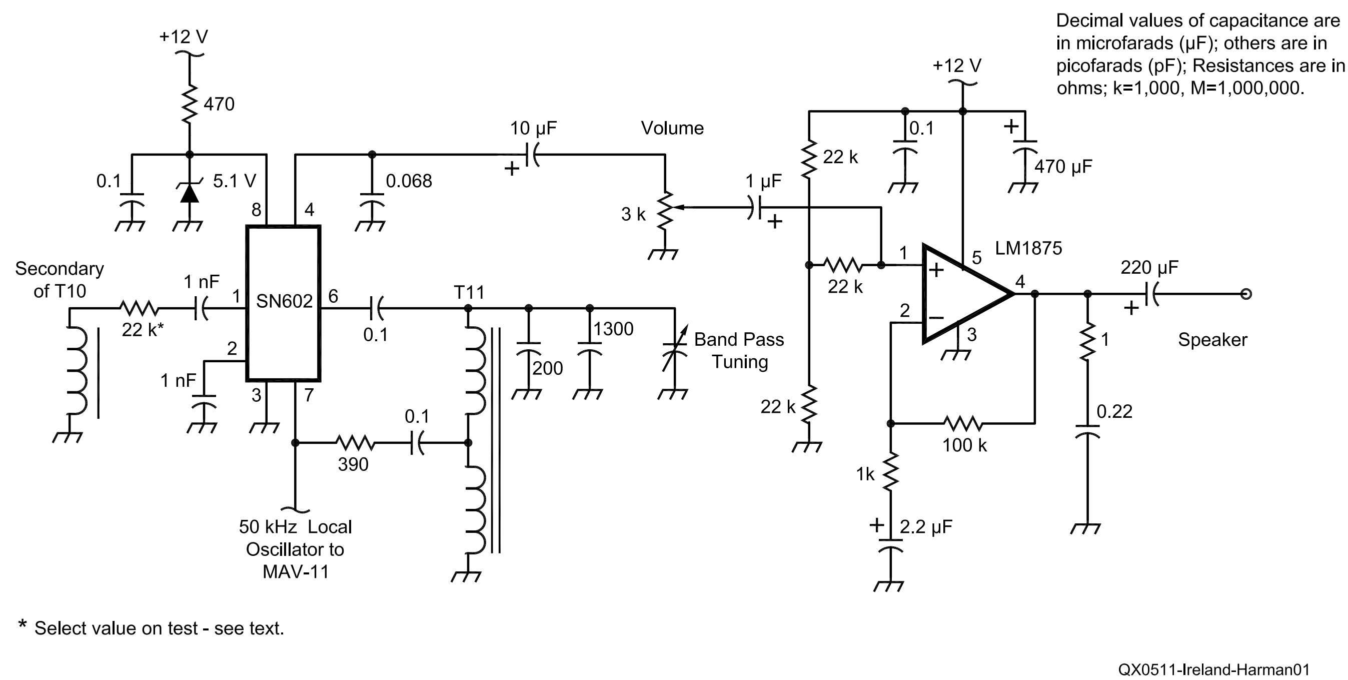


Fig 1—Circuit of product detector and AF amplifier for Harmanized R4C.

out of noise, particularly under contest conditions.

VK6VZ also discovered that Tom Rauch, W8JI, had carried out further modifications to the Drake R4C, replacing all of its tube mixers with high performance double balanced solid state devices and its permeability tuned front end with individually switched band-pass filters.[3] Tom had also totally rebuilt the AGC system. In reality he had kept the Drake receiver architecture, but totally rebuilt the receiver with modern solid-state devices.

As a result, after an enormous amount of work, W8JI produced a receiver with a measured 2 kHz blocking dynamic range of 127 dB—considerably better than any amateur band transceiver currently on the market.

One evening, in the shack of VK6APH, who in an earlier incarnation had designed RF receiving systems professionally, the two authors decided to carry out a project to rebuild/re-engineer a "junker" R4C owned by VK6VZ to have *at least* the performance of the Sherwood R4C and include some of the techniques used by W8JI.

However, there were two major snags: many of the solid state components used by NCØB and K8RRH were no longer available and W8JI had never written up his work for publication.

VK6APH then sought advice from W8JI on aspects of the circuit design and started design work in earnest. The Harmanized R4C uses the Sherwood 600 Hz first IF filter, along with the IF gain redistribution and product-detector design principles established by NCØB and K8RRH. The 12 V regulated supply uses the design principles of an R4C modification by Howard Sartori, W5DA, originally published in *73* magazine back in 1979.

The authors wish to formally thank NCØB, K8RRH, W8JI and W5DA, without whose ideas the Harmanized R4C would not exist.

A Moderate Rebuild

The Harmanized R4C occupies the middle ground between the original NCØB/K8RRH modifications and the radical rebuild carried out by W8JI. The moderate rebuild described here uses cheap and readily available components and offers a blocking dynamic range at 2 kHz of 95 dB—better than any commercial Amateur Radio available in 2004 with the exception of the Ten-Tec Orion and the FlexRadio SDR1000.

It should be noted that this "moderate rebuild" is of reasonable technical complexity and needs to be carried out by someone with considerable experience in building or modifying radio equipment. While this feature is reasonably comprehensive, it does not give a "blow-by-blow approach and carrying out these modifications involves careful reading and interpretation of both the Drake and Harmanized R4C circuit diagrams.

Major Features

In summary, the Harmanized R4C has the following features:

- High performance LM1875 audio amplifier (THD at 20 W/1 kHz of 0.015% typically).
- 12 V regulated supply (+12 V and –12 V) for solid-state circuitry.
- SN602 product detector.
- SN602 as R4C third mixer.
- Minicircuits SBL-3 double balanced mixer (+7 dBm LO) as R4C second mixer.
- MAV-11 MMIC to increase IF gain after the R4C second mixer.
- Sherwood 600 Hz first IF CW filter, with relay switching so the stock 8-kHz-wide first IF filter is retained for SSB/AM operation.
- Completely new three-speed hang AGC system. The AGC voltage is applied after the second-IF crystal filters and allows the gain of all the stages before the R4C's third mixer to run at maximum. The AGC has one speed for very fast CW (35 wpm-plus), plus one for slower CW and one for SSB.

It should be noted that the R4C

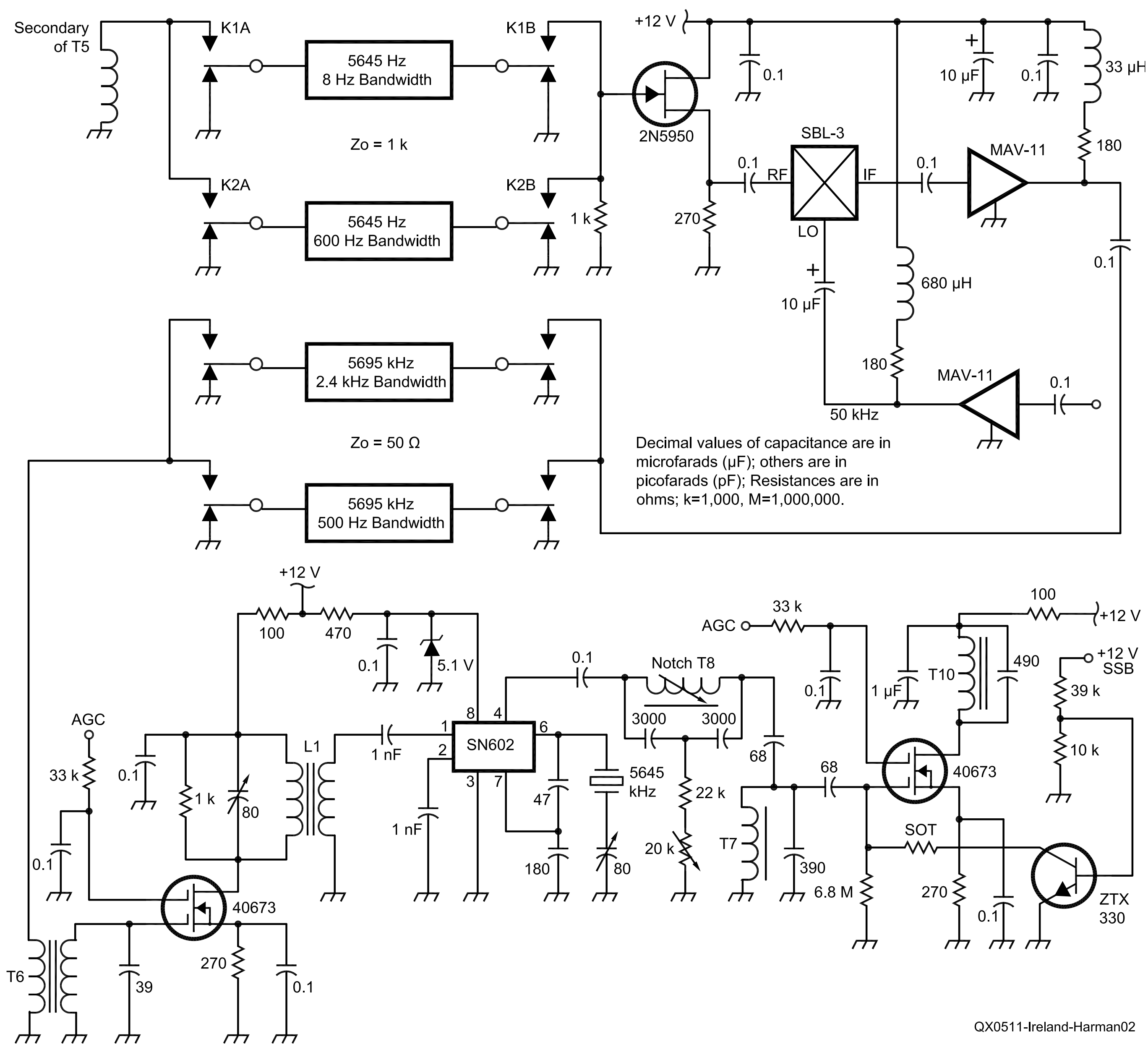


Fig 2—Circuit of IF filter switching, IF gain redistribution and second mixer.

used for the modifications was a late model, with 6EJ7 first and third mixers. The Harmanized R4C preserves the original 6BA6 RF amplifier and its associated permeability-tuned RF circuitry, along with the 6EJ7 first mixer, on the basis that these work reasonably well, but all other tubes in the signal path are replaced by solid-state circuitry.

The 6BA6 RF amplifier has had resistor R3 removed to increase the screen grid V_s on the tube and improve its intermodulation performance—a modification suggested by W8JI.

One interesting side effect of the removal of most of the tubes is the stability of the R4C's transistorized permeability tuned oscillator seems improved, perhaps because of the subsequent reduction of heat inside the chassis!

In its second IF, prior to modification, the R4C had been fitted with the optional Drake 500 Hz CW crystal filter, supplementing the stock Drake 2.4 kHz eight-pole crystal filter. While the shape factor of the 25-plus year-old 500 Hz filter isn't great, the filter is a particularly effective selectivity aid in a "pile-up" of CW stations when combined with the Sherwood 600 Hz first IF filter and the R4C's Passband Tuning. VK6VZ has found this combination more effective in CW pile-up situations than the IF Shift/Width controls on his FT-1000MP, which is fitted with 500 Hz filters in its second and third IFs.

VK6VZ rebuilt the R4C's permeability-tuned oscillator tuning mechanism, using information provided by K1VY on the Internet,[4] which greatly improved the ease of tuning.

Hamfests of all shapes and sizes

are great places to find "junker" R4Cs.

12 V Regulated Power Supply

This modification is an update of the excellent article by Howard Sartori, W5DA, in the June 1979 issue of *73* magazine[5] and involves the replacement of the existing R4C 12 V dc supply for solid state devices. The new supply uses the 17 V-0 V-17 V secondary of the existing R4C mains transformer, four 1N4007 rectifier diodes with 7812 and 7912 regulator chips.

The original R4C generated +12 V dc by reducing the 150 V supply to 12 V using "drop-down" resistors. This means of producing 12 V resulted in over 15 W of heat being coincidentally generated.

As well as getting rid of this extraneous source of heat, the new +12 V regulated supply, using an LM7812 regulator (see Fig 3) also greatly reduces the noise and hum that were generated by the old supply.

In order to supply the hang AGC system used by the Harmanized R4C, an additional –12 V regulated supply is needed. This is provided using an LM7912 regulator—see also Fig 3.

Audio Amplifier

The LM1875 is a monolithic 20 W power amplifier IC, in a TO220 package, with an open loop gain of 90 dB. The circuit (see Fig 1) is designed to suit speakers from 4 Ω to 8 Ω impedance and enables the R4C speaker matching transformer (T13) be removed, making the anti-VOX facility (which uses the same transformer) redundant.

The volume control shown in Fig 1 is the existing Drake AF Gain 3 kΩ potentiometer.

Product Detector

A detector in a receiver needs to have good isolation between its input and output. Because the original two-diode detector in the R4C is not double-balanced, it allows its audio output to leak back to the detector input, which then modulates the third (50 kHz) IF stage.

A simple, active double balanced product detector using the SN602 mixer integrated circuit is substituted for the two-diode detector—see Fig 1. The Drake components CR2, CR3, C83 and C84 are removed, with C83 being replaced with a jumper.

For those who may raise an eyebrow at the use of a lowly SN602 as

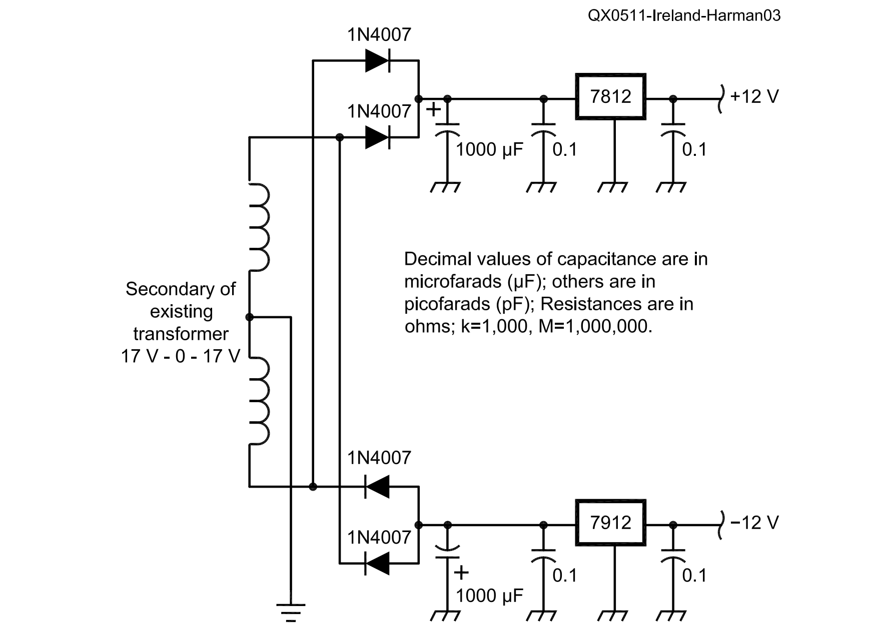
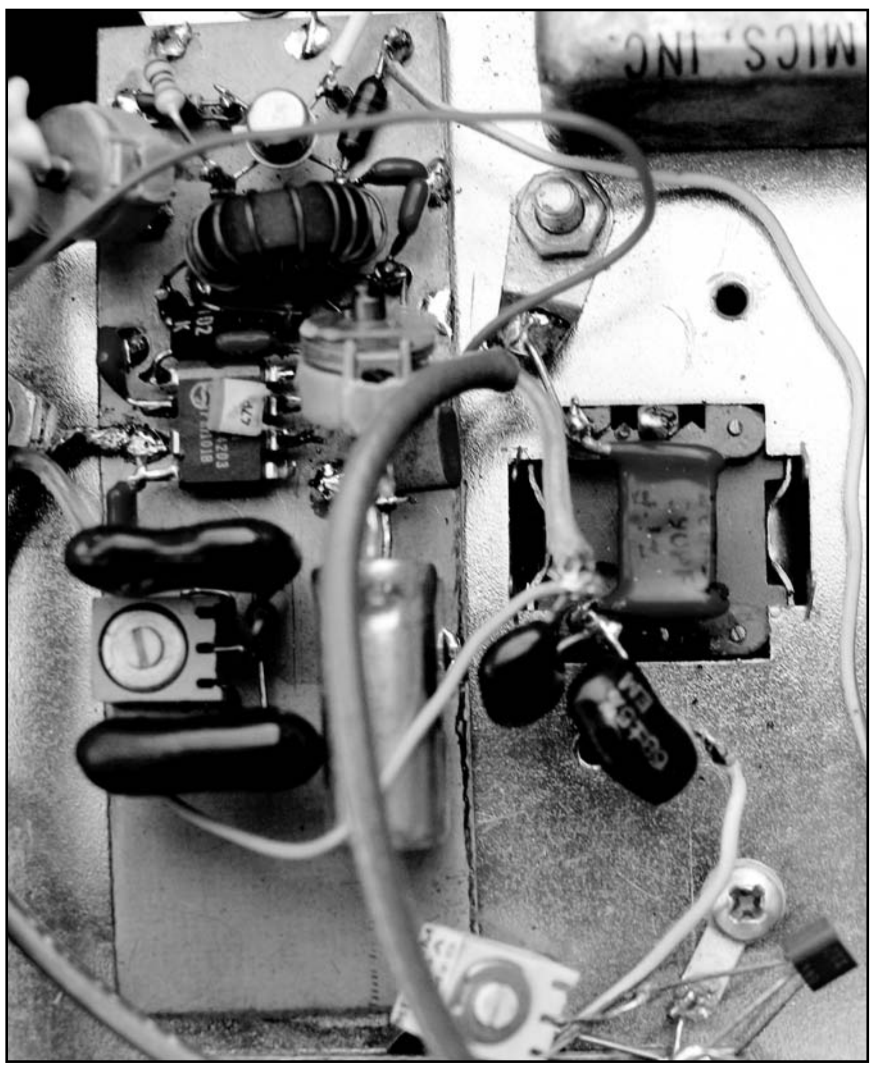


Fig 3—Circuit of +12 V and –12 V regulated power supply.



SN602 product detector PCB with Drake T10 (top left) and T11 (lower left).

Components

Unless otherwise specified, resistors are 1/4 W, 5% tolerance carbon-composition or film units.

Audio Amplifier (see Fig 1)
1 Ω, 1k, 22k (3), 100 kΩ resistors.
1 µF, 2.2 µF, 220 µF, 470 µF 50 V electrolytic or tantalum capacitors.
0.1 µF, 0.22 µF ceramic capacitors.
LM1875 20-W power amplifier, TO-220 package.
Small heatsink to suit LM1875.

Product Detector (see Fig 1)
470 Ω, 22 kΩ resistors.
0.001 µF (2), 0.068 µF, 0.1 µF (3) ceramic capacitors.
10 µF, 50 V electrolytic or tantalum capacitors.
SN602 mixer integrated circuit.
5.1 V, 1 W Zener (1N4733/BZX85C5V1 or similar).

Redistributing the IF gain/Sherwood 600 Hz filter /replacing second and third mixers (see Fig 2)
Notes: T5, T6, T7, T8 and T10 are original Drake transformers—see R4C circuit diagram/manual. The 5645 kHz crystal is an existing R4C component, as is the 80 pF variable capacitor (actually C59, which is 7-60 pF). The 390 pF shown across T7 is the existing C49. The 490 pF across T10 is the existing C65. If your R4C is not fitted with the optional Drake 500 Hz second IF (5695 kHz) filter, the Inrad 2603.2 filter can be used instead.

100 Ω, 270 Ω (3), 100 Ω, 180 Ω, 470 Ω, 1k (2), 10k, 22k, 33k, 39k, 6.8M, SOT (see below) resistors.
20k trimpot, 0.125 W.
SOT resistor. This is "selected on test," so that when the narrow Drake 500 Hz second IF crystal filter is switched in circuit, the S meter reads the same as when the wider 2.4 kHz filter is selected.
0.001 µF (2), 0.1 µF (8) ceramic capacitors.
39 pF, 68 pF (2), 47 pF, 180 pF mica capacitors.
80 pF variable trimmer capacitor—see L1 below.
10 µF (2) 50 V electrolytic or tantalum capacitors.
L1 uses the ferrite core from Drake T16. The core is wound with a primary (in the drain of a 40673 MOSFET) of 13 turns of #22 SWG wire and a secondary of 6 turns of #22 SWG wire. L1 is tuned by an 80 pF trimmer capacitor, which is adjusted for maximum signal.
33 µH radio frequency choke.
680 µH radio frequency choke.
2N5950 field effect transistor.
ZTX330 transistor.
40673 MOS field effect transistor (2).
SN602 mixer integrated circuit.
Sherwood CF-600/6 six-pole 600 Hz CW filter (5645 kHz). This is currently US$135 from Sherwood Electronics—see **www.sherweng.com**.
12 V two-pole sub-miniature relays. Suitable relays can be obtained from Radiospares—see **www.rs-components.com.au**, stock no. RS345-038 (ultra miniature PC relay 12 V dc).
Mini-circuits SBL-3 mixer (see **www.alltronics.com**). The address is PO Box 730, Morgan Hill, CA 95038-0730; e-mail **ejohnson@alltronics.com**.
MAV11 amplifier chip (2). These can be obtained from Down East Microwave (**www.downeastmicrowave.com**), stock no. 1104.

