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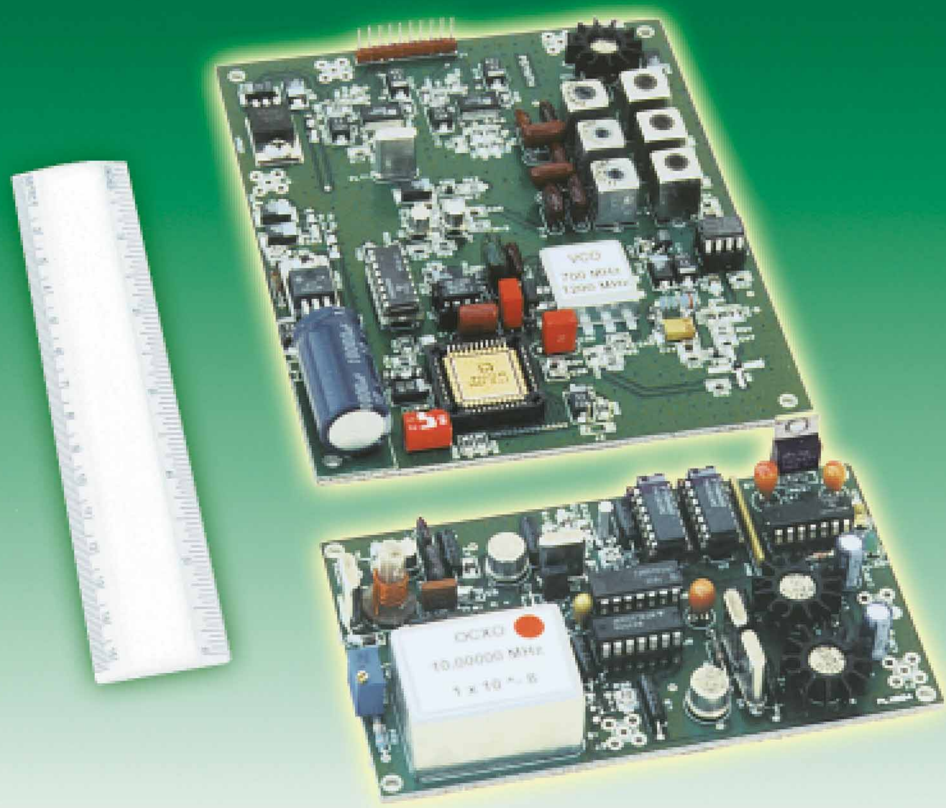
Forum for Communications Experimenters

May/June 2001



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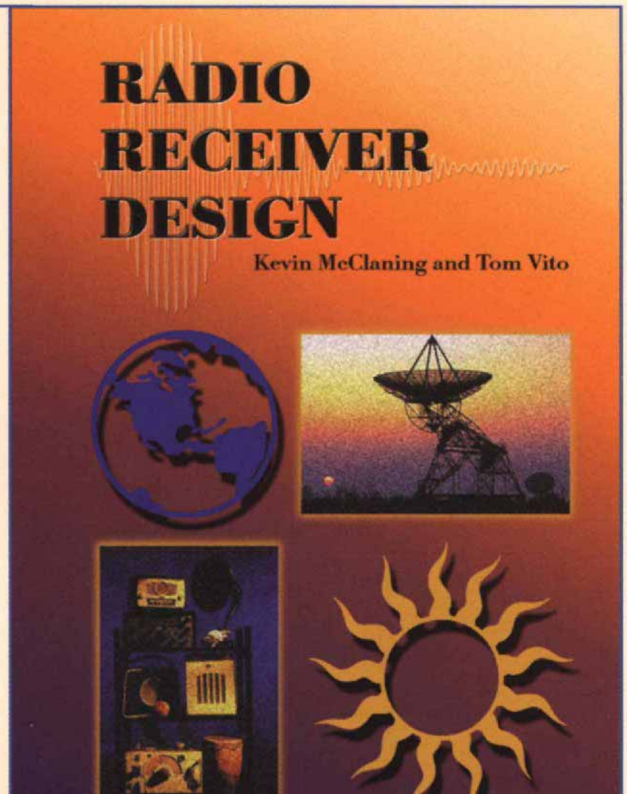

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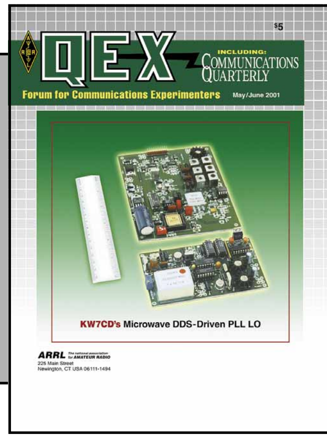

About the Cover

KW7CD's LO for an HF transceiver with a 75-MHz IF, read about it on page 3.

Features

Columns

May/June 2001 *QEX* Advertising Index

THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interests in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

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1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,
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Empirically Speaking

Frequency-synthesis devices are now available with performance only dreamed of even five years ago. They are enabling radio designs that would otherwise be impossible. They bring spectacular specifications within reach of the commercial designer and Amateur Radio experimenter alike. Sub-micron technology using various chemistries and topologies has produced some startling results.