AGC system
Note: T10 is the original Drake transformer—see R4C circuit diagram/manual.

Regulated PSU (see Fig 3)
1N4007 diodes (4).
1000 µF, 50 V electrolytic or tantalum (2) capacitors.
0.1 µF (4) ceramic capacitors.
7812 +12 V regulator chip.
7912 –12 V regulator chip.

Hang AGC (see Fig 4)
100 Ω, 220 Ω, 470 Ω, 1k, 3.3k, 4.7k, 10k (8), 15k, 56k, 100k (2), 1M, 6.8M, 22M resistors.
2k, 500k trimpots, 0.125 W.
0.01 µF, 0.1 µF (8) ceramic.
1 µF polystyrene.
BC548 (2) transistors.
2N3819 (2) field effect transistors.
LM324 integrated circuit.
1N914 (4) diodes.
5.1 V, 1 W Zener diode (1N4733/BZX85C5V1 or similar).

the product detector should note that it is protected by the narrow bandwidth of the filters preceding it. The SN602 is also used as the product detector in the receiver section of the well-respected Elecraft K2 transceiver.

The "T11" shown in Fig 1 is the Drake BFO transformer, while the secondary of the transformer, shown in series with the 22 kΩ resistor and the 1 nF capacitor that are connected to the input of the SN602, belongs to the Drake T10. The 200 pF capacitors and the 1300 pF capacitor shown are part of the original R4C Band Pass Tuning.

The value of the 22 kΩ resistor may need to be changed/optimized ("selected on test") once the Harmanized R4C AGC system is operational, in order to prevent overdriving the input of the SN602.

Redistributing the IF Gain/ Sherwood 600 Hz Filter/ Replacing the Second and Third Mixers

The Harmanized R4C is fitted with two crystal filters in both the first and second IF sections. The first IF (5645 kHz) filtering is switched between the stock four-pole 8-kHz-wide Drake SSB/CW filter and the Sherwood CF-600/6 six-pole 600-Hz CW filter, using 12 V miniature double-pole relays—see Fig 2. Similarly, the second IF (5695 kHz) filtering is switched between the stock Drake 2.4 kHz-wide SSB/CW filter and the optional Drake 500-Hz CW filter.

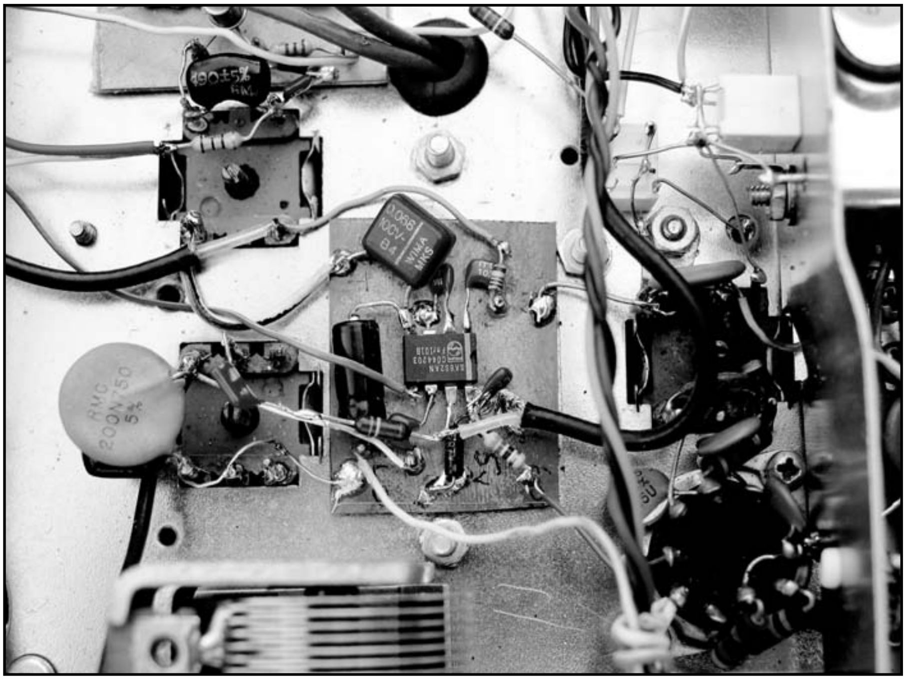
The two relays that control the first IF filter selection are switched using a miniature single-pole single-throw switch, mounted on the far left-hand side of the Harmanized R4C front panel, immediately to the left of the XTALS switch. The two relays used to select a second-IF filter are switched using the existing R4C MODE rotary switch.

If you have a R4C without a second-IF CW filter, suitable 500 Hz and 250 Hz filters can be obtained from Inrad.[6]

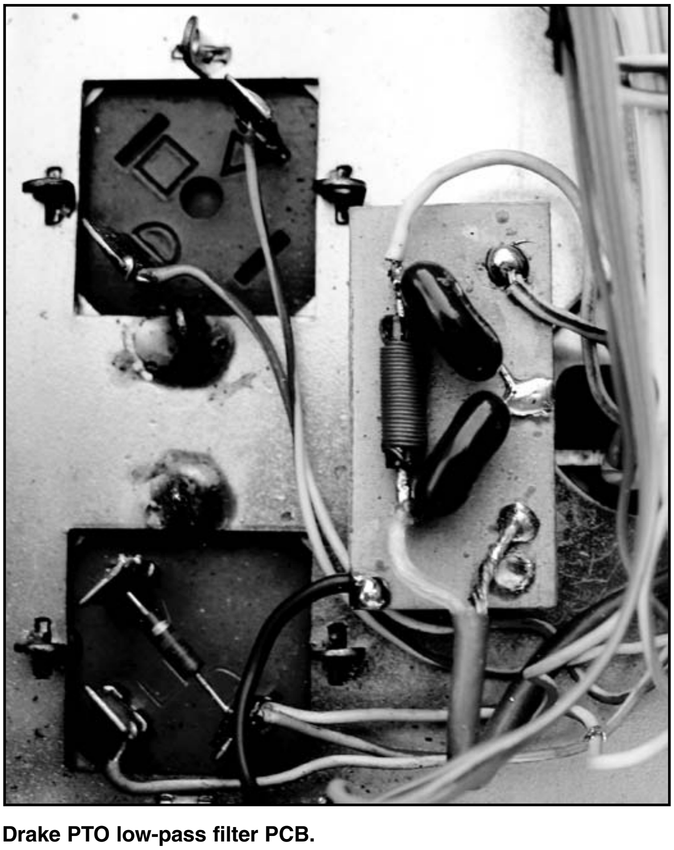
The R4C second-mixer tube (6BE6) is replaced by Mini-circuits SBL-3 double-balanced mixer and a MAV11 amplifier chip. The MAV11 effectively provides a broadband termination to the SBL-3 without the need for a diplexer. Another MAV11 provides sufficient local-oscillator drive (at 50 kHz) for the SBL-3. Note that the gain before the SBL-3 second mixer is reduced from that of the original Drake second mixer circuit.

The R4C third-mixer tube (6EJ7) is replaced with a SN602 mixer chip, preceded by a 40673 dual-gate MOS FET amplifier. The crystal injection for the third mixer (at 5645 kHz) is provided by an oscillator circuit, using part of the SN602 mixer chip.

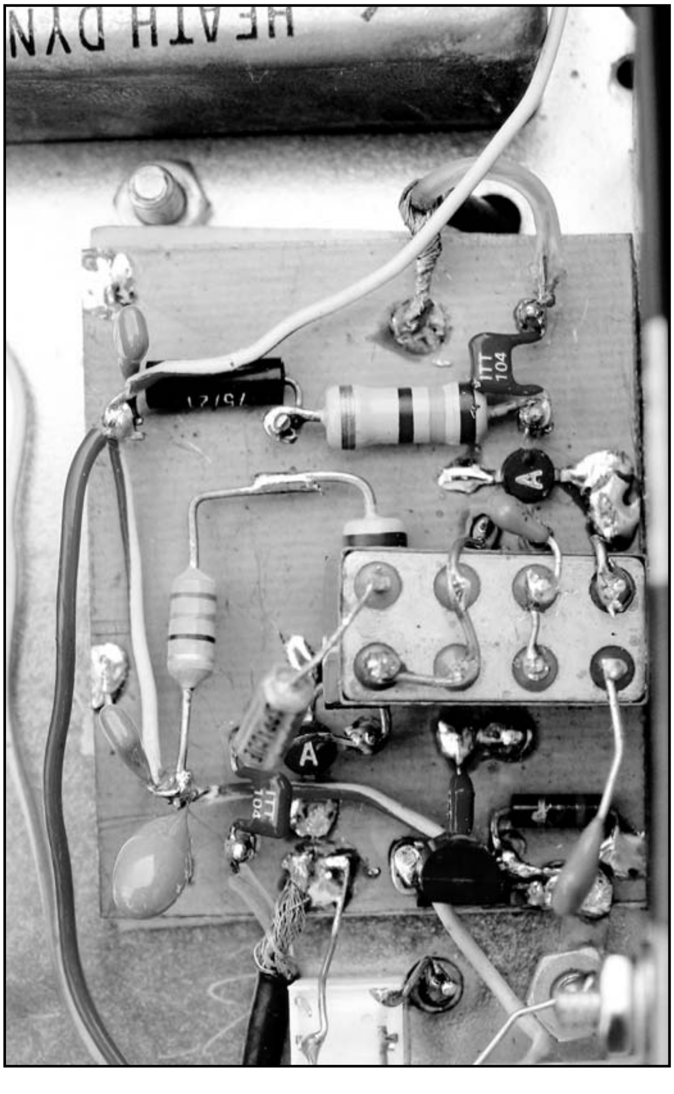
Again, the use of an SN602 is appropriate, owing to the narrow bandwidths of the filters preceding it.



SN602 third mixer PCB with Drake T7 at center and SBL-3 mixer (at right).



Drake PTO low-pass filter PCB.



Mini-Circuits SBL-3 second-mixer PCB.

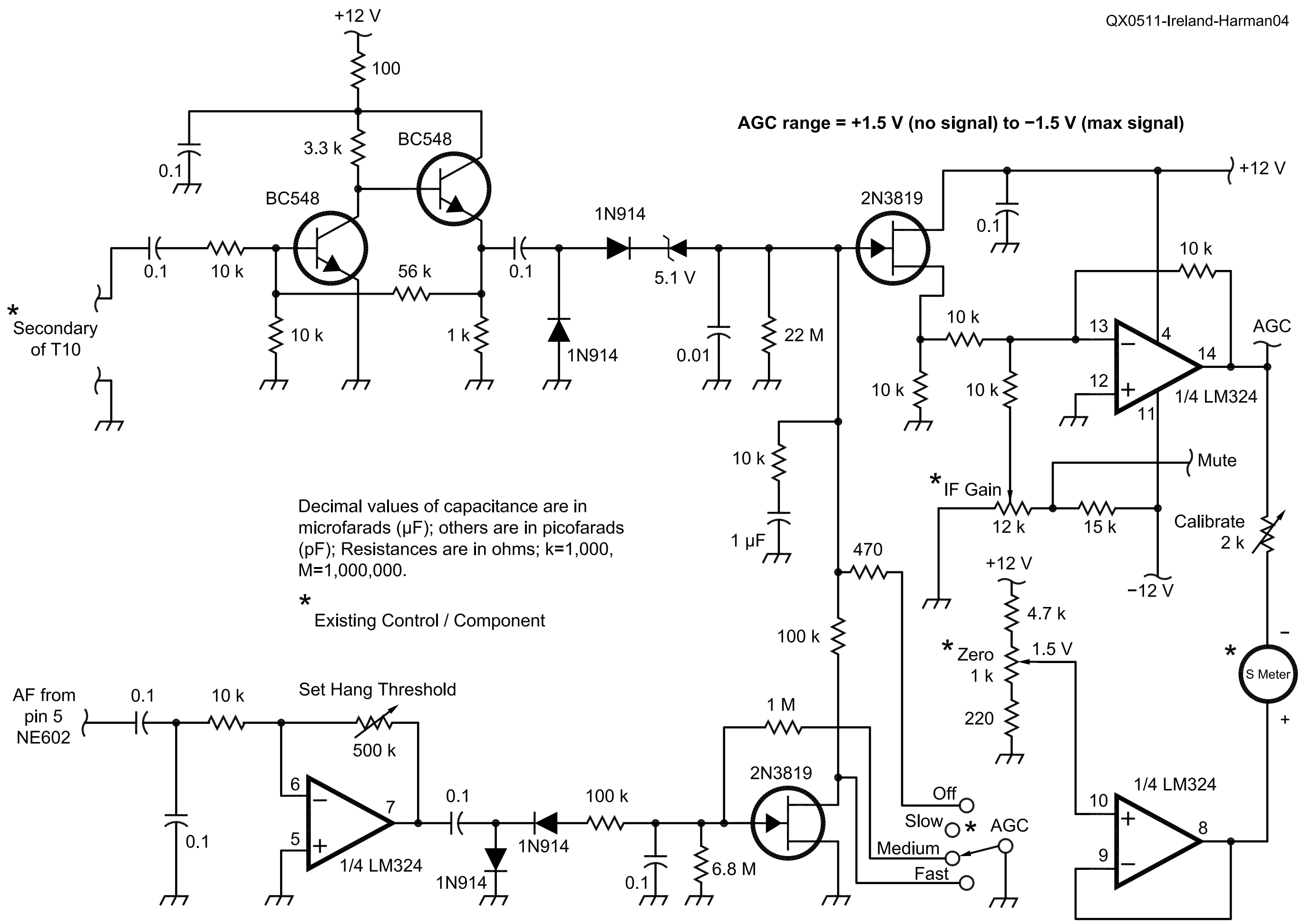


Fig 4—Circuit of hang AGC circuit.

The resistor marked SOT in Fig 2 is "selected on test" so that when the narrow Drake 500 Hz second IF crystal filter is switched in circuit ("CW .5") the "S" meter reads the same as when the wider 2.4 kHz filter is selected. This is achieved by switching in the 500 Hz second IF filter, tuning in a steady carrier, noting the "S" meter reading, switching to the 2.4 kHz second IF filter ("SSB") and adjusting the SOT resistor so it gives the same "S" meter reading.

AGC System

The hang AGC system uses the aforementioned 12 V dc regulated power supply that provides +12 V and –12 V supply rails—see Fig 3.

The AGC system provides three basic settings, apart from "off," and uses the existing Drake rotary AGC switch. In the "fast" setting, a conventional fast attack and decay system is implemented, while the "medium" and "slow" positions use a full hang AGC system. When receiving SSB signals, the "slow" position is normally selected and, on strong signals, this results in almost total silence between pauses in speech.

The actual hang AGC system (see Fig 4) uses a pair of BC548 transistors that provide a voltage gain of approximately six times, in order to amplify the 50 kHz IF signal to a sufficient level to be diode-detected. The two 1N914 diodes voltage-double the IF signal to produce a dc voltage proportional to the signal level. The 5.1 V Zener diode delays the AGC action until the signal input level has reached approximately –100 dBm, that is, the AGC knee level.

The 0.01 µF capacitor and 22 MΩ resistor provide the long time-constant for the hang AGC system. The 2N3819 provides a high input impedance and low output impedance to drive the unity-gain inverting AGC amplifier.

The inverting input of the LM324 provides a convenient point to inject a variable negative voltage to set the IF gain, with the original Drake RF gain control being used for this purpose.

The AGC decay time is determined by the second 2N3819, connected to the gate of the first FET. With the AGC switch in the "fast" position a 100 kΩ resistor is placed in parallel with the 22 MΩ resistor, providing a very fast attack and decay characteristic and effectively turning the hang action off.

At the other AGC settings, the 2N3819 controls the hang time, dependent on the charge on the 0.1 µF capacitor connected to its gate. This capacitor is charged from a dc voltage derived by rectifying the incoming audio taken directly from the product detector.

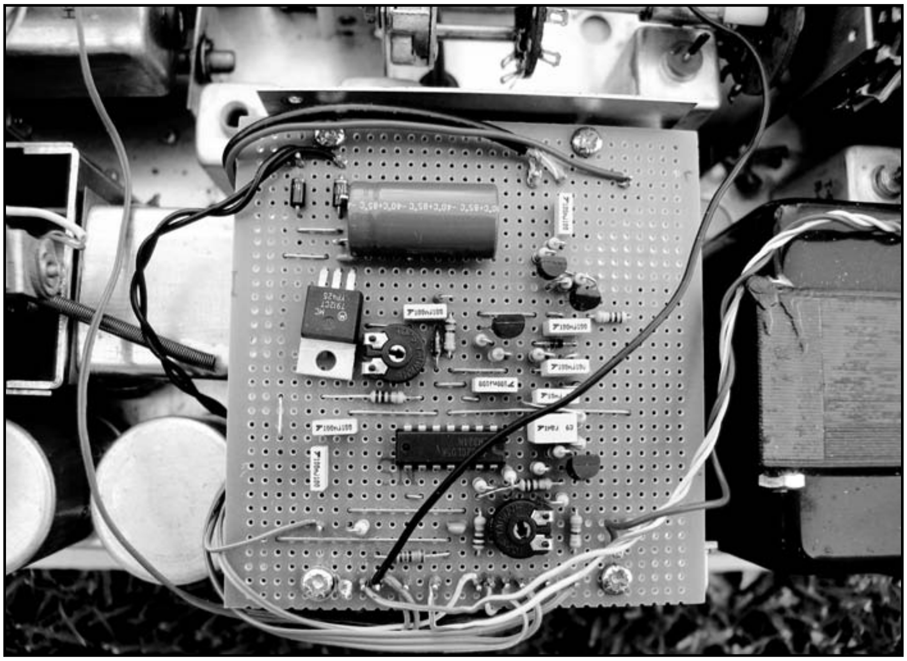
The actual hang threshold can be set to the operator's preference by adjusting the gain of the AF amplifier, using the 500 kΩ variable resistor.

The AGC voltage is applied to the second gate of the two 40673 dual gate MOSFETs. The AGC range of these two stages is quite adequate for the majority of received signals and only requires the use of the RF gain control on the strongest of signals.

The perfectionist would have delayed the AGC applied to the first 40673 but, in practice, it was found that this had minimal effect on the AGC characteristics.

While additional AGC range could be provided through the addition of a PIN diode attenuator earlier in the signal path, the fact is that if the AGC was applied in this manner, it would be milliseconds too late when a narrow IF filter is used and, consequently, would do nothing to improve the AGC performance of the receiver.

The type of AGC system used in the Harmanized R4C has been described in both the classic *Solid State Design for the Radio Amateur* by Wes Hayward, W7ZOI, and Doug DeMaw, W1FB, and more recently in its successor *Experimental Methods in RF Design* by Wes Hayward, Rick Campbell, KK7B, and Bob Larkin, W7PUA.



AGC system, including power supply.

Construction

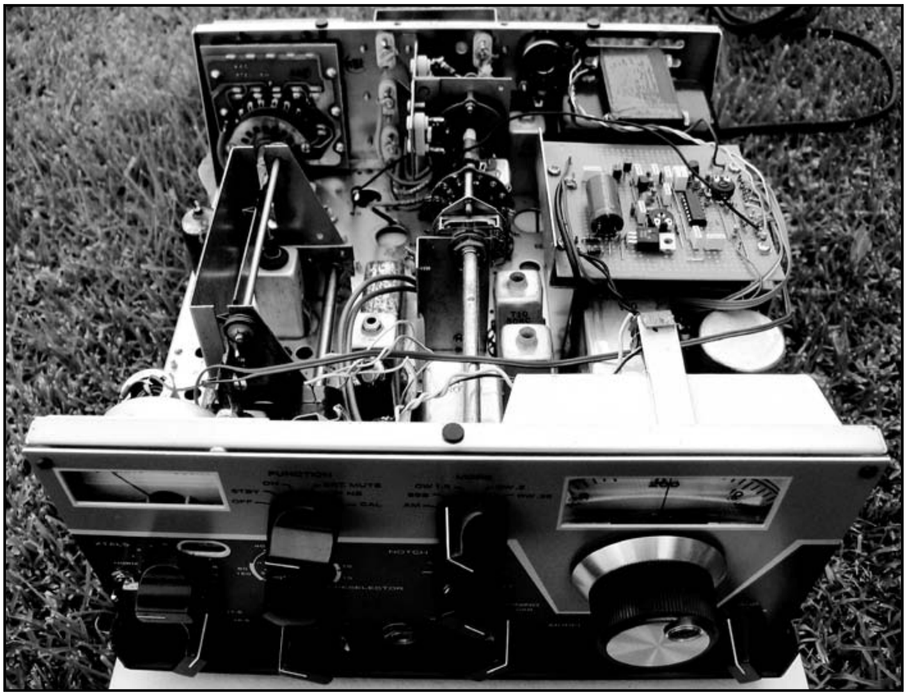
The best way to carry out the modifications described above is to do them one at a time and then retest the receiver with the idea of keeping a working receiver throughout the modification process and see its performance progressively improve.

The order of the modifications shown in the "major features" section is the order in which they were carried out by VK6APH.

Most of the construction is carried out "dead bug" style on four small pieces of single-sided PCB and mounted underneath the chassis. The photographs accompanying this article show how this was done. Wherever possible, the original Drake components were used, particularly low-value mica capacitors.

The audio amplifier PC board is mounted in the space between the Drake 4-NB (noise blanker) mounting brackets, while the AGC system is built on a piece of matrix board and mounted on top of the 4-NB mounting brackets—see photo.

The +12 V power supply is mounted on the edge of the R4C chassis, adjacent to the crystal calibrator, and uses the chassis as a heatsink. The –12 V power supply coexists on the same circuit board as the AGC circuitry.



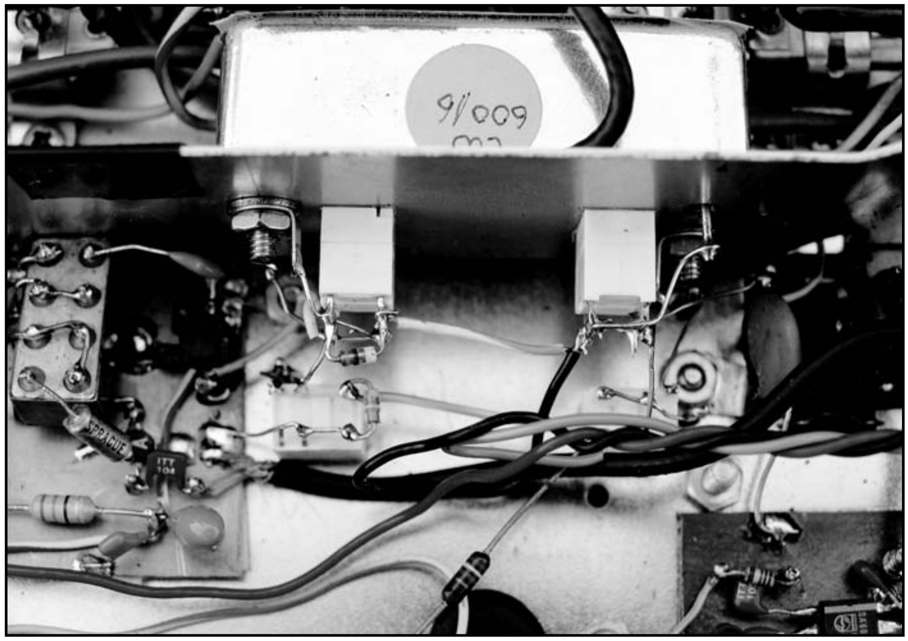
Front view of Harmanized R4C. The small toggle switch on the far left hand side selects the first IF crystal filter (either the Drake four-pole 8 kHz filter or the Sherwood six-pole 500 Hz filter).

The low-pass filter on the output of the Drake Permeability Tuned Oscillator (PTO) (C118, L3 and C119) has been removed from its original position and rebuilt on a small piece of PC board, which is mounted on the underside of the Harmanized R4C shown in the photo on the last page of this article.

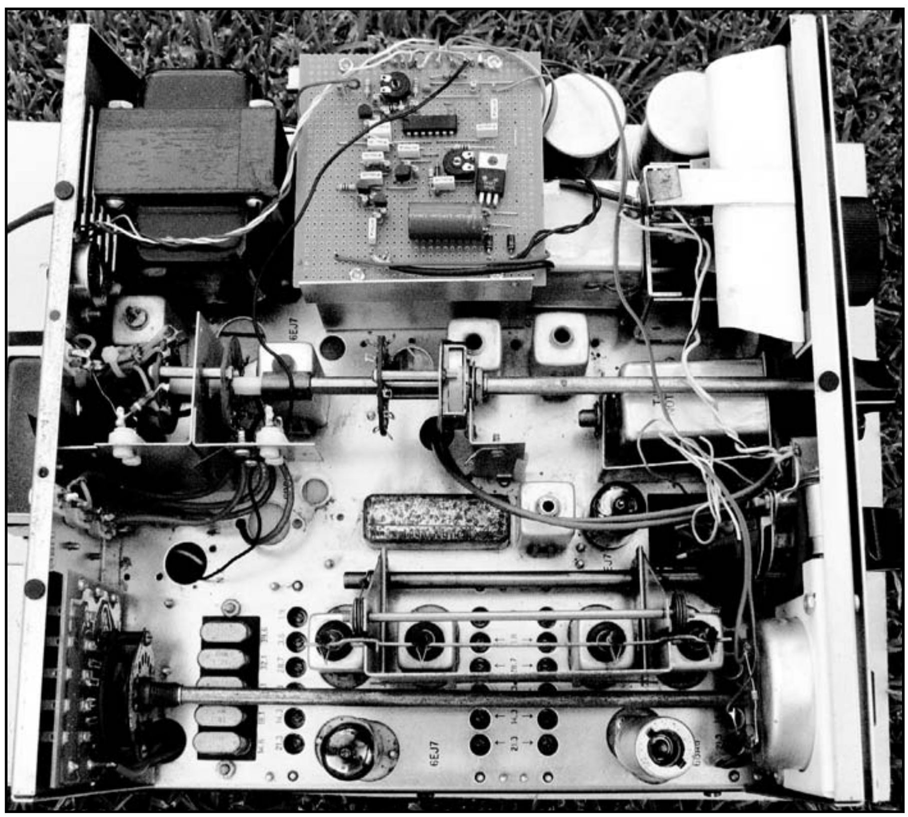
VK6VZ has high-resolution digital photographs of top and bottom views of the Harmanized R4C (showing construction of the majority of the circuit boards) that can be viewed on the *QEX* Download Files Web site (**www.arrl.org/qex/files/**).

Conclusions

The strong-signal performance of the Harmanized R4C is excellent, with its 95 dB blocking dynamic range at 2 kHz (with its Sherwood 600 Hz first-



Sherwood CF600/6 filter and relay switching.



Top view of the Harmanized R4C .

IF and Drake 500 Hz second-IF filters in circuit) keeping it free from cross modulation in even the busiest of CW contests. As a comparison, the Yaesu FT-1000MP—the most popular modern transceiver used by DXers and contesters—has a 2 kHz BDR of about 73.5 dB[7] with its 500-Hz IF filters in circuit some 20 dB less than the Harmanized R4C.

The 20 dB improvement in BDR can make a huge difference in successful weak signal DXing, particularly on 1.8 MHz, where local signals can be very strong. On his Web site, W8JI, widely acknowledged as *the* authority on 160 m reception, notes:

"Over a period, I've found 85 to 90 dB dynamic range to be about the most that is ever needed. In a simple [reception] installation with a single Beverage antenna, 80 to 85 dB IM3 and blocking dynamic range is probably enough. Receivers with less than 80 dB IM3 and BDR probably compromise a reasonably good station's capabilities."

Using a receiver like the Harmanized R4C with extremely low phase noise and a good BDR under CW contesting conditions is a revelation—you can truly tell the quality/distortion levels of the signals you receive (and some will be truly horrible!) and clearly hear the differences between good and bad quality signals.

You will also hear a gap—and a large drop in noise level/mush—between two good quality strong signals and truly experience what "good dynamic range" means.

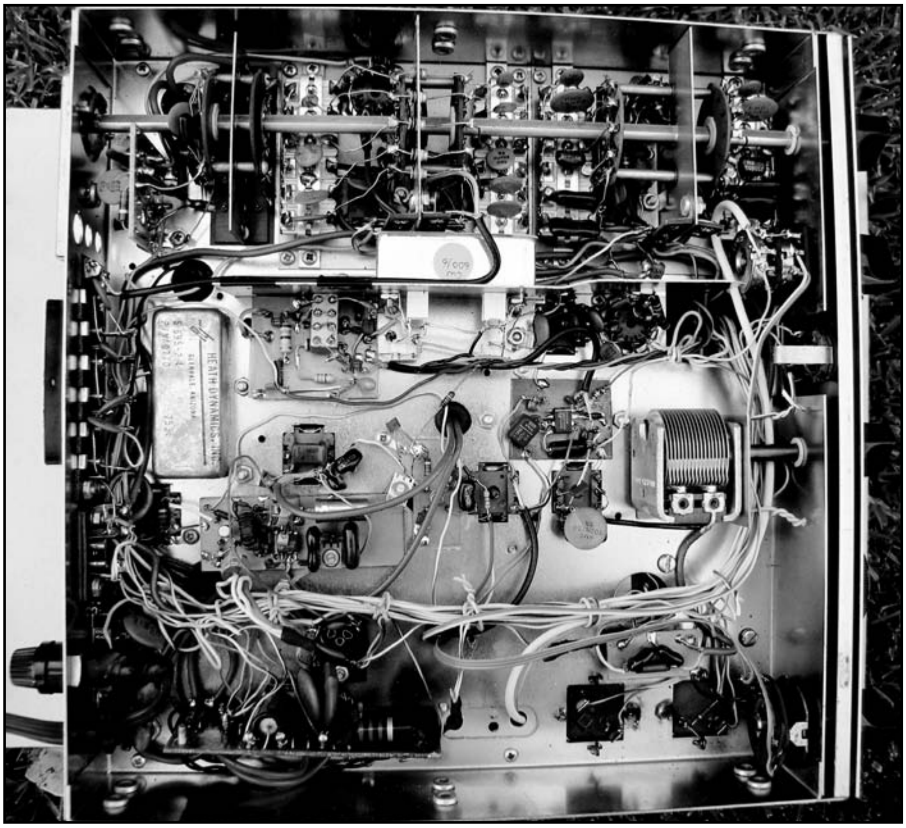
The AGC design of the original R4C made it sound very good on SSB—as one observer noted it almost seemed to "breathe"—but poor on CW. This was because of the long group delay caused by the R4C's crystal filters in the first and second IF stages, coupled with the AGC being applied before the second-IF crystal filters.

As a result, the delay of the original AGC was excessive, causing it to "overshoot." As W8JI says, "Any AGC arriving at the RF amplifier or first IF amplifier is several ms after the fact" when a narrow CW filter in the second IF is selected—the narrower the filter, the longer the group delay.

As stated previously, in the Harmanized R4C the AGC voltage is applied *after* the second-IF crystal filters and allows the gain of all the stages before the R4C's third mixer to run at maximum, eliminating the problem identified by W8JI.

The hang-AGC system designed by VK6APH improves on the performance of the original Drake R4C AGC on SSB, with the better readability and "ease on the ear" advantages that a hang system provides. The slower of the two CW positions will suit most contest and DX operation and offers a huge improvement in readability of CW signals over the original Drake R4C AGC.

While the old-fashioned ergonomics provided by the Harmanized R4C may not suit everyone, VK6VZ and



Bottom view of the Harmanized R4C .

VK6APH find its responsiveness to different signal levels make it much more pleasurable and less tiring to listen to than modern synthesized multiple-IF transceivers, whose AGC systems try to make all levels of signals sound S9 and are consequently very dynamically "flat."

Notes

[1]"Present Day Receivers—Some Problems and Cures" by J. Robert Sherwood, WBJGP (now NCØB), and George B. Heidelman, K8RRH, *Ham Radio*, Dec 1977. "New audio amplifier for the Drake R4C" by J. Robert Sherwood, WBØJGP, and George B. Heidelman, K8RRH. *Ham Radio*, Apr 1979.

[2]The Sherwood 600 Hz 6-pole CW filter for the first IF of the R4C is available from Sherwood Engineering (**www.sherweng.com**). For the SSB operator, LSB and USB filters for the R4C first IF are also available—see the Web site for further details.

[3]See "R4C heavy mod" in the table and Note 5 on the table in "Receiver Tests" at W8JI's Web site (**www.w8ji.com/receiver_tests.htm**).

[4]The article "Drake PTO: Smooth and Silky" by Neil A. Rosenberg, K1VY, can be found at **www.imiwebs.com/drakepto/**.

[5]"High Performance Receiver Add-Ons" by George Sartori, W5DA, *73* magazine, Jun 1979. The part of this article relating to the modification to the R4C 12 V dc supply can be downloaded from **www.mods.dk/view.php?ArticleId=2551**.

[6]International Radio ("Inrad"). See **www.qth.com/inrad/r4c-2if.htm**.

[7]Taken from the table "Receiver Tests" at W8JI's Web site, **www.w8ji.com/receiver_tests.htm**.

Photos by Ernie Harper, VK6TN.

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Steve Ireland has been a scientific journalist for over 20 years and is a former editor of the UK's Ham Radio Today *magazine. He was first licensed as G3ZZD at the age of 16 and emigrated to Australia in 1989, where he became VK6VZ. He has a BA honours degree in English from the University of North London (1982) and also holds an Ordinary National Certificate in Engineering (1975). His main interests are restoring vintage ham radio equipment, writing technical articles for ham magazines and low-band DXing. He has DXCC on 160 meters.* □□

8877 "Lite"—A 50-MHz 20-Pound Travel Amplifier

K5AND lets us in on how to break your score without breaking your back.

By Dick Hanson, K5AND

As those of you who do expeditions and mountaintop contesting can attest—carrying a kilowatt amplifier on your quest is *so* much fun. Over the years since 1987, I have built or modified five travel amps, all for 6 meters. The first was a highly modified Tokyo Hy-Power HLK-1A, followed by a Yaesu FL-2100, followed by a single 3CX800 (for W6JKV), followed by a pair of 3CX800s, followed by a single 8877 with separate power supply, followed by a single 3CX800 with built-in high voltage supply with a smaller (lighter) high voltage transformer.

These amplifiers all had one thing in common—they were a real challenge to transport. The last one was the best of the lot with respect to both footprint and weight, but still weighed in at 30 pounds, minus blower, plus 12 pounds for the luggage bag for a total weight of 42 pounds. I wanted a carry-on package that would weigh less than 30 pounds and, available wall voltage notwithstanding, an amplifier that would deliver 1500 W under normal conditions.

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Time for an Extreme Makeover?

It's pretty hard to "shrink" a conventional high-voltage power supply, even with careful sizing of the transformer, smaller filter capacitors, etc, but it is the power supply that offers the most potential for shedding pounds.

Ever since seeing a prototype regulated high-voltage switching supply in the Command Technologies booth at Dayton several years ago, I've dreamt about imbedding a high voltage switcher in an amplifier. Watts Unlimited was the first commercial entry into the high-voltage switching arena, so after months of studying the product and asking questions of its creator, I decided to try one.

Most "new" technologies have pros and cons, so you must evaluate both.

The "pro" side of the Watts Unlimited PS-2500A[1] is:

- At 10 pounds, it's a lightweight.
- It provides reasonable voltage regulation from idle to full-load (within about 10%).
- It's electrically *quiet*—this supply is only on during transmit.
- It's capable of supplying enough power for a 1.5 kW RF output amplifier,
- Instead of providing a fixed bias in the cathode (assuming triode tubes) and then changing the bias from cut-off to operate with a cathode

[1]Notes appear on page 46.

relay, in this amp, *there is no cathode relay*. Since there is no relay to switch from cut-off to operating bias, the tube will draw idle current as soon as the PTT circuit is activated. This eliminates at least several components from the amplifier.

"Cons" to keep in mind include:

- The supply cannot be turned on without a load. To do so would be to guarantee a failure of the output capacitors, as the resulting high voltage would soar over their 4 kV rating. The manufacturer recommends a minimum idle current of 150 to 250 mA at turn-on. So if you're going to use a Zener diode for the operating bias, you should select a Zener that will make the tube draw at least 150 mA at idle.
- Another thing to consider is that the supply does not come "on" when you turn on the ac power switch. It requires an external nominal 5-12 V dc source to switch it on. The user needs to provide a dc on-off source voltage activated by the PTT circuitry. While this is not an obstacle, you need to be aware of this requirement in your planning.
- The power supply includes two "fail safe" modes which will take the supply off line. Additional safety circuitry is highly recommended to prevent the amplifier from being keyed in the absence of anode voltage, which could damage the tube. This additional protection can take the form of a grid-trip circuit or an absence of B+ circuit.

Design Criteria

Let me say now that this is more of a *concept* article than a complete nuts and bolts, step-by-step how-to article. That said, if you have built an amplifier before, you should feel right at home with this material.

Because of my requirement for 1.5 kW RF output, and because of the regulation characteristics of the power supply, it seemed that a single 8877 would be the easiest tube to implement. Also given consideration were 3CX800s. For this application, the single 8877 best matched the power supply idle current requirement.

With the 8877 idling at 150 mA, the instant the supply is switched on, the anode voltage is 3400 V. At full load of 850 mA, the anode voltage drops to 3 kV. The power supply remains cool and electrically quiet while "dormant" even though 240 V ac is present. When the PTT is activated, the supply goes from 0 V to full output in a few milliseconds.

Features of the amplifier are:

- It fits in a relatively small box—6.5×13.5×12 inches (HWD, less blower). See Figs 1 and 2.
- It weighs a mere 20 pounds, including the built-in power supply.
- There are no protrusions from the cabinet to break in transit.
- There are no knobs, screwdriver adjustments are provided for the plate and load capacitors.
- A flush plug and socket is provided for the ac line entrance connections.
- Plate and grid meters are protected

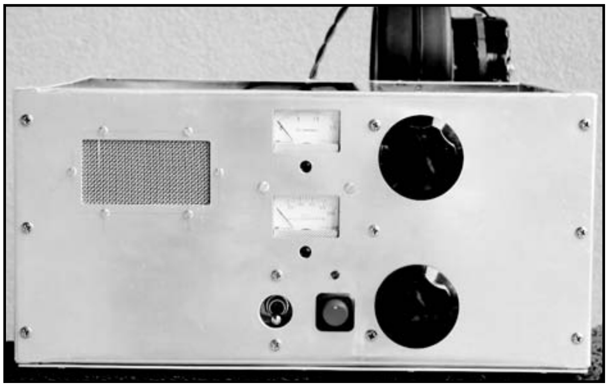

Fig 1—Front view, showing exhaust port on left, plate current meter on top at center, grid current meter at center, recessed power switch and amplifier IN/OUT switch with ground-trip LED above. On the top, the removable knob on the upper right is PLATE TUNE. The lower removable knob is PLATE LOADING. The air intake for the power supply is on the bottom of cabinet. The RF compartment is pressurized so that 95% of the air exhausts through the anode cooler and the rest exits thru the cathode compartment.

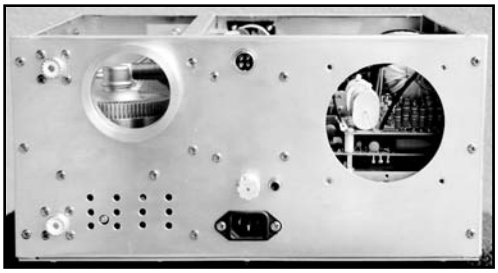

Fig 2—Rear view, showing RF input and output connectors at bottom left and top left, the mounting flange for the blower, cathode compartment air exhaust (under blower flange), ac power connector for blower and fan in upper center. The fuse, RCA PTT connector, ac mains connector and hole for the muffin fan at far right. Both openings are "screened" with stainless screening. Teflon chimney with rubber "extension" for pressurizing anode compartment.

Fig 3—The amplifier packed in its Travel Pro bag and ready to go.

behind the front panel.
- A 3 minute time delay holds off PTT until the cathode warms up.
- It is provided with built-in RF input/output relay switching.
- A vacuum plate capacitor and physically small meters help shrink the cabinet.
- It fits inside a rolling carry-on Travel Pro bag (8 pounds)—these bags are made of Teflon-coated ballistic nylon and are lightweight, durable, strong and water-resistant. Best of all, they have really great wheels and an extendable handle. See Fig 3.
- This amplifier may not win any beauty contests, but it must be rugged.

Building the Amplifier

For me, the hardest part of a project is planning the metal work, which of course is dictated here not only by the size of the components, but also by the airline size limitations for "carry-on baggage." Charlie Byers of Byers Chassis supplied the chassis and its interior metal work. I had some more metal work done by a local machine shop, and I used imbedded captive nuts for cabinet assembly rather than gambling on my fat fingers trying to get a nut on a screw. Details of the cover sheet metal work are shown in Figs 4 and 5.

Since this is a single band (in fact, almost single frequency) amp, it does not require much tuning. So if you can eliminate the protruding tuning knobs—why not? I made a couple of knobs with 1/4 inch shafts and then filed the ends of the shaft stubs to resemble a flat-blade screwdriver end. These assemblies are temporarily inserted into the plate and load panel bushings on the front panel so that you can in fact tune with a knob in "set and forget" fashion; the knobs are then removed for transport. The cathode tuning capacitor is also a screwdriver adjustment, and once set, will hold tuning over at least 1 MHz.

This amplifier uses a conventional π circuit instead of a π -L configuration to save a little more space. I always carry a high-power ICE 6-meter filter in the suitcase, so I have no worries about second harmonic rejection. The blocking capacitor is an "857" type, rated at 15 kV and a lot of current.

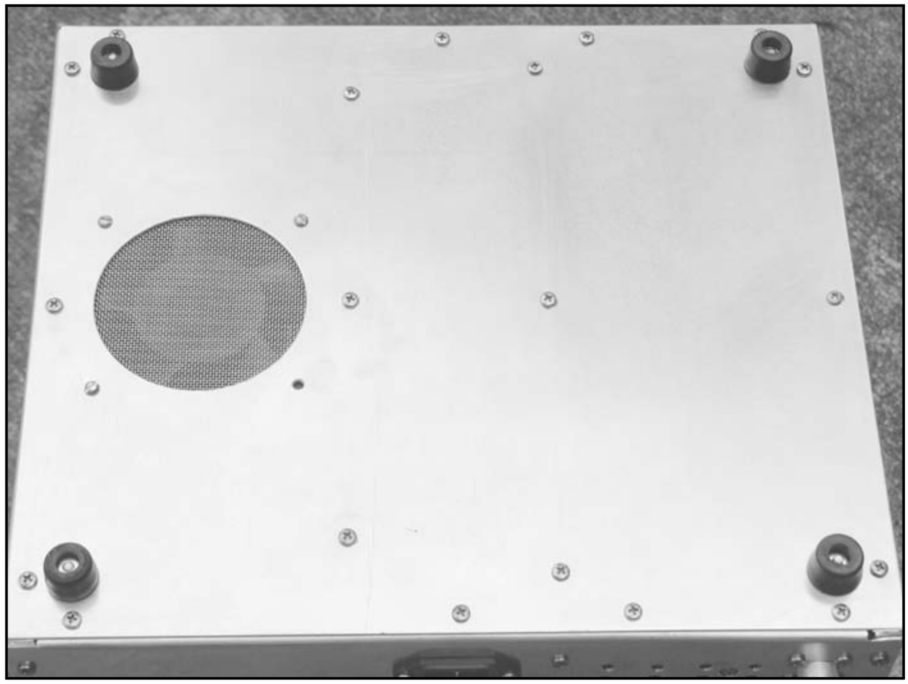

Fig 5—Bottom cover sheet metal.

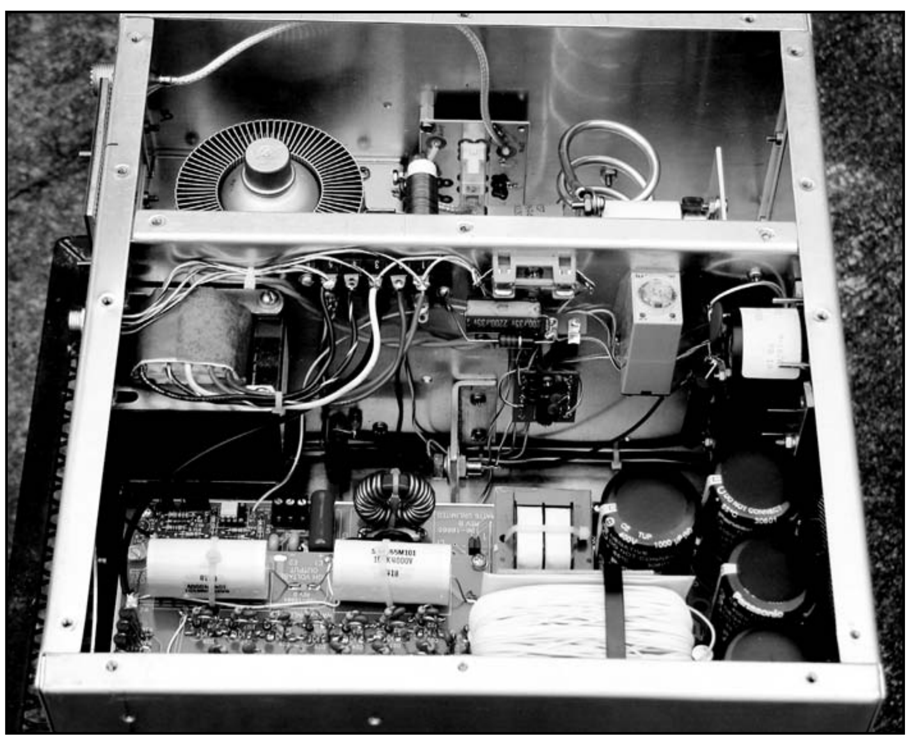

Fig 6—Another view of RF deck, showing the 8877, anode clamp and input/output relay-switching board, featuring the Schrack relays. Also note the Teflon RG-142 coax for the output. This coax will handle the 1.5 kW without meltdown! Also shown is part of the control circuitry—the Omron time delay relay and Tripplett 120G series plate and grid meters. The fuse at the center is for the amplifier cathode circuit. The filament/control transformer at the left is an Ameritron 406-1419-3J.

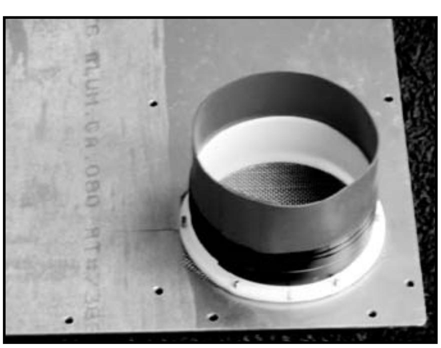

Fig 4—Top cover sheet metal detail including chimney mounting.

The RF input/output relay board is the same one I have been using for a number of years now. The SPDT Schrack relays have very robust contacts. The amplifier off-line VSWR may be "tuned out" using a thru-line capacitor ground trick shown me by Pat Stein, N8BRA, of Command Technologies and mentioned in an earlier article.[2] See Fig 6.

The three-minute time delay for cathode warm-up is provided by a common Omron H3YN-2, 0-10 minute delay relay; the delay is set for three minutes. It seems like an eternity when the band is open and the amplifier is warming up!

The power cord set is a heavy duty IEC arrangement featuring a standard three-prong male chassis mount connector mated to a 3-wire, 12 gauge power cord about 8 feet long. The plug end is left unterminated so that you can put on whichever male plug may be required for the country you're visiting. Again, this "connector" approach on the amplifier was used to eliminate protrusions which can be damaged in shipment. The ac mains are turned on and off by a recessed DPST switch rated at 10 A at 250 V ac. The filament and control voltage transformer is an Ameritron unit; PN 406-1419-3J; $49.95.

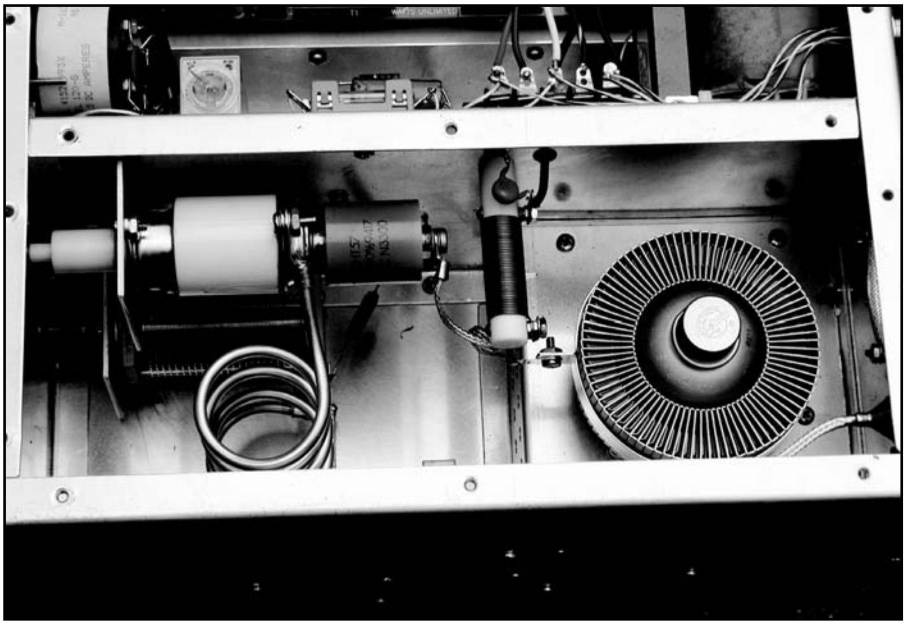

Fig 7—Larger view of the tank circuit, showing the silver-plated tank coil, 3 to 30 pF vacuum variable capacitor, 857 blocking capacitor, RF plate choke and, almost buried at the bottom, the 200 pF loading capacitor.

The plate RF choke is wound on a 1/2 inch Teflon rod, using 42 turns of 20 gauge Formvar insulated copper wire. The bifilar filament choke is 10 turns of 14 gauge Formvar insulated copper wire wound on a 2 inch long 1/2 inch diameter ferrite rod. See Fig 7.

The cathode choke is a Z-50 (7 µH) unit. Don't forget to install another Z-50 RF choke from the output loading capacitor terminal to ground to protect you from a shorted blocking capacitor. L1 and L3 are described in the RF deck section.

Tripplett 120-G series meters are used for the plate and grid metering since they are physically small and are high quality devices. They are protected from abuse by being mounted behind the front panel. The grid meter is a 100 mA unit and the plate meter is a 1 A unit.[3]

Grid over-current protection is provided by Q2 and its associated circuitry. The 5 kΩ potentiometer in the circuit allows the trip current to be set in the range of 40-150 mA, which is adequate for the 8877. When the preset current is reached, Q2 conducts, which in turn closes K4. One set of it's contacts locks the relay ON and turns on the TRIP LED; the other set of contacts opens the +15 V control bus to the PTT circuit, thus taking the am-

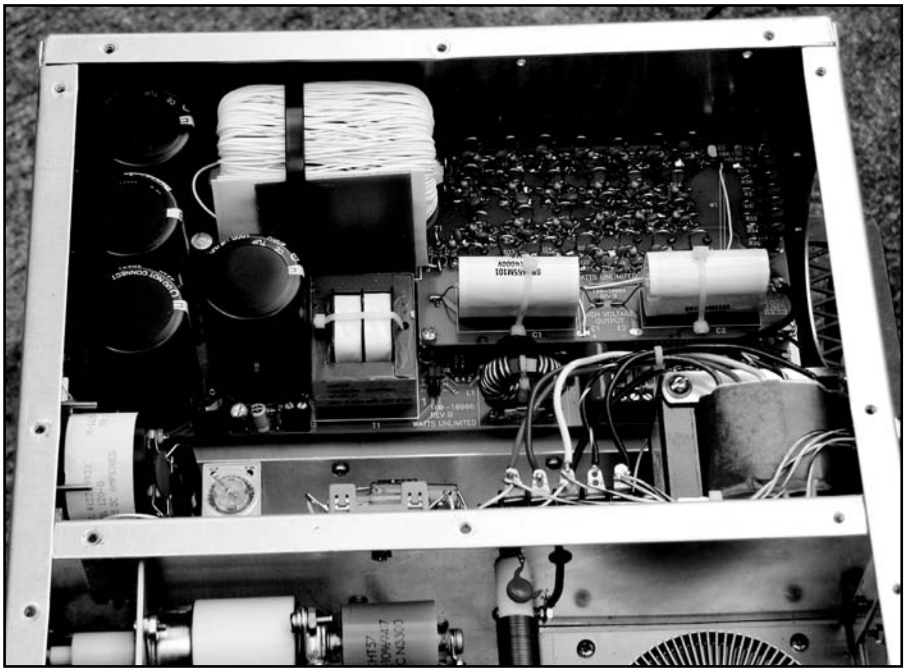

Fig 8—This shot shows the PS-2500A high voltage switching power supply. The white object at the top left is the high voltage transformer. The two high voltage output capacitors are at the center right. The total output capacitance is 0.2 µF at 4000 V dc.

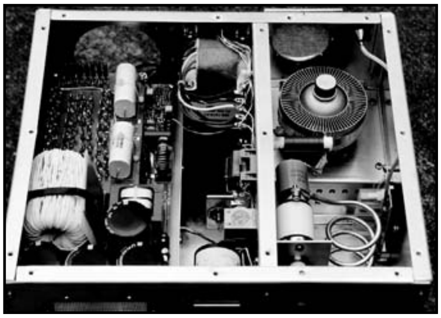

Fig 9—Top view, showing the Watts Unlimited PS-2500A switching power supply at far left. The center section houses the control circuitry, including filament transformer, time delay relay and metering. The RF compartment is on the far right.

plifier off-line until the over current condition has been resolved. Depressing the normally closed push-button switch resets the trip circuit.

To pre-set the grid-trip value, apply +5 V dc through a 500 Ω potentiometer to the junction of the grid and plate current meters. Adjust the temporary pot for some not to exceed value of grid current, say 100 mA for example. With the pot at maximum resistance, gradually turn it towards minimum until K4 trips and the TRIP LED comes on.

Specifications

There are no surprises here. With 40 to 50 W of drive, the amplifier will put out 1500 W with 2500 W dc input. The anode voltage is about 3000 V under a load of 850 mA. Grid current runs between 40-60 mA with this loading and anode voltage.

If you have an MFJ-259 or similar antenna analyzer, all the preliminary tuning can be accomplished before applying the high voltage. Getting the cathode and tank circuits "in the ballpark" before applying drive is always a good feeling.

Installation of PS-2500A Power Supply

The good news is that there are very few connections required to interface to this unit. The sobering news is that any mistakes are either costly, dangerous or both. Fortunately, the manufacturer has done a very nice job with documentation—the manuals are excellent. I would suggest taking extra time reading and understanding the manual especially regarding the following points (see Figs 8 and 9):

- AC connections, including neutral
- B minus
- HV turn on-off
- B plus connection
- Metering
- Jumpering

The supply factory default wiring is connected for 240 V ac mains as shown in Fig 10. The default jumpers are also set for the B minus connection tied to chassis ground. Most lin-

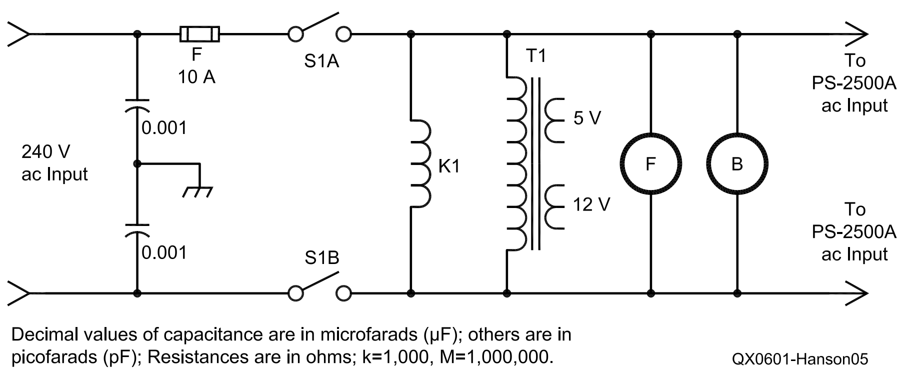

Fig 10—Schematic of power connections.

B—230 V 50 CFM blower.
F—230 V muffin fan.
K1—Omron DPDT 10 minute time delay relay.
S1—DPDT switch; 10 A at 240 V.
T1—Ameritron AL-1500 filament transformer, 230 V to 5 V at 10 A.

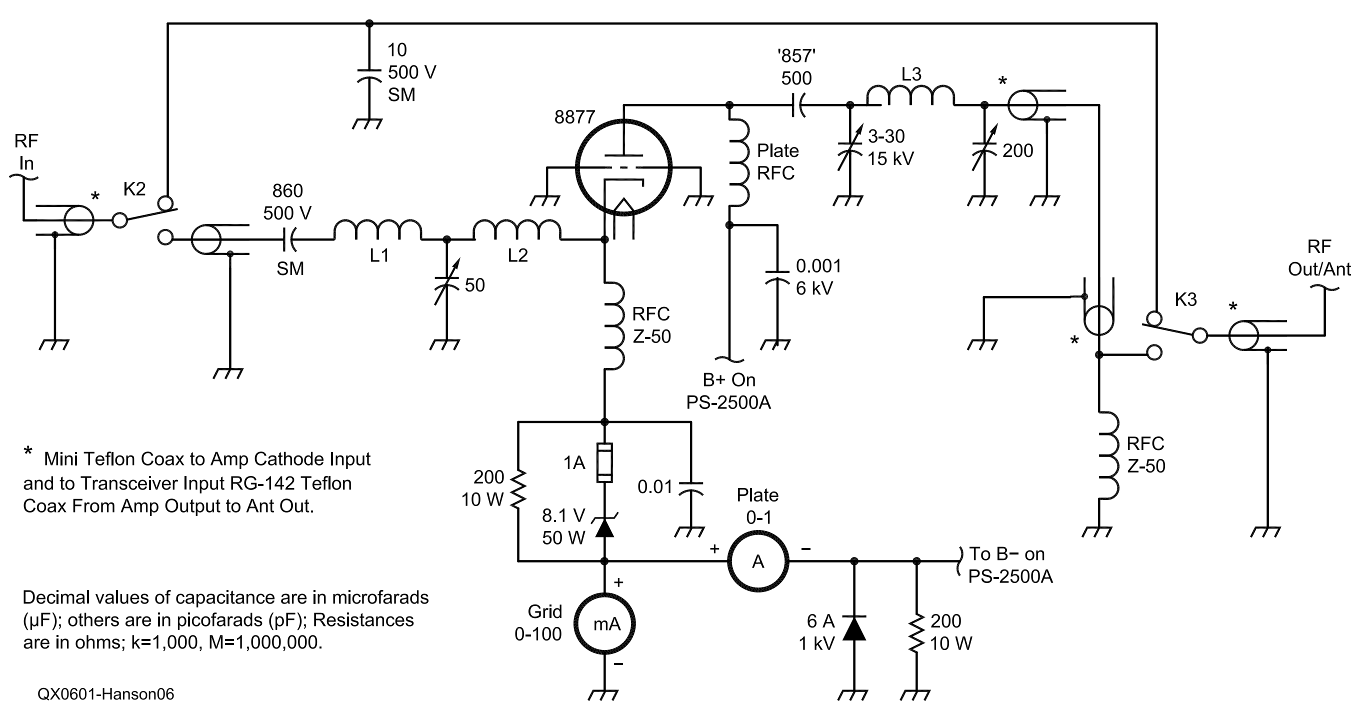

Fig 11—RF deck schematic. L1 and L2 are 9 turns 16 gauge, 1/2 inch diameter by 1 inch long. L3 is 4 turns, 3/16 inch tubing, silver plated 1 1/2 inch diameter by 3 inches long. The plate RFC is 12 turns 20 gauge Formvar close wound on a 1/2 inch teflon rod. The filament choke is 10 turns 14 gauge formvar each, bifilar wound on a 1/2 by 2 inch ferrite rod.

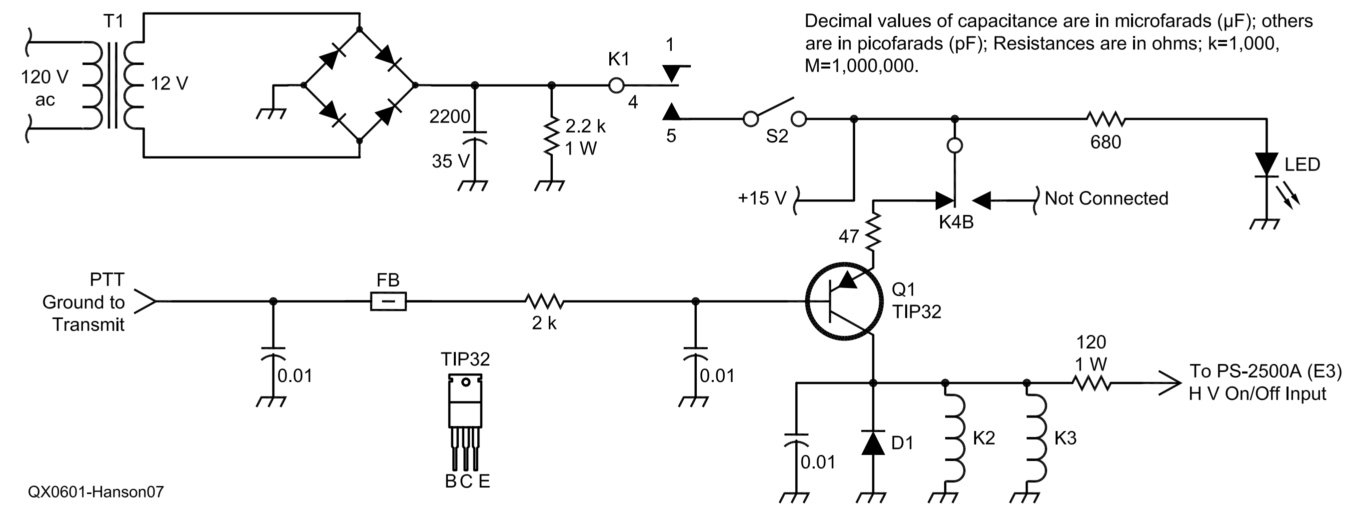

Fig 12—Schematic of control wiring. K2 and K3 are Schrack SPDT 12 V dc relays.

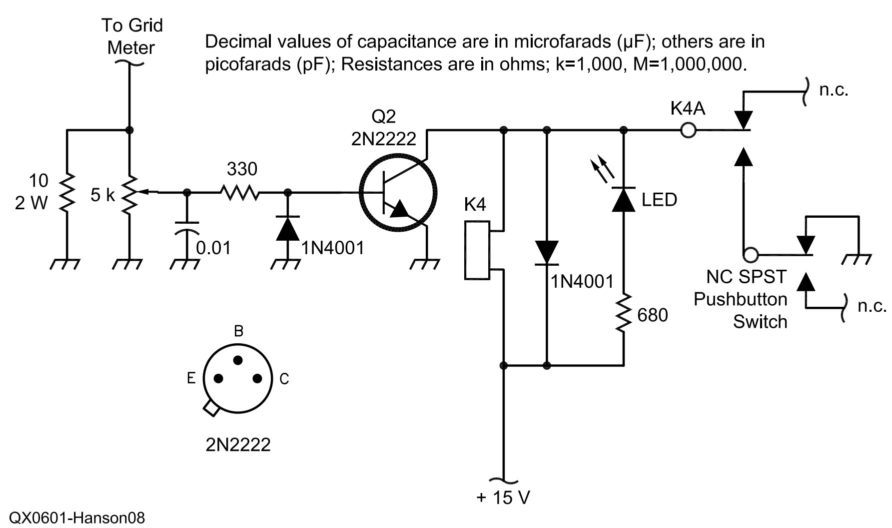

Fig 13—Schematic of optional grid circuit protection scheme. K4 is a DPDT 12 V dc relay

ear supplies have the B minus above ground by several hundred ohms to provide for metering. That is the option I chose. This means you need to remove the B minus jumper to ground. The negative terminal of the plate meter is connected to B minus through a 200 Ω, 10 W resistor in parallel with a 6 A 1 kV diode, both going to chassis ground. This affords both current measuring as well as protection for meters and power supply components.

I chose to "key" the supply on and off with +12 Vdc supplied by the PTT circuit.

Schematics and instructions for the power supply are not furnished here but are available for download from the manufacturer. An RF deck schematic is shown in Fig 11, control circuitry in Fig 12 and the optional grid protection circuitry is shown in Fig 13.

In closing, remember: *Failures on a mountaintop, or in some foreign country are easier to prevent than to repair* (ancient ham proverb).

Notes

[1]The PS-2500A may be obtained from: Watts Unlimited (Tim Hulick), 886 Brandon Lane, Schwenksville, PA 19473-2102; 610-764-9514; **www.wattsunlimited.com**.

[2]*2002 Proceedings of the Southeastern VHF Society*, p 254.

[3]The current metering subsystem is that shown in *The ARRL Handbook*, 2005 Edition, p 18.39.

Dick Hanson, K5AND, has been licensed since 1954. He holds a BA from the University of Texas and following 18 years with Hewlett-Packard in medical electronics has been the owner and CEO of Southern Staircase in Atlanta, Georgia. He has been building mostly VHF antennas and amplifiers for 30 years. He has lugged them to 19 DXpeditions, prompting the development described in this article. □□

Antenna Options

By L. B. Cebik, W4RNL

Do I Need More Gain?

On 2 meters, we find both horizontally polarized and vertically polarized antennas in keeping with the two main activity clusters on that band. At the lower end of the band, point-to-point communication dominates, along with some EME reflected-wave communication. In the main, these activities use horizontal polarization. In the upper part of the band, repeater and related mobile activities dominate, with a reliance on vertically polarized antennas. Near the middle of the band, we find a narrow frequency spread used for satellite communications. The variety of antennas used for this service tends to be a mixture, with some circularly polarized antennas. Similar patterns apply to most of the amateur VHF and UHF bands.

Unfortunately, many amateurs carry over HF antenna experiences into the VHF and UHF bands. Hence, there seems to be only one question that dominates poor results with an existing antenna: how do I obtain more gain? Longer Yagis and exotic antennas come to mind as the surefire answers to all inadequate communications problems. (We shall bypass the "more power" answer to the same question.)

For some situations, an antenna with more forward gain might be the answer. But higher gain may not always imply a longer Yagi with more elements. For many cases, the answer to our need for effective communication may lie elsewhere. In these notes, we shall review some information that is readily available but scattered. In the end, we may opt for more antenna gain, but only as a secondary feature of other antenna properties that we too often overlook. We shall confine ourselves to ordinary communication on 2 meters—as a band on which we can focus attention and make comparisons. Extreme terrestrial DX and EME communications will have to be topics for another day.

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Height

Sometimes the answer to an antenna problem is not gain but height. There are two immediate clusters of reasons for needing more height.

Local Clutter: Local clutter consists of all objects that may block, absorb, refract and reflect RF energy so that it cannot reach its target. Fig 1 provides a simplified sketch of the situation. Both organic and inorganic structures can get in the way of RF energy that we want to reach a certain station, regardless of whether we are using a vertically or horizontally polarized antenna. Trees and shrubs vary in their energy deflection abilities, depending on local weather, type of flora, and the season. Wet wood is usually more ionized than dry wood. Some species of trees have a higher metallic content than others.

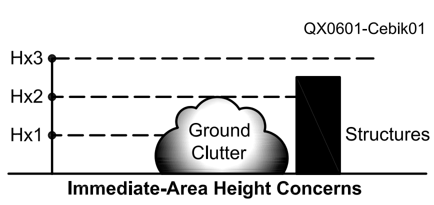

Fig 1—Some immediate-area reasons for needing more antenna height.

The more modern a human-built structure, the more likely it is to have more metal that can block our communications attempts. An old Victorian mansion might have a single ac circuit and wires in the floor of the upper story. A modern house has computer, telephone, TV and other cables adding to extensive house wiring that likely comes down the walls from an attic area. As well, foil-lined attic insulation is common. Finally, structural steel is working its way into the modern home, not only as support beams, but also as a replacement for the traditional 2-by-4. We need not move to a high-rise apartment structure to be surrounded by metal.

In these kinds of situations, raw antenna gain may not help us much at all. As the figure suggests, we need to place our antenna above the clutter. If the clutter is at a moderate distance from our antenna location, we can sometimes move the antenna site laterally to clear most of the problematical signal deflections. However, in most cases, nothing succeeds like height. We shall return to the local-area height question, but first, we should note the second category of concerns.

The Horizon: How close the radio horizon is to us often surprises newer VHF and UHF operators. Since VHF and UHF communication normally is line of sight (and just a little more), the height of the antennas at both ends of the line determines how far apart we may be and still effectively communicate. Fig 2 shows the most basic outline of the situation. If the two antennas have different heights, each will have a different distance to the grazing point, that is, the point at which the RF energy encounters the ground. We can estimate either D1 or D2 from a standard equation (shown in *The ARRL Antenna Book*, 20th Ed., p 23-6).

$$D_{mi} = 1.415\sqrt{H_{ft}} \quad \text{(Eq 1)}$$

The equation shown is actually a short form of a slightly more complex equation, a version of which appears in *Reference Data for Engineers*, 8th Ed., p 33-14.

$$D_{mi} = \sqrt{1.5KH_{ft}} \quad \text{(Eq 2)}$$

The equations are easily converted from miles of distance and feet of height to kilometers and meters. However, let's look more closely at the new element in the second version of the equation. K is the effective Earth radius. The value is about 1.333 for the temperate latitudes, but may vary from 0.6 to 5.0 depending on where in the world we may be. The simplified equation produces accurate results only for the latitudes in which the value of K is 4/3.

We may easily turn the equation around to see how high an antenna must be for a given distance to the radio horizon.

$$H_{ft} = \frac{D^2_{mi}}{1.5K} \quad \text{(Eq 3)}$$

Lest we think that achieving a significantly greater distance to the radio horizon is a linear matter of raising the antenna by so many feet, notice the fact that the equation uses the square root of height in finding the distance. Doubling the antenna height will only increase the distance to the radio horizon by a factor of 1.4. Table 1 correlates some common antenna heights and the distance to the radio horizon.

Of course, the distance to the horizon (D1) is not the distance to the most remote distant station that we can contact. That station will also have an antenna height and resulting distance (D2) to its radio horizon. So the actual communications distance is D1 + D2. Atmospheric bending of signals may add perhaps 10% to the raw calculation. However, broadcast antennas (for example, FM or television) usually add a term to their height plans so that station signals clear the Fresnel zone, the region in which diffraction from objects in the signal path may yield interfering waves. Nevertheless, for point-to-point communications, the average amateur is severely limited in efforts to increase the communications range by adding more tower sections. Little wonder that FM repeater work relies heavily on placing the repeater antennas on the highest tower with space for rent or donation.

In most cases, achieving enough antenna height to clear most local clutter takes precedence over adding to the antenna gain. Only if we can establish at least marginal communications with the desired target station will added antenna gain provide significant signal strength to convert marginal signals into reliable ones.

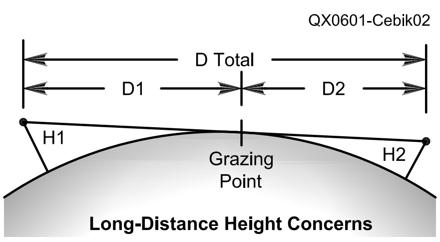

Fig 2— A distant reason for needing more antenna height.

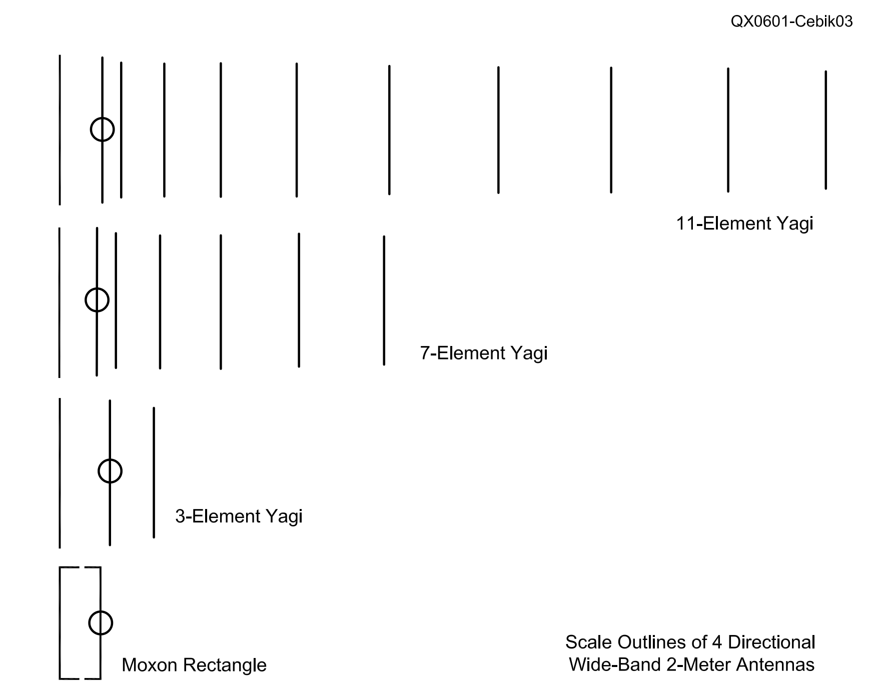

Fig 3—Outlines of 4 antennas that we shall use as samples. Drawings are to scale.

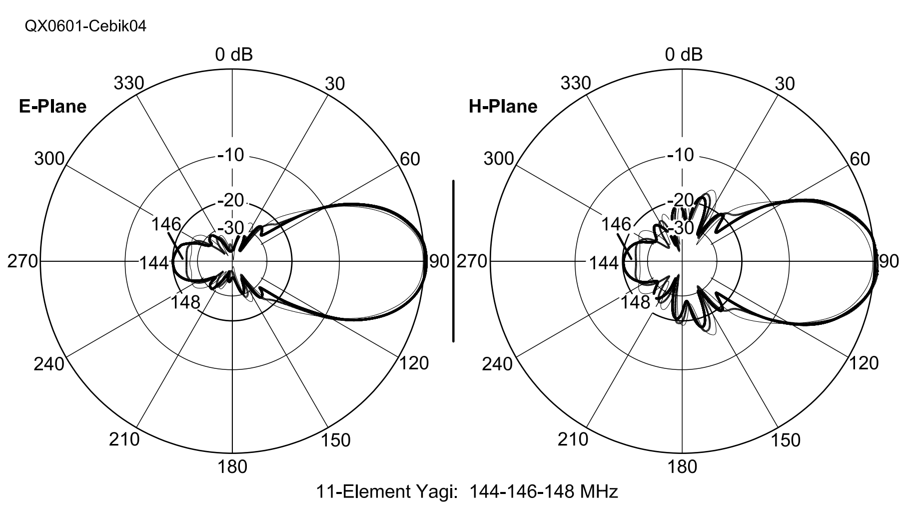

Fig 4—Free-space E-plane and H-plane patterns of the 11-element Yagi.

Gain and Beamwidth

We often overlook an important property of antennas in our quest for maximum gain and front-to-back ratio. Antennas also exhibit a beamwidth that can be very useful to us in obtaining the coverage that we desire and the blocking of signals from undesired directions. We shall deal eventually with all four of the antennas in Fig 3, but initially we shall look most intently at the three Yagis in the group. The outlines are to scale. Table 2 lists the dimensions for the Yagi antennas in inches. The dimensions assume that the beams use either a nonconductive boom or that the elements are well insulated and isolated from a conductive boom. The *EZNEC* models of these antennas are available from the ARRL Web site.[1] The models use a free-space environment and the elements will form a horizontally polarized antenna if we place them over ground. However, you may rotate the elements to create a vertically polarized antenna and adjust the array height as desired.

All three Yagis are credible performers for their boomlengths and number of elements, and all use ½" diameter elements for uniformity. I have selected Yagis that cover the entire 2-meter band so the comparisons are fair throughout. The change of gain across the band is minimized, and the front-to-back ratio is at least 20 dB for all models at all frequencies sampled. The two larger Yagis have feedpoint impedance values very close to 50 Ω. The 3-element version has a natural driver resonance between 25 and 30 Ω. The model uses a ¼-λ matching section to bring the model source impedance above 40 Ω at the design frequency (146 MHz). The 36-Ω transmission line can be composed of parallel sections of 72-Ω coax cable. Should you wish actually to construct the antenna, you may also shorten the driver and add a beta (hairpin) matching component. In all cases, the beams cover the band with under 2:1 50-Ω SWR. Table 3 lists the modeled performance of the Yagis at 144, 146, and 148 MHz in free space to provide some fundamental data.

We should first review some common information about Yagis with different lengths. The most significant physical fact is that the boomlength tends to increase faster than the element count. The 7-element Yagi has 2.3 times the number of elements as the 3-element Yagi, but the boom is 3.4 times

[1]See **www.arrl.org/qexfiles/1X06_AO.zip**.

Table 1—Some calculated distances to the radio horizon based on antenna height.

Antenna Height (feet)	*Distance (miles) to the radio horizon*
10	4.5
20	6.3
30	7.8
40	8.9
50	10.0
100	14.2

Table 2—Dimensions of 3 Yagi antennas discussed in the text.

All antennas use 0.5" diameter elements and presume a non-conductive boom. The "boom" dimension indicates the cumulative spacing, and the "element" dimension is the total element length, both in inches.

Element	*3-Element Yagi*		*7-Element Yagi*		*11-Element Yagi*	
	Boom	*Element*	*Boom*	*Element*	*Boom*	*Element*
Reflector	0	40.12	0	40.73	0	39.77
Driver	13.39	38.00	10.14	39.01	11.53	38.73
D1	25.10	34.42	14.66	35.81	16.08	35.98
D2			26.49	35.16	27.66	35.32
D3			42.49	35.24	42.58	35.39
D4			62.99	35.05	62.66	35.21
D5			85.37	33.04	87.08	34.23
D6					115.77	33.35
D7					145.53	32.67
D8					176.45	31.99
D9					202.31	31.20

Table 3—Modeled (NEC-4) wide-band performance of the 3 sample Yagis.

Gain values are for free space, and the front-to-back ratio gives the 180° value. The E-plane is in plane with the elements, and the H-plane is perpendicular to them.

3-Element Yagi			
Frequency MHz	144	146	148
Gain dBi	7.63	7.77	7.97
Front-Back Ratio dB	21.68	36.36	24.93
E-plane Beamwidth degrees	65	64	63
H-Plane Beamwidth degrees	108	107	103
Impedance (R ± *j*X Ω)	35.1 + *j*6.0	41.1 – *j*2.2	41.5 – *j*17.1
50-Ω SWR	1.46	1.22	1.52
7-Element Yagi			
Frequency MHz	144	146	148
Gain dBi	11.53	11.63	11.48
Front-Back Ratio dB	22.60	30.66	21.62
E-plane Beamwidth degrees	48	46	44
H-Plane Beamwidth degrees	57	55	52
Impedance (R ± *j*X Ω)	41.6 + *j*1.5	47.2 + *j*6.8	51.5 – *j*5.2
50-Ω SWR	1.21	1.16	1.11
11-Element Yagi			
Frequency MHz	144	146	148
Gain dBi	14.06	14.18	13.92
Front-Back Ratio dB	20.21	24.46	25.31
E-plane Beamwidth degrees	38	37	36
H-Plane Beamwidth degrees	42	40	39
Impedance (R ± *j*X Ω)	43.0 – *j*4.1	45.1– *j*0.8	48.7 – *j*10.2
50-Ω SWR	1.19	1.11	1.23

longer. The 11-element Yagi has 3.7 times the elements of the 3-element Yagi, but on a boom 8.1 times longer. Gain does not keep pace with the increases in either the element count or boomlength. Moving from 3 to 7 elements nets us a 3.9-dB gain increase, but adding a similar number of elements to wind up with 11 nets us only about 2.6-dB additional gain.

Next, let's examine some data that many VHF Yagi users overlook. The data tables include the E-plane and H-plane beamwidth values: that is the number of degrees between the half-power points on the radiation pattern. The E-plane beamwidth is very close to what we would obtain operating the beam horizontally over the ground, while the H-plane value is close to what we can expect for beamwidth when operating the antenna vertically over ground.

If we begin with the 11-element Yagi, shown in Fig 4 for all three sampled frequencies in both planes, we find no great difference between the E-plane and H-plane beamwidth values—about 3°. However, if we shorten the beam to 7 elements, as shown in the overlaid patterns of Fig 5, the beamwidth difference grows to 9°. When we clip 4 more elements from the Yagi, as in the patterns in Fig 6, the difference between the E-plane and H-plane beamwidth values climbs to about 43°.

To make the information even more graphic, Fig 7 overlays all 3 beam patterns in each plane for 146 MHz. The E-plane patterns show the gain increases with boomlength. However, the beamwidth decreases by only 27° as we move from 3 to 11 elements. In the H-plane, the differential in beamwidth values is 67°. When using the Yagis over ground in the horizontal position, the change in beamwidth usually signals only greater ease or difficulty in aiming a rotatable installation. In contrast, the much larger change in beamwidth presents us with some interesting potentials for vertically oriented Yagis over ground, even if we create a fixed installation.

Before we explore that potential, let's add one more beam to the collection, the Moxon rectangle. This compact 2-element array has some interesting properties in conjunction with the Yagis in our collection. Table 4 lists the dimensions, with reference back to Fig 3, which shows the outline of the beam. Dimension A represents the two parallel long element sections. B is the driver tail length, while D is the reflector tail, where "tail" indicates the

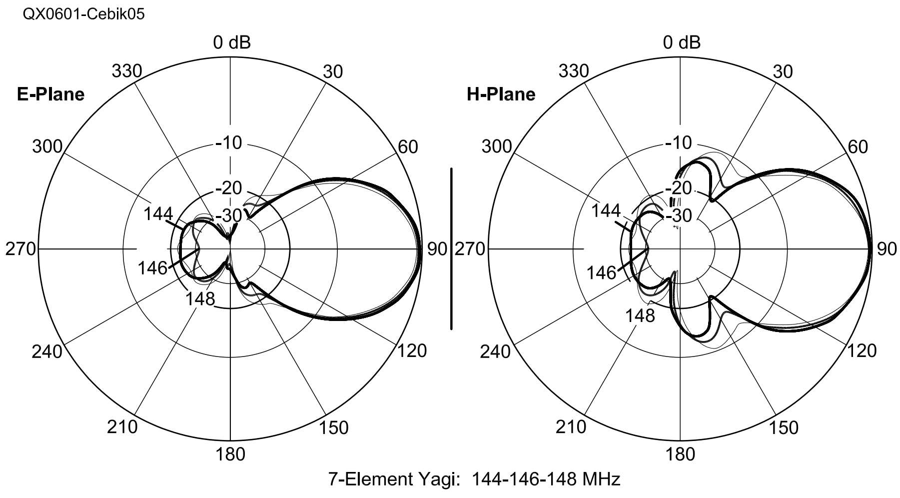

Fig 5—Free-space E-plane and H-plane patterns of the 7-element Yagi.

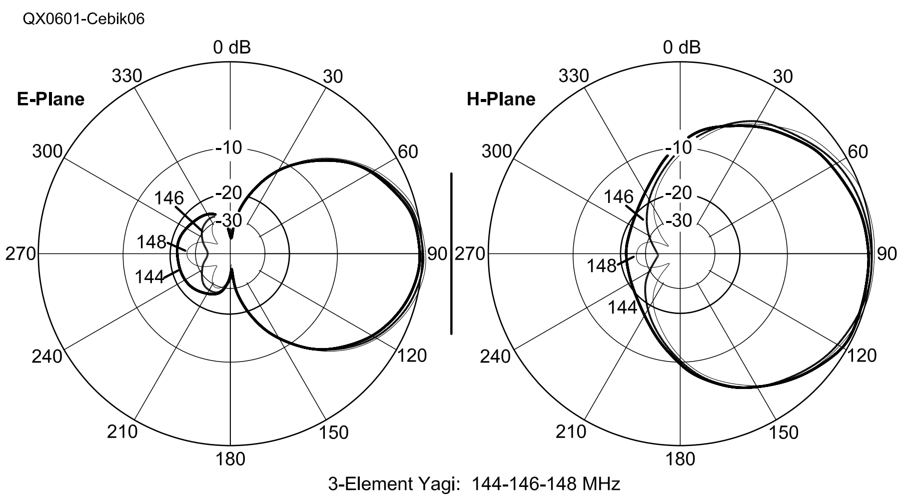

Fig 6—Free-space E-plane and H-plane patterns of the 3-element Yagi.

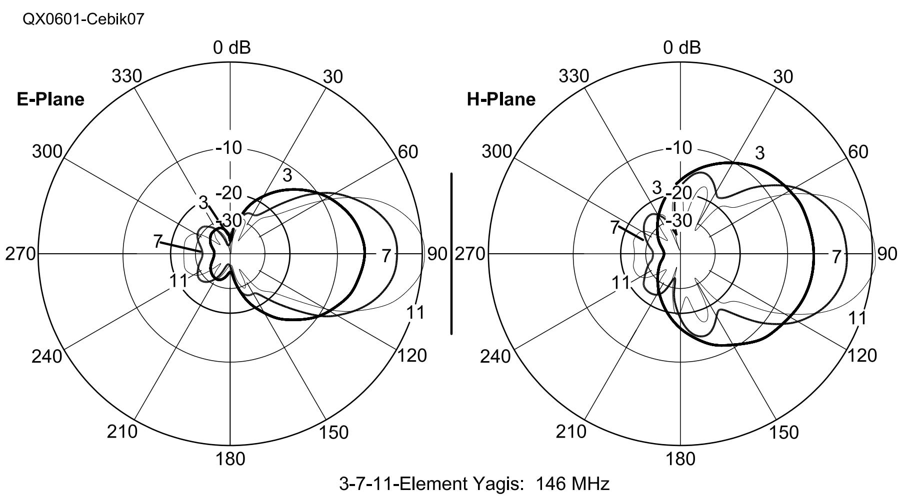

Fig 7—Comparative free-space E-plane and H-plane patterns of the 3 Yagis at 146 MHz.

Table 4—Dimensions of a 2-meter Moxon rectangle using ¼" diameter elements.

See text for an explanation of the dimension letter designations. Dimensions are in inches.

Dimension	*Length*
A	28.98
B	3.71
C	1.48
D	5.60

Table 5—Modeled (NEC-4) wide-band performance of the Moxon rectangle.

Gain values are for free space, and the front-to-back ratio gives the 180° value. The E-plane is in plane with the elements, and the H-plane is perpendicular to them.

Frequency MHz	*144*	*146*	*148*
Gain dBi	6.26	5.94	5.64
Front-Back Ratio dB	20.54	35.18	19.55
E-plane Beamwidth degrees	78	79	79
H-Plane Beamwidth degrees	134	144	153
Impedance (R ± *j*X Ω)	41.9 – *j*11.6	53.2 – *j*1.7	62.9 + *j*5.8
50-Ω SWR	1.36	1.07	1.29

portion of the elements that point toward each other. Dimension C is the gap between tails. The sample Moxon uses ¼" diameter elements.

Table 5 provides the modeled performance data for the Moxon rectangle. The gain is modest, but the front-to-back ratio is very high for a 2-element wide-band beam. The direct 50-Ω feedpoint covers the entire 2-meter band easily. More significant perhaps are the patterns in Fig 8. The E-plane beamwidth is 79° at mid-band, about 13° wider than for the 3-element Yagi. However, the greatest growth in beamwidth occurs in the H-plane pattern, which is 34° wider than the corresponding 3-element Yagi pattern. The cardioidal pattern provides a 144° beamwidth at 146 MHz.

Where we do not require much gain, the Moxon rectangle provides some very promising potentials. Horizontally, the antenna makes a good field unit that we might hand-steer without much difficulty. However, the final promise for the antenna may well lie in its service as a fixed vertically polarized antenna for certain types of home-station repeater operations.

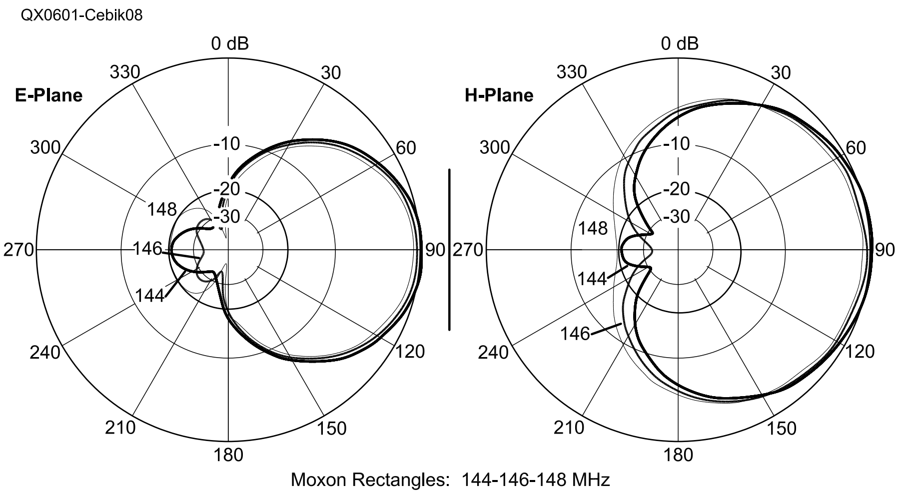

Fig 8—Free-space E-plane and H-plane patterns of the Moxon rectangle.

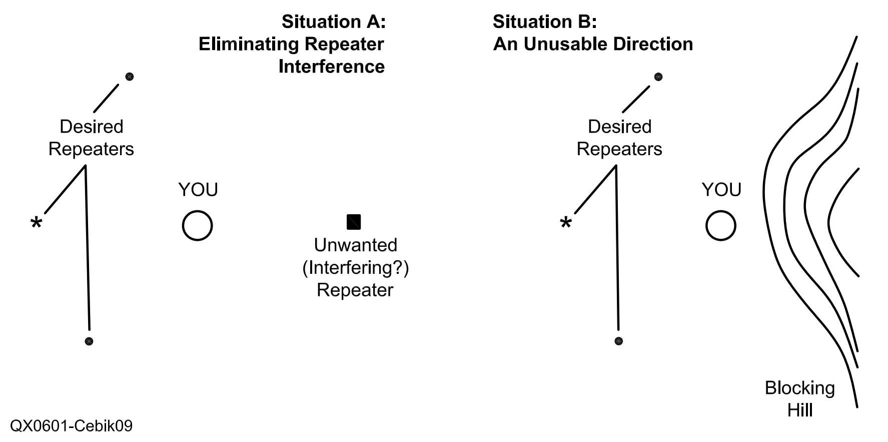

Fig 9—Two reasons for using a directional vertically polarized antenna.

Beamwidth as a Primary Property

At many locations, omni-directional antennas for repeater operations may be the wrong choice. Fig 9 presents only two such scenarios. The left portion of the figure shows a situation in which there may be a repeater station to the east that either causes interference or which we simply do not wish to access while working with one or more of the repeaters to the west. The situation calls for a directional antenna with a reasonably good front-to-back ratio to reduce signals to and from the east. At the same time, the beamwidth of the antenna should be wide enough to permit contact with the entire set of repeater stations to the west.

On the right, we have a hill or other major terrain obstruction that prevents operation to the east. In this case, we might use an omni-directional antenna. However, such an antenna will provide lower gain. As well, reflections from the hill may create interference patterns. The use of a directional antenna with the proper pattern shape would permit us to control the reflections in a useful way.

We have surveyed four different antennas in terms of the H-plane beamwidth. Which one will serve best in a given application depends on the operational needs. Fig 10 presents overlaid H-plane patterns for all four antennas at 146 MHz. There are four different sets of needs indicated by the dots or locations of the desired repeater stations. (Note that, although I am working in terms of repeaters, any vertically polarized station within the active field may be included.)

The upper left situation shows four widely spaced stations, two of which fall outside the beamwidth of any of the Yagis. If gain is not a major consideration, then the Moxon rectangle may best fulfill this need. Indeed, one need not use this antenna design only when there is a need for a very wide beamwidth. The upper right portion suggests a situation calling for somewhat more gain but a lesser demand on beamwidth. Hence, the 3-element Yagi may be the antenna of choice. The two lower scenes present calls for still higher gain and decreasing beamwidth requirements. The 7- or 11-element Yagis may best fill these roles.

The lessons from this exercise are two. First, for any proposed antenna, we should understand not only the potentials for gain and front-to-back ratio. We should also understand both the E-plane and H-plane beamwidths. Second, once we understand the beamwidth capabilities of antennas, we may select one that will provide us with designer-coverage for a given communications situation. In many cases, we may avoid the expense and maintenance worries of using a rotator by selecting the right antenna for the desired field of coverage.

More Gain, Same Beamwidth

So far, we have resolved antenna issues by relying on antenna properties other than forward gain. There may be cases in which we need height, beamwidth and additional gain. As we saw in comparing the Yagis, the higher the gain potential, the narrower the beamwidth. Suppose we need to meet the demands of one of the top two scenes in Fig 10, but with more gain than we can obtain from a single Moxon rectangle or 3-element Yagi.

The answer does not lie in making a longer Yagi. For every increment of boomlength that we add, we lose a proportional amount of beamwidth. For vertically polarized Yagis, the rate of decreasing beamwidth is greater than for horizontally polarized Yagis. Of course, we can accept the narrower beamwidth and resort to the rotator. However, we should first explore a strategy that allows us to enjoy the simplicity of a fixed installation.

One effective strategy is to use a 2-stack of whichever antenna we select. The mechanical trade-offs between a longer-boom Yagi with 3-dB higher gain and a stack of two shorter Yagis are about even at 2 meters and above. To create a stack, we shall have to extend the mast by about 7' (1 λ), but the individual antennas will place less stress on the mast than a single Yagi with a boom perhaps 3 times as long. Since we would have to rotate the longer Yagi to cover the same field, let's try the stacking route.

A good separation between vertically polarized stacked beams is about 1 λ, center-to-center (or, for reference, feedpoint-to-feedpoint), as shown in the outlines in Fig 11. Since all of the feedpoint impedances are 50 Ω, you may use a pair of 75 Ω cables, one from each feedpoint to the midpoint between them. Each line should be an odd multiple of ¼ λ at about 146 MHz (taking the velocity factor of the line into account). The two resulting 100 Ω impedances in parallel match the main 50 Ω cable. Other schemes are possible, but this one is time-tested. Phase-line losses require the use of the best quality coax that you can afford.

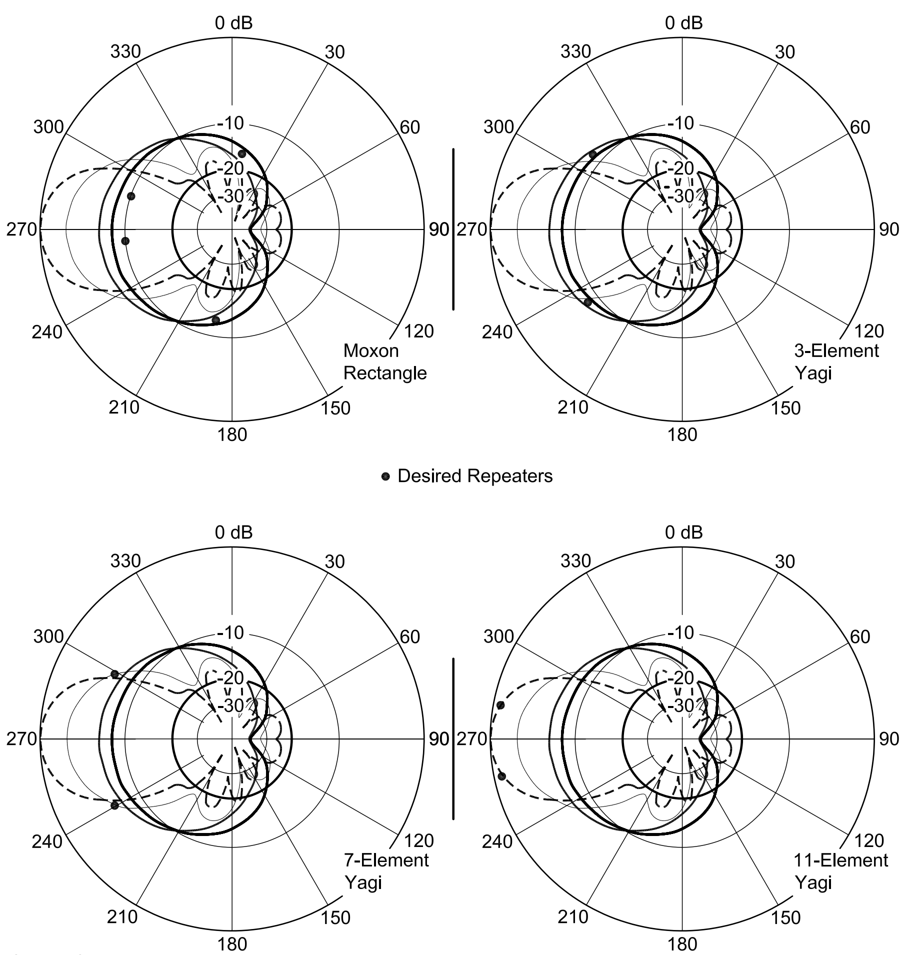

Fig 10—Four repeater scenarios and 4 antenna solutions.

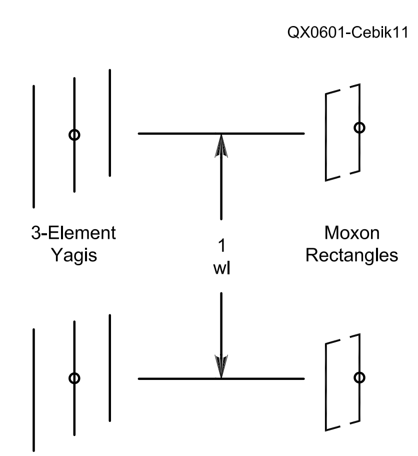

Fig 11—The outlines of 2-stacks of 3-element Yagis and Moxon rectangles.

Table 6 provides comparative information on the modeled performance of the two types of antennas, with information on single-unit and 2-stack versions. With a 1-λ separation between the two antennas in each stack, we net about 3.2-dB of added gain over single-units at the lower height. As well, the 2-stack has a slightly lower TO angle, and that fact has positive implications for point-to-point use of the array. The antennas are each far enough above ground and far enough apart that the feedpoint impedance values for the 2-stack do not significantly change relative to the impedance of a single unit.

For the present discussion, perhaps the most important fact is that neither beam loses any horizontal beamwidth when placed in the 2-stack just described. Fig 12 compares the single-unit and 2-stack azimuth patterns for each antenna type. 2-stack gain in each case marks the outer limit of the patterns. Hence, the Yagi may give the illusion of showing a pattern that is smaller in area. However, the real difference lies in the narrower beamwidth for the higher-gain Yagi, relative to the Moxon rectangle.

Our exercise presumed that we needed vertical polarization for the desired communications. It also set up scenarios in which we wanted to null out a general direction and direct our transmitted energy (and our receiving sensitivity) over part or all of the remaining horizon. The techniques that emerged gave priority to the H-plane beamwidth of the antenna candidates in devising a way to meet the need. Gain became a secondary property. Within the limits of the scenario, attaining more gain required methods other than simply making a longer Yagi with more gain.

Table 6—Comparative performance at 146 MHz of a single unit and of a 2-stack, using vertically polarized 3-element Yagis and 2-element Moxon rectangles.

2-stacks are separated by 1 λ center-to-center. Base height is 20', with the second antenna 7' higher.

3-Element Yagi	*Single Unit*	*2-Stack*
Gain dBi	11.46	14.64
TO angles	4.2°	3.6°
Front-Back Ratio dB	36.04	31.63
H-Plane Beamwidth	107°	108°
Impedance (R ± *j*X Ω)	41.1 – *j*2.2	40.8 – *j*3.2 (x2)
Moxon Rectangle	*Single Unit*	*2-Stack*
Gain dBi	9.64	12.90
TO angle	4.3°	3.7°
Front-Back Ratio dB	34.58	25.90°
H-Plane Beamwidth	144°	147°
Impedance (R ± *j*X Ω)	53.3 – *j*1.6	54.5 – *j*3.5 (x2)

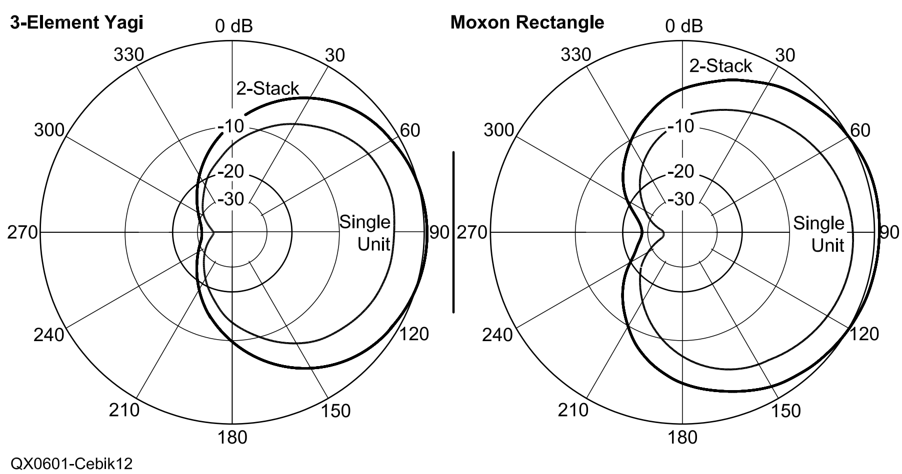

Fig 12—Single-unit and 2-stack azimuth patterns at a 20' base height for vertically polarized 3-element Yagis and 2-element Moxon rectangles. Base height of the upper antenna is 27'.

Conclusions

At VHF and UHF, there are numerous communications activities that call for the highest gain possible. Long-distance point-to-point work may call for stacks and squares of the longest-boom antennas feasible installed as high as possible. EME work may call for similar antennas, but height is less of a problem, since we shall point the antennas upward. Still, the quest for gain rules these activities.

However, we may fail to meet most basic communication needs if we only think of gain, and especially, if we think of gain only in terms of longer Yagis with more elements. The first task is to employ all of the ingenuity at our disposal in raising the antenna above the local ground clutter. The next step is raising the antenna to a height that places our signal above the radio horizon relative to our targets.

As we move into special needs—such as those presented by home-station contacts via repeaters—gain may once more take a back seat to other antenna performance parameters. In the sample case, the horizontal beamwidth of various antennas proved to be more important than raw gain in developing a solution. Even when we needed more gain, a longer antenna was not the best route to achieving it.

These notes are not in any way final answers to the questions that we have explored. Instead, they are initial options designed to expand thinking about point-to-point communications. For example, Yagis and Moxon rectangles are not the only antenna types that might meet our needs. Planar and corner reflector arrays are available and might allow easier construction, especially at 432 and 1296 MHz. Even the broadband batwing dipole array might meet some needs.

The first step in choosing the option that is correct for a given situation is an analysis of the situation itself. The second step is a full understanding of the antenna performance properties relevant to the needs of the situation. The more complete our understanding of available antenna designs, the more likely that we shall be to select a workable option as a solution. □□

Tech Notes

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Quantifying Measurement Uncertainty

Background

Now is the time and this is the place for a detailed discussion of measurement uncertainty, since we've published enough on it without a full explanation. We're serious about elevating our measurements into the realm of science.

Uncertainty is inevitable in every measurement. It may depend as much on the observer as on the equipment. Let's begin with some definitions of those things that contribute to uncertainty.

Definitions

Precision relates to the repeatability and consistency of measurements. We may define *accuracy* as the degree to which measurements agree with some value of known precision. Values of known precision represent the best self-consistent measurements we can make. Those points of reference are generally referred to as *standards*.

Perhaps a more useful pair of definitions distinguishes between *systematic* and *random* effects. Systematic effects are generally associated with equipment inaccuracy or measurement scale; random effects are generally associated with repeatability. In turn, the two effects are loosely associated with what we call *type-A* and *type-B* uncertainties. Type-A uncertainties are those that can be directly evaluated by statistical means; type-B uncertainties are those evaluated by other means. Quantification of composite uncertainty generally requires that type-B components be converted to a statistical basis for inclusion in calculations.

Any nonlinearities in the measurement process can cause random effects to appear systematically, and vice versa. That brings up the important distinction between measurement error and uncertainty. Error is the difference between a measurement result and the "true" value of what's being measured; uncertainty is an expression of confidence in the measurement result. But who's to say what's a true value except by measurement?

To take a practical example, let's say you're measuring a transmitter's output power. Power represents the rate of energy transfer and the standard unit is the watt. How is a watt defined? Well, it's defined in terms of other units that, taken together, form the basis of our coherent International System of Units (SI units). Those basic units and their abbreviations are:[1]

- meters, m (length)
- kilograms, kg (mass)
- seconds, s (time)
- amperes, A (electric current)
- kelvins, K (temperature)
- candelas, cd (luminous intensity)
- moles, mol (amount of a substance)

All other units may be expressed in terms of those basic units. The watt, for example, may be expressed as kg-m^2/s^3.

And how are those basic units defined? Well, they're defined by measurements of certain fundamental physical constants. Now you may be wondering how we can live with what seems to be a circular argument: that we define standards by measurement, and that we make our measurements by comparison with standards. The key is that those fundamental constants consistently appear in many different formulations of how the world appears to work. They include the speed of light in a vacuum (c), the mass and charge of the electron (m_e and e), Planck's constant (h), and the fine-structure constant (α). Furthermore, our system of basic units and standards is self-consistent across the range of formulations that describe how everything "ticks."

[1]Notes appear on page 56.

Aren't those basic units still somewhat arbitrary, though? Instead of the meter, we could use a basic unit of length called the *poindexter* (P) instead. But then we could be confident that 1 meter equals k poindexters, where k is an arbitrary constant. All other basic units would have to be renamed and somehow scaled to conform to equations that comply with observation, like Maxwell's equations, special relativity and so forth. Because every measurement we make depends on our quantification of those fundamental constants, it's critical that their values be known to high accuracy.

Metrology[2]

The Measurement Equation

In the communications field generally, we're not measuring fundamental physical constants. Our measurements, therefore, are functions of some other quantities. Whatever it is, each function is called a *measurement equation:*

$$y = f(x_1, x_2, x_3 \ldots x_N) \qquad \text{Eq 1}$$

where y is the thing being measured (the *measurand*) and $x_1 \ldots x_N$ are N other quantities, which somehow relate to underlying physical laws, equipment inaccuracy, known correction factors, variability in observers and everything else that contributes to uncertainty. Each test setup has its own measurement equation. To formulate one, we begin with the physical laws governing an experiment.

Take a second example: the measurement of the composite single-sideband noise of a transmitter. Most refer to that as phase noise but the term composite refers to the measurement of both AM and PM noise. Let's say we feed a CW transmitter's output into a spectrum analyzer (through an attenuator!) and examine the display. The transmitter is adjusted to provide its best constant-amplitude, constant-frequency signal.

We know that the analyzer mixes that signal and filters the product—perhaps several times—before presentation. We want a result in terms of power per unit bandwidth, relative to the power at the transmitter's nominal frequency. Since uncorrelated noise powers add, the analyzer's local oscillators (LOs) increase the displayed result because of their own single-sideband noise. In linear terms, a first approximation to the measurement equation would be:

$$P_n = P_T + P_{LO_1} + P_{LO_2} + \ldots P_{LO_N} \qquad \text{Eq 2}$$

where P_n is the measured noise power density in W/Hz, P_T is the transmitter's noise power density and $P_{LO1} \ldots P_{LON}$ are the noise power densities of the analyzer's LOs, all at the specified offset from the center frequency. Solving for the desired quantity, we get:

$$P_T = P_n - P_{LO_1} - P_{LO_2} - \ldots P_{LO_N} \qquad \text{Eq 3}$$

All terms must be regarded as estimates. Eqs 2 and 3 omit uncertainties associated with the equipment and the observer. Clearly, the uncertainties in terms on the right side of the equation contribute to the uncertainty of the result. Expressing the power densities logarithmically in dB relative to the powers at nominal center frequencies, as is usual, doesn't allow us to directly add them; but by taking the log of both sides of Eq 3, we could write an equation expressing the proper relationships. In addition, we've assumed that the LO noise power densities are uncorrelated and that assumption may be unwarranted. We've not yet included any effects from analyzer filter bandwidth accuracy, sample averaging and other sources.

A simple approach is to ensure that analyzer noise specifications exceed the measured result by a comfortable margin. A margin of 10 dB would result in an uncertainty from analyzer noise contribution of about ±1 dB. That ratio of instrumentational limits and measurement level is often called the *accuracy ratio*. Inaccuracies in the display itself and in the observer may add to that. We'd want to repeat the test several times to determine precision.

Uncertainty Computation

The uncertainty in the above example includes both type-A and type-B uncertainties. We use both statistical and empirical methods: Statistics tells us to add the noise powers (type-A) but we obtain the analyzer's noise contributions from independent measurements made by the manufacturer that we cannot statistically analyze separately (type-B).

A type-A statistical estimate of the uncertainty associated with each input variable in Eq 1 may be obtained by starting with the *mean* (average) of several different observations taken under identical test conditions. For N different observations of x, the mean is:

$$\bar{x} = \frac{1}{N}\sum_{m=1}^{N} x_m \qquad \text{Eq 4}$$

and the standard uncertainty associated with x is its *standard deviation:*

$$\sigma(x) = \left[\frac{1}{N(N-1)}\sum_{m=1}^{N}\left(x_m - \bar{x}\right)^2\right]^{\frac{1}{2}} \qquad \text{Eq 5}$$

where $N>1$. The uncertainty of a set of multiple observations is therefore obtained through averaging, as in an analyzer's video averaging function, which tends to minimize uncertainty. The use of the Greek letter sigma is common for standard deviation.

Combining Standard Uncertainties

To find the *combined standard uncertainty* of a measurement setup, compute the positive square root of the estimated variance, $\sigma_c^2(y)$, which is obtained from:

$$\sigma_c^2(y) = \sum_{m=1}^{N}\left(\frac{\partial f}{\partial x_m}\right)^2 \sigma^2(x_m) + 2\sum_{m=1}^{N-1}\sum_{k=m+1}^{N}\frac{\partial f}{\partial x_m}\frac{\partial f}{\partial x_k} cov(x_m, x_k) \qquad \text{Eq 6}$$

where f is the measurement function and cov(x_m, x_k) is the estimated covariance associated with inputs x_m and x_k.[3] Eq 6 is derived from a first-order Taylor-series approximation of the measurement equation and is the usual way of combining standard deviations and covariances. It is sometimes referred to as the *equation of uncertainty propagation*.

Engineering statistics is grungy! Note the following basic characteristics of Eq 6, though. Partial derivatives $\partial f / \partial x_m$ are called the *sensitivity coefficients*. The uncertainty of the combined result depends on the rate at which the measurement result varies with each input. Eq 6 also includes a term that takes into account how the combined uncertainty varies with the correlation of uncertainties in the inputs.

Eq 6 reduces to a very simple relationship in our example of uncorrelated noise powers that add. $\partial f / \partial x_m = 1$ and cov(x_m, x_k)=0, so the uncertainty of the result is simply the positive square root of the sum of the squares of the individual uncertainties:

$$\sigma_c(y) = \left[\sum_{m=1}^{N}\sigma^2(x_m)\right]^{\frac{1}{2}} \qquad \text{Eq 7}$$

For a 3rd-order intercept point measurement, $\partial f / \partial x_m = 3$ and cov (x_m, x_k)=0, yielding:

$$\sigma_c(y) = \left[9\,\sigma^2(x_1) + \sum_{m=2}^{N}\sigma^2(x_m)\right]^{\frac{1}{2}} \qquad \text{Eq 8}$$

where the input x_1 is the equal power of each of two off-channel signals input to a receiver. Uncertainties include, but may not be limited to: 1) the type-A or -B inaccuracies of signal generators, attenuators, spectrum analyzers, cable and connector losses and so forth used in the actual measurement; 2) the inaccuracy of equipment used to set reference levels during preparations for the measurement, and 3) the standard deviation associated with the repeatability of the measurement. That last item may be calculated using Eq 5 for multiple measurements that use identical test setups.

Expressing Uncertainty

Over the years, we've approached international consensus on a uniform way to express uncertainty in measurement. In 1981, the International Committee for Weights and Measures (CIPM) recommended a method that has gained widespread acceptance, including at the National Institute of Standards and Technology (NIST) in the US. It incorporates the statistical methods above, as well as a thing called the *coverage factor, k*, that describes an interval over which measurements achieve a desired level of confidence. Detail of coverage factor is not strictly necessary for our purposes here, since we're only interested in declaring uncertainty within reason. Coverage factor is important where very accurate measurements must be supported with statistical evidence.

What's necessary for both casual and professional communications workers is a declaration of uncertainty *for all measurements*, regardless of magnitude. For consistency, the number of significant figures in reported results should correspond to the stated uncertainty. For example: 101 ± 2 dB, not 101. 2 dB. Finally, definitions of measured parameters should be in plain English without reference to a test procedure.

Important note: This procedure produces a statistical uncertainty that represents a certain confidence level in a measurement. It is not a worst-case analysis. It says, in effect, "I am this confident about the accuracy and precision of my measurement but there is a small chance it could be worse than that." Corrections to data because of known errors should be included in a measurement result but not as part of the uncertainty calculation. Corrections have uncertainties, though, and those must be included in the composite uncertainty.

Acknowledgments

We have to acknowledge uncertainty in our measurements as readily as we have to deal with it in other facets of our lives. Even when measuring fundamental physical constants, on which every other measurement is based, Werner Heisenberg showed that the very act of measurement affects what is being measured. Note that we could have substituted the word noise for uncertainty almost everywhere without losing much meaning in the above discussion.

Many thanks to Joe, K1JT, and Leif, SM5BSZ, for reviewing my scribblings. Their comments and suggestions are always on-target.

Notes

[1] *Reference Data for Radio Engineers*, 6th ed, Howard W. Sams & Co, 1975, New York.

[2] *Uncertainty of Measurement Results*, National Institute of Standards and Technology (NIST), 2000, **http://physics.nist.gov/cuu/Uncertainty/index.html**; also see B. Taylor and C. Kuyatt, *Guidelines for Evaluating and Expressing the Uncertainty of NIST Measurement Results*, NIST *TN 1297*, 1993, and other publications at **http://physics.nist.gov/Pubs/pdf.html**.

[3] *Guide to the Expression of Uncertainty in Measurement*, International Organization for Standardization (ISO), 1993; reprint with revisions, 1995. This document was published in the name of seven international standards organizations that supported its development: Bureau International des Poids et Mesures (BIPM), International Electrotechnical Commission (IEC), International Federation of Clinical Chemistry (IFCC), ISO, International Union of Pure and Applied Chemistry (IUPAC), International Union of Pure and Applied Physics (IUPAP) and International Organization of Legal Metrology (OIML). □□

Letters to the Editor

A Better Antenna-Tuner Balun (Sep/Oct 2005)

Dear Editor,

Andrew Roos's article, "A Better Antenna-Tuner Balun," was interesting and informative, but there were a couple of mistakes. First, the ideal transformer does not follow his first rule that winding currents are equal and opposite. There is nothing about transformer operation that implies it. If that were true, an open-circuited secondary on any transformer would rise to arcing voltage in an attempt to maintain equal currents, but this obviously does not happen. In an ideal transformer the winding voltages are equal (at the turn ratio) because of their shared flux, but not necessarily winding currents by any means. This is a serious misconception and may influence conclusions in the text; I have not examined that.

It is an unfortunate misnomer that the common-mode choke has been called a current-mode balun. There is really no inherent balancing action to it at all; it is truly just a common-mode choke. With a floating load, the load is as unbalanced as the source. If the middle of the load is grounded, the load will be balanced (as Roos points out); but if you're going to ground the middle of your load then many transformer configurations will provide a balanced load.

This is not a particular feature of the so-called current-mode balun. The grounded load center almost never applies to antennas. As Roos points out, antenna coupling to source ground is generally high-impedance and asymmetrical. The common-mode choke will reduce common-mode currents if impedance levels are appropriate. Except in specific cases this does not result in a load balanced with respect to ground.

Regards, Gerrit Barrere, KJ7KV, **gerrit@exality.com**

Dear Mr Barrere,

Thank you for your letter. I am pleased that you found my article interesting and informative.

Most of the sources that I consulted give the following equations for an "ideal transformer":

$V_s/V_p=N_s/N_p$ and $I_s/I_p=N_p/N_s$

where V_p and V_s are the voltages on the primary and secondary; I_p and I_s, the currents flowing in the primary and secondary; and N_p and N_s, the number of turns on the primary and secondary. For a 1:1 transformer, $N_s=N_p$ so $V_s=V_p$ and $I_s=I_p$. It merely remains to note that the signs of I_p and I_s are chosen such that the currents are flowing in opposite directions.

The physical assumption that underlies this understanding of an "ideal transformer" is expressed by Professor Roger King, Interim Chair of the Department of Electrical Engineering and Computer Science at the University of Toledo, as follows:

> *The ideal transformer does not generate, dissipate, or store energy. Therefore the instantaneous power leaving the transformer is the same as that entering. This could be said in other words by saying that if one were to draw a box around an ideal transformer and sum the power flows into (or out of) the box, the answer is zero at every moment in time. ("The Ideal Transformer" at* ***www.eecs.utoledo.edu/~rking/circuits/IdealTrans.pdf****).*

This does not imply that the voltage across an open secondary will be infinite. Rather, the input impedance looking into the primary will be infinite, so no current will flow in the primary, and the voltage across the secondary will be the same as that across the primary.

This interpretation seems to be shared by the majority of authors. However, I do note that at least one reference only requires that an "ideal transformer" conserve energy over the entire cycle, rather than at every instant **(www.saburchill.com/physics/chapters/0061.html)**. With this interpretation, an ideal transformer may still store energy in reactive fields, in which case I agree that the currents flowing in the primary and secondary are not necessarily equal and opposite.

Fortunately this discussion is of semantic interest only, as the effect of allowing the transformer to store energy in reactive fields would be to add a reactance in parallel with one of the transformer windings (or both, the effect is the same). This is precisely what I did with the "winding reactance" Z_w in the article. This reactance was taken into account in all the derivations, so the results are valid irrespective of which definition of "ideal transformer" you choose to adopt.

It seems to me that what counts as a "balun" depends on what it is you are trying to balance. If a load is driven using a common-mode choke, then the currents flowing into and out of the load will be equal and opposite, ie balanced, although the voltages across the load will not necessarily be balanced with respect to ground. If the load is driven using a Ruthroff ("voltage") balun then the voltages across the load will be balanced with respect to earth, although the currents flowing into and out of the load will not necessarily be balanced. (Both cases are subject to the limitations due to the winding reactance, which are treated in detail in my article.) Although I do not know the history of the term "balun," it seems to me that contemporary usage, which allows these devices to be called "current" and "voltage" baluns, respectively, is quite accurate and descriptive of their effects.

73, Andrew Roos, ZS1AN, **zs1an@qsl.net**

Dear Mr. Roos,

Your points about conservation of energy are quite right, and that is a good way to approach the problem. It does ignore the practicalities of finite magnetizing current, however, which can be very important in resonant transformers, transformers at their low operating frequency limit, etc.

The current ratio being fixed at the turn ratio is sort of a "secondary" effect (excuse the pun). It is due to load current being transferred from the secondary to the primary, at the turn ratio, and doesn't include magnetizing current (which applies to the primary only). It is really derived from the voltages of the windings being locked at the turn ratio due to Faraday's law, and is easily misconstrued as being fundamental when dealing with just a two-winding transformer.

Consider a three-winding ideal transformer with the same turn count on all windings, and one primary. Connect different resistors across the two secondaries. Of course the winding currents will be different, and they will not have anything to do with the turn ratio. The sum of the secondary currents will equal the primary current, but only this sum will have anything to do with the turn ratio.

What does happen is that the secondary winding voltages will be the same as the primary. The currents will be derived from that basic relationship. There isn't anything about ideal transformer operation which forces currents to match. This is just a derived relationship from the winding voltages.

To the second point, I agree that

"balance" means different things in different contexts, to different people. But the term "common-mode choke" emphasizes one of the points in your article: that its effectiveness depends on the impedance levels in the rest of the circuit. It is part of a filter; it does not perform a balancing function inherently, and thus should not be called a balun at all. A "choke" is only as effective as the impedances it works into, whereas a "balun" (ie, a voltage balun) forces balance regardless of the rest of the circuit. I think it's important to discriminate between "choking" (filtering) and "balancing."

Best regards, Gerrit, KJ7KV

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

Doug,

I have been looking for a good digital solution to the S/N problem of HF voice transmission and your approach is definitely not it! Worse than the irritating flat affect of synthetic voice output (I'd rather a miss-tuned [sic] SSB duck), is the misery of training a speech recognition program. If all ham licenses [sic] required either voice training a computer or passing a Morse code test, hams would flock to CW.

I can see it now. Instead of complaining about an operator's lousy fist, we will all recognize the operator by the errors his voice recognition system is making.

If you want to do things right, analyze 10,000 QSOs, break them down into categories (could there be more than 10?), and transmit only words to "fill in the blanks" (eg, handle, wx, antenna height). That would provide lots of data compression. Even better, put all those details for each ham on an Internet database and let an online computer fill in those details. Now we are talking data compression.

Perhaps the future is to have the recipiant [sic] download a Java applet to the sender's computer and then the two hams would only transmit "interesting" data, each based on the other ham's filter. Uninteresting segments could be sent as "yada yada yada."

If you do intend to pursue this Votrax mode of communications, I suggest you limit the fixed vocabulary to about 200 words. Linguistic research (references upon request) shows that even adults use very small vocabularies for 80% of conversations. The rest of the words should be "built up" as a new library adaptively during the QSO. You could create a library of libraries and try to match one as the QSO progresses. I still say you should dump the robotic voice and let the recipient read the screen. He/she can more naturally add inflections in his/her head while reading. Ultimately, something gets lost in translation when the jokes are told by Gort.

Andy Mitz, WA3LTJ, **arm@helix.nih.gov**

Hi Doug,

I enjoyed your article in the current *QEX* and was describing the technique to my XYL. The concept was quite familiar to her, since she was a very proficient shorthand typist many years ago. It seems that source coding has a long history.

In shorthand they go further than efficient coding of individual words by treating many standard phrases used in business correspondence. In our case I wonder if coding phrases like "my QTH is...," etc, would be helpful. I also understand that shorthand can accommodate the use of hard and soft vowel sounds and many standard abbreviations are used, "ack" for "acknowledge" and so on.

73, Ron Skelton, W6WO, **ronskelton@charter.net**

A Direct-Conversion Phasing-Type SSB Rig (Nov/Dec 2005)

Hi Doug,

Many years ago, I designed and built a phasing receiver (R. P. Haviland, "An Advanced Amateur Receiver," *CQ*, Oct 1950). Because of work, moving and changed interests, I never got around to keeping the design modernized. So it was with great interest that I read KQ6F's article. He has done a superb job of using modern components to realize his design aims. I do suggest some extensions, based on what I learned from my early designs. While these are primarily to extend the range of signal types that can be demodulated, one offers the possibility of SSB improvement.

The first addition is to add a switch at the output of the phasing networks ending in U5, coupling to two output audio amplifiers instead of just one. The switching should allow "Both to I_{out}, to Q_{out}, to $I_{out}+Q_{out}$, to $I_{out}-Q_{out}$, One to I_{out}, the other to Q_{out}." This converts the receive chain from an SSB/CW type to one which will properly demodulate upper and lower sideband SSB, CW, AM, narrow-band FM independent sideband AM or FM and DSB suppressed carrier. Wider-band FM can be usable, with some signal loss.

This is not optimum for double-sideband signals. For these, a carrier should be inserted at the proper frequency and phase. Changing U2 to a loop controlled VFO is part of this. My original design used only phase detection of the received carrier, which is optimum for AM and FM, but poor if there is great carrier suppression. There are several possibilities for these. One is to multiply I_{out} by Q_{out}, using the low-pass filtered output to control the VFO, the dc component becoming the lock. Another is to add to this a narrow-band stage, to output the residual carrier that exists for all signals. Still another is to design this for the 60 or 120 Hz "hum" modulation, which, however, is becoming less prevalent as switching power supplies become more common. While some experimentation is indicated here, a good system would improve SSB by restoring voice frequency accuracy.

If the various modes of modulation are analyzed for communication capability in a crowded band, it will be found that the real gain of SSB lies in the elimination of the carrier, not in the single sideband feature (R. P. Haviland, "A Comparative Study of Communication Systems Using Different Modulation-Demodulation Techniques," *E.B.U. Review*, June 1969.). An important feature in AM reception appears when the two audio outputs are fed to stereo speakers: The distortion due to selective fading disappears; instead the signal appears to wander across the space between the speakers. With stereo headphones, the wander appears to be in the space between the ears. The double-sideband systems have an advantage when interference shows up: Switching to the sideband reject mode can reduce or even eliminate the QRM.

I have not attempted to state the changes to the transmit chain. CW, AM, SSB, FM and reduced carrier operation would be possible with the correct switching.

I submit that this form of optimum reception of signals deserves wider use.

R. P. (Bob) Haviland, W4MB, **bobh@iag.net**

Doug,

Rod Brink's article was very interesting. However, I had some trouble figuring out how the AGC worked until I realized there is an error in Fig 6. The output pin, 7, of U8B should go to R52 (as it does) but it should not go to the +11.5 V power supply. In the configuration shown the AGC cannot work and it may also be that U8B could blow out, depending on the internal design of the device.

Ernie Moore, VE3CZZ, **ejmoore@trytel.com**

Hi Ernie,

You're right. Break the connection between U8B pin 7 and the supply voltage.

73, Doug Smith, KF6DX, **kf6dx@arrl.org**

Quadrature Phase Concepts (Nov/Dec 2005)

Doug,

I puzzled over your article on phasing mixers. I finally focused on Fig 2 and the text that applied to it. You asked for an alternate explanation, so here goes.

Fig 2 is a classic "image-reject" mixer. With the sum of the two phased mixer outputs as you show, the mixer outputs at 5.55+7.25 MHz, if unfiltered, would add and produce an output a 12.8 MHz. As shown the difference at 1.7 MHz will cancel, so no surprise at the "Output not Significant" label on the sum output block. If you replace *sum* with *difference*, now the mixer outputs at 12.8 MHz are the ones that will cancel. No need for a 1.7-MHz filter to eliminate them. At the same time, the two 1.7-MHz components will add. Just by using a difference, voila: the desired 1.7 MHz will be present with no other output.

The 1.7-MHz filters between the mixer outputs and the phase-shift networks are a confusion factor: You don't need filters there. If you want more rejection of 12.8 MHz, just one filter on the sum output would perform the same function without the need for wideband phase matching of two filters to preserve the summing and cancellation this mixer can provide. But now on to what you're really thinking about, rejecting the opposite sideband of existing signals.

As shown, Fig 2 is phased to be what is sometimes called a "USB" mixer. That is, it's normally designed to produce the sum output frequency and cancel the difference frequency. With a difference block instead of sum block connected to the mixer outputs, it's sometimes called a "LSB" mixer since it produces the difference frequency and cancels the sum. I suspect this "eliminating the USB" in a "USB mixer" terminology got your hopes up, thinking it might somehow be able to selectively reject already-generated SSB signals. No such luck. The USB of the original signal generation doesn't have any relationship to the USB of this mixer.

A signal at 7.25 MHz will be translated down to 1.7 MHz. Signals at 7.251 MHz (USB if its carrier was 7.25 MHz) and 7.499 MHz (LSB if its carrier was 7.25 MHz) will be translated down to 1.701 MHz and 1.699 MHz, respectively—still USB and LSB with respect to the new 1.7-MHz center. That's because *both* of those signals, being above the 5.55 MHz oscillators are, to the two mixers, both on its upper sideband. It shifts everything linearly. This mixer, with or without filters, cannot alter an AM, USB, LSB, CW, slow-scan or RTTY signal in any way, it just shifts it down. The "LSB" it's rejecting is signals at 5.55-1.7 MHz, on 80 m. To beat on a dead horse, 80 m is this mixer's LSB, and that has nothing to do with whether any of the signals are USB or LSB.

The only way you can reject an opposite sideband signal is for it to be on the opposite side of the local oscillator. To do this your carrier needs to be midway between the USB and LSB of that signal—in other words, at the frequency where the original carrier would have been: direct conversion. Or perhaps better yet, make a receiver with two image-reject mixers: the first one to translate 40 m down to 1.7 MHz (rejecting 80-m signals), the second at 1.7 MHz to reject the opposite sideband.

Regards, Bill Carver, W7AAZ, **bcarver@safelink.net**

Doug,

Gary Heckman's article shows just how important the statement of a problem can be to the solution. I remember (dimly) examples in calculus class where a teacher filled the black board with a messy equation, and by choice of a substitution, reduced it to something simple by selecting just the right sub-expression to factor out before going on. I never would have gotten to a solution by myself.

In this case, quadrature mixing as a method of eliminating part of the RF spectrum is indeed a fact accomplished by thousands of pieces of equipment every day. Probably the most common use is the "image-reject mixer" that uses quadrature techniques to eliminate the unwanted image from a mixing process—particularly useful when the image lies close to the desired frequency and thus is hard to filter.

The reason Gary's analysis did not produce a validating proof is that the process depends on the frequency and phase relationship between the local oscillator and the incoming signal(s). The incoming signal therefore must be described in terms of the oscillator signal. Referring to Fig 2 in the article, the accompanying text describes the incoming signal frequency as F and the oscillator frequency as O. It turns out that for any incoming frequency F, that frequency can also be described as $O+A$ where A is the difference (in frequency and phase) between F and O. Like the math tricks my professor did in class, this substitution is helpful precisely because we suspect where we want to end up. The desired output of the mixer is A, the difference between F and O, so stating F in terms of O and A is useful.

Note that A can be positive (if F is above O, as for a USB sideband) or negative (if F is below O, as for a LSB sideband). Since A is a frequency (indeed the frequency we anticipate will come out of the circuit) one should ask what the meaning of a negative frequency is. As Gary points out, the mixers in this case are folding the spectrum around the frequency of the oscillator (O). As F gets closer to O, A gets lower, as F passes O, though A goes through zero and then gets higher. This concept is well known in frequency analysis work. In fact the concept of negative frequency (and the more generalized concept of complex frequency) is an integral part of the fast Fourier transform. From trigonometry, $\cos(-A) = \cos(A)$, but $\sin(-A) = -\sin(A)$. It is this sign inversion that, combined with quadrature detection, removes all frequencies either above or below O.

73, Wilton Helm, WT6C, **whelm@compuserve.com**

Antenna Options (Nov/Dec 2005)

Hi Doug,

I very much enjoyed L.B. Cebik's series on antenna modeling software in *QEX*. Table 2 in Part 2 of the series implies that 4NEC2 works with NEC-2 and not with NEC-4. This is not the case. Arie has worked with several of his NEC-4 licensed users to provide NEC-4 compatibility in 4NEC2. I am one of those users and wished to point out one other nice feature of 4NEC2 that I use in conjunction with NEC-4.

In 4NEC2, if one turns on the option: *Main->Settings->Other settings->Wait for DOS box* and appropriately edits 4nec2.bat in the 4NEC2 package with a pause statement and changes the paths of the NEC input and output files, then 4NEC2 can be used as a preprocessor and postprocessor for a NEC engine run on another platform.

I use *Linux*, primarily, and I build most of my NEC-4 engines under *Linux*. With the above scheme, I can generate NEC-4 input files with

4NEC2, run NEC-4 with those input files and give the NEC-4 output files back to 4NEC2 and continue with the 4NEC2 output graphics for the NEC-4 output—very nice.

Furthermore, 4NEC2 runs under *wine* under *Linux* making this whole process very easy. It also runs under *vmware* under *Linux*. But one can do this procedure with separate *Windows* and *Linux* platforms if he desires. And of course, one may use his own licensed *Windows* NEC-4 engine without changing 4nec2.bat. Most importantly, I wanted to make it clear that 4NEC2 provides NEC-4 solutions as well as NEC-2 solutions for LLNL licensed NEC-4 users.

You can find out all about 4NEC2 and download it from **home.ict.nl/~arivoors/.** I have no association with 4NEC2 other than being a very satisfied user. The software has provided me with very nice pre- and post-processing for NEC-4 that I would not have otherwise.

73, Harry McGavran, W5PNY, **w5pny@arrl.net**

Doug,

I am glad to learn of the added capabilities of 4NEC2 of which I had not been fully aware when writing the series. This is information that deserves to be shared with *QEX* readers, whether they are *Windows* or *Linux* users. Since 4NEC2 is free for the download, it is a nearly ideal experimental platform for those who desire to see what adaptations are possible in the NEC arena.

Indeed, there are a considerable number of *Linux* users among the amateur community. You might consider an article providing some detailed information for present non-*Linux* users (mostly we *Windows*-bound types) on how to use *Linux* effectively, its advantages, its limitations, etc., and adapting the NEC-4 core to the *Linux* mode using the 4NEC2 interface (in and out) might make a good focal point for such a piece. Your steps are straightforward, but there is a good bit of background that is worth explicating.

Thanks again for the update.

L.B. Cebik, W4RNL, **cebik@cebik.com**

Reduced Bandwidth Voice Transmission (Nov/Dec 2005)

Doug,

What you are looking for [here] is good polarization isolation. That can't be achieved by crossed dipoles. Here some info about it.

The amount of "cross-pol" from any antenna is not just 20 dB. It is a very complex function and a difficult amount to determine. The amount of cross-pol is different for different locations in the antenna pattern. It is also different for different antenna types. Crossed dipoles might have the most amount of cross-pol. There are several areas in the pattern to consider.

Boresight axis: This is where cross-pol is minimized. Here –20 dB cross-pol might be possible. However, roving off the boresight axis even by a degree or so increases the cross-pol level. If cross-pol is measured, there in a null exactly on boresight (assuming matched antennas, and no ground interference). The cross-pol pattern looks very similar to a difference pattern of a monopulse radar antenna.

Off axis on principal axis: Cross-pol in this area is minimized, but I would not expect it to be 20 dB down.

Off axis in the 45-degree planes: Cross-pol sidelobes (also known as Condon lobes) exist in these planes. It is possible for the amount of cross-pol to exceed co-pol on the cross-pol sidelobes. Yes, a single dipole generates a lot of cross-pol on the 45-degree axis.

A better antenna choice: A reflector antenna would be a better choice to reduce cross-pol. For an axi-symmetric reflector, using a circular polarizer, cross-pol is given by the "return loss" on the polarizer ports, if the return loss on all the ports is equal. In fact, it is possible to have almost no cross-pol sidelobes with an axi-symmetric reflector. That subject is covered in the literature under the subject "Integrated Cancellation Ratio," (ICR) as it was a method used to reduce weather clutter from radar signals. This works because rain returns an almost entirely a cross-pol signal. The circular-polarizer part here is important. If a linear polarizer is used, the Condon lobes always exist.

In polarimetric radar, use is made of both the cross- and co-pol signals. It is possible to discriminate rain, snow, ice, hail, etc, to a fine degree if the ICR of the antenna is good enough. Something better than –30 dB (two-way) is good enough. There is a report out there that shows raindrop alignment and misalignment during a thunder discharge, over about Springfield, Massachusetts, using a polarimetric radar located in Sudbury, Massachusetts.

Keep up the good, challenging, articles.

Jim Ussailis, W1EQO, **ussailis@equinox.shaysnet.com**

Empirical Outlook (Nov/Dec 2005)

Hi Doug,

Thanks for a provocative editorial regarding "Emergency Preparedness and the New Federalism." You brought up many areas of concern directly affecting amateurs responding to Katrina. I agree that better coordination is required and hopefully ARRL will work directly with the Department of Homeland Security to resolve these problems.

But in the last half of your editorial, you digressed into engineering and funding issues beyond the purview of Amateur Radio. You implied that the flooding might not have occurred because "Congress and the president clearly need to get their acts together."

Funding, design, construction and maintenance of a levee system is not a single-year issue. The US Army Corps of Engineers has been responsible for these levees for over 40 years. I heard a radio interview with a ranking Corps officer who stated the FY2005 funding was only reduced from the requested level, not totally eliminated. He stated the lack of 100% funding was not the cause of the levee failure, as the deleted funding was for ship movements, not levee construction.

The failure analysis is ongoing but to date, they have found two causes:

1. Hundred year-old trees growing on the dry side of the levee had roots seeking water by growing under the levee; when the hurricane topples the tree, a large root ball rips out a chunk of the levee, causing seepage, then levee failure.

2. Continued subsidence of the Mississippi River delta soil as much as 1 inch per year in some areas. Levees constructed 20 years ago are now 13 to 21 inches below as-built grades.

The news media kept reporting Hurricane Katrina was a Category 5 when about 250 miles south of New Orleans. Ocean buoys reported waves over 60 feet high before the buoy disappeared. Massive damage was done to offshore oil platforms. But as Katrina moved north, the winds diminished to Cat 3 (as measured by hurricane-hunter planes).

As Katrina moved inland, the wind speeds fell more, especially in the northwest quadrant of the eye (New Orleans area). As soon as conditions permitted, Louisiana State University weather researchers recovered wind-speed recordings from ground-based instruments. To their surprise, east Lake Ponchatrain had only 98 mph winds (Cat 1) and West Lake less than 70 mph, below Cat 1. But the news media continued to report New Orleans devastated by Cat 3 storm. Wrong!

These ground wind speeds can be verified by comparing the downtown New Orleans buildings to those in the last storm to hit southwest Florida. The Florida buildings had structural damage and windows blown out. The New Orleans buildings had little wind damage but flooding made them uninhabitable. In fact, many had rooftop billboards that were intact.

What are my information sources? I work with Chevron employees and contractors who responded from California for storm damage repair. They sent e-mails with attachments from local TV stations, newspapers, etc. Chevron is self-insured and they attempt to gather information for their own failure analysis of fallen radio towers, etc. Most of the information has not been on the major news media.

So this begs the question: If the city, parish and state governments cannot properly respond to a barely Cat 1 storm, what would have happened during an actual Cat 3 or stronger hurricane?

Amateur operators cannot change their response due to poor planning, incompetent authorities, political bickering or lack of funding. We just do what is necessary. Thanks for a great magazine.

73, Dave Light, KG6YSZ, **jlight@bak.rr.com**

Hi Dave,

And thanks for your thoughtful letter. The meaning of the last part of my editorial is this: The federal government's primary job is to protect its citizens. They've had their chance to design, build and maintain a system to do that in New Orleans and elsewhere and have failed.

You're right that the huge funding cut in fiscal 2005 wasn't the proximate cause of levee failures. It was a matter of inadequate engineering and execution over the decades.

It's time to let the locals do the job, give them what they need and relieve the "feds" of any responsibility but funding. In other words, involuntarily terminate them for cause.

You know what's said about an ounce of prevention. We're hams but we're also US citizens and many of us are engineers. We *can* dispense with poor planning, incompetent authorities, political bickering and lack of funding. I believe we deserve a chance to show that we can do better. Give us the money and let us organize our own flood-control work. Our locals will appreciate the jobs and the ability to better communicate with those in charge. Only then should we judge county and parish responses.

73, Doug

Octave for Signal Analysis (Jul/Aug 2005)

Hi Doug,

Oops! Ray Mack, WD5IFS, is right. There is an error in my letter (p 61) in the Nov/Dec issue of *QEX*. The return of fft() when presented a cosine wave will be real and when the time domain function is a sine wave, the return will be imaginary.

While considering the phase angles of the results from *Octave*, you might notice that the sinusoids I use in the article only approximate the theoretical results for sinusoids which have zero crossings at time zero (sine) or local extrema at time zero (cosine). This stems from the fact that *Octave* uses one-based vectors and matrices like *FORTRAN*. Other languages, such as *C*, are zero-based, and some, like *Mathcad* and various flavors of *BASIC*, make the base an option. There's been a continuing controversy over that issue since the early days of programming and I won't pursue it here. Searching the Web yields widely varying opinions on the matter.

To develop a sinusoid using either cos() or sin() with zero offset in phase, N samples of the time domain waveform should extend from 0 to N–1. If the one-based element numbers of *Octave* are used as times, the first time, rather than being zero, will be offset in time slightly and the sine or cosine wave will begin with a slight phase offset that will carry through to the transform into the frequency domain. I could have changed that in the code in the article, but that would have added an additional complication to explain and I chose to leave it out. Changing the phase offset to zero changes neither the validity nor the accuracy of the computations and it will be noticed only when you look at the transform of a sinusoid (using either the sin or the cos function) by itself. Only the phase is changed, not the magnitude yielded by abs(), and we discarded the phase for the purposes of the article.

73, Maynard Wright, W6PAP, **m-wright@eskimo.com**

Help for Oscillator Failure in the HP8640B (Sep/Oct 2005)

We inadvertently omitted the sidebar that is referenced on p 38, col 2 par 1 of this article. Here it is—sorry, John!—*Doug, KF6DX*

What if the Transistor Itself Has Failed?

So you have gone through the voltage-at-the-transistor-socket checks and find that you have the proper plus and 20 V dc? Well, it is possible that you may have a failed oscillator transistor. HP (now Agilent) references this NPN RF transistor as Q1 in the RF oscillator assembly A3. The part number originally assigned to this part is 5086-4282. If you look up this part on the Agilent web site, it will refer you to a newer part number under the Agilent numbering system (1GS-4214), but it will also tell you that it is obsolete. In private email communications with Agilent service, they stated that they were unable to provide any parametric information on the obsolete transistor.

This transistor needs to be an NPN bipolar device in a TO-72 case (although perhaps TO-18 would work—only two leads, the collector and emitter, are needed) with a transition frequency (f_t) high enough to have adequate gain to oscillate up to 550 MHz or so and to survive the roughly 40 Vce applied by the external circuitry. It also needs to have the base connected to the case internally, because the base is grounded in this application, and the ground is made by physical contact between the case and the protective hex cap. In a brief, non-exhaustive search of major manufacturer components, I've not been able to find anything that meets these requirements. This may have been a proprietary HP/Agilent component. If anyone has successfully found a replacement for this device, *QEX* would be a great place to tell the rest of us about it.—*John Klingelhoefer, WB4LNM, 1500 Kingsway Dr, Gambrills, MD 21054* □□

In the next issue of QEX/Communications Quarterly

Fritz Raab, W1FR, brings us a simple, high-efficiency class-D transmitter for LF and HF operation. The unit can be adapted to operate on any frequency between 100 kHz and 3 MHz simply by changing the output filter. It uses inexpensive MOSFETs. Drain voltage is 165 volts and efficiency is close to 90%. Check it out! □□

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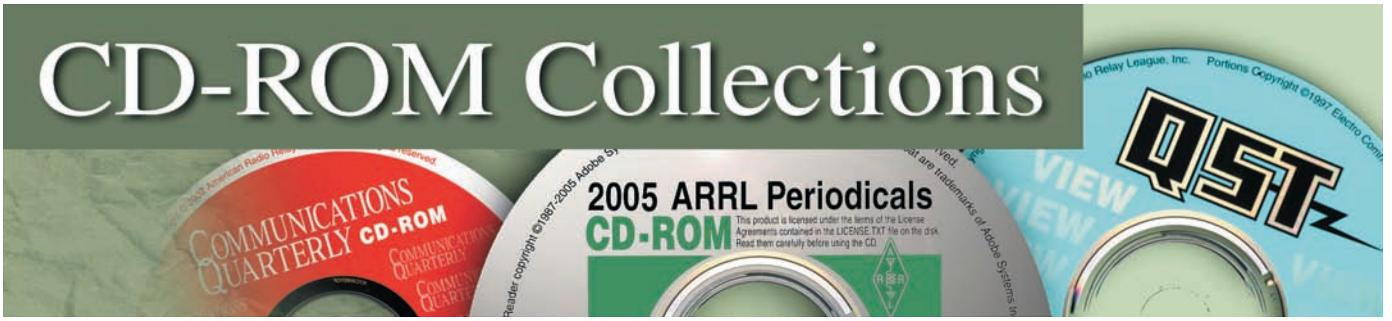

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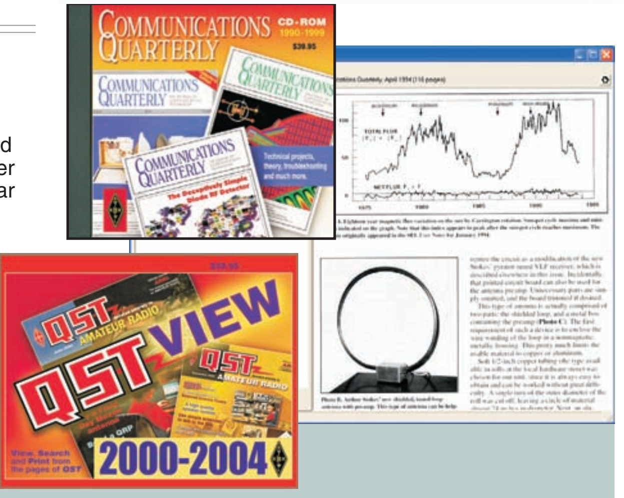

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