The state of the art is such that we must search for a better term than DDS for some of those devices. Circuit integration has advanced to the point that almost everything needed to produce a digital radio exciter may be included on one chip. While that is alluring, we are finding that the need for flexibility forces us to distribute signal-processing tasks across several discrete parts. One logical partition occurs between the generation of signals in digital form and their subsequent conversion to analog form and vice versa. That point of interface is likely to endure. Possibilities for transmitters are maximized when hardware supports the largest variety of coding schemes, modulation formats and communications protocols. It's no wonder then, that we see a drive toward general-purpose signal-processing hardware, even in this age of radios on a chip.

I tend to dwell on signal processing in receivers because it is usually much easier to transmit a signal than to receive it. What has DSP done for us lately in receivers? Well, some improvements have naturally come at the extremes of signal strength. At the large-signal end of things, analog mixers still dominate performance. Before too long, it may be that we'll see practical designs that eliminate mixers altogether to achieve surprising dynamic range. At the small-signal end, DSP is already getting some astounding results.

Experimenters are applying concepts of long-time integration and time-frequency analysis, gleaned from other fields of research, to the reception of signals buried far below the noise level. Long-standing distance and QRP records are being shattered with those techniques. Maximum-entropy deconvolution methods have reached prominence in recovery of data from remote spacecraft, as well as from telescopes in astronomy, from RF to gamma rays. I expect Amateur Radio to benefit from application of those and other methods to weak-signal work. What is possible is nothing short of astonishing.

Here's another thing I think is astonishing: With all that wonderful progress being made elsewhere, that there should be so much disagreement about the source impedance of an operating RF amplifier. Finding the s_{22} of an amplifier is a fairly common procedure among working engineers and data on real circuits are not scarce. Leaders in Amateur Radio, industry and academia have made many measurements and they are there for your perusal. As our recent coverage of this topic winds down, please remember that we have tried to focus on what is new. Will we see anything new on the subject after today? Only time will tell.

In This Issue

Cornell Drentea, KW7CD, returns with the second part of his "Beyond Fractional-N." He provides details of his recent designs along with the promise of further improvement. We include a bit of information about state-of-the-art synthesizer chips. H. Paul Shuch, N6TX, brings us a look at the search for intelligent, extraterrestrial radio signals. It is perhaps surprising how similar that is to the search for intelligent signals originating on this planet.

Bob Freeth, G4HFQ, takes much of the mystery out of working with sound cards under *Windows*. Walter Schulz, K3OQF, also looks to the heavens, but with a different emphasis. He shows how to accurately get your bearings using surveying techniques. Walter Maxwell, W2DU, gets his say on amplifier source impedance; he seeks to bolster his view of things. John Stephensen, KD6OZH, explores effects of mixer-port terminations and supplies ideas about how to get the most out of your receiver's front end. Bill Walker, W5GFE, has written some programs that reduce the output of *NEC* antenna modeling software to graphs. You *Linux* and *Unix* users may find them of special interest.

Peter Bertini, K1ZJH, presents an interesting discussion that appeared in *RadComm* not long ago. Zack Lau, W1VT, presents a low-cost 222-MHz helical band-pass filter in his RF column. Visit Dayton—it's a blast! *—73, Doug Smith, KF6DX;* **kf6dx@arrl.org**.

Beyond Fractional-N, Part 2

A DDS-driven PLL at microwave frequencies affords superior performance in up-converting 75-MHz-IF HF transceivers

By Cornell Drentea, KW7CD

In Part 1 of this series,[1] we saw current and future methods of implementing frequency synthesizers for HF transceivers. We learned about brute-force synthesizers, PLLs and fractional-*N* (dual-modulus) types; DDSs and DDS-driven PLLs, as well as DDS-only synthesizers and all their combinations. We also learned that while it is relatively easy to design brute-force, low-resolution synthesizers, it is much more demanding to design their broadband, high-resolution counterparts. The tradeoffs are significant and complex. In addition, we learned that a DDS-driven PLL remains the preferred solution in high-resolution, wide-band frequency synthesis today. This is so primarily because of its simplicity and its ability to eliminate complex multiple loops, mixers and other brute-force synthesis techniques and the attendant implications for spurious sidebands.

Here in Part 2, we will use the knowledge we gained from Part 1 to implement the design of a DDS-driven PLL in an up-converting HF radio with a 75-MHz IF. This synthesizer operates at 10 times the required frequency range to improve its phase-noise performance. It was designed by the author in upgrading his previously developed Star-10 transceiver (see Part 1, Fig 5B). The complexity of this design and its boards preclude home duplication of the circuit boards; etching patterns are not available.

Design Criteria

Initially, a DDS-driven PLL operating directly at the LO range of interest (77 to 105 MHz) was contemplated for this application. As indicated before, in HF communications, a phase-noise performance of –90 to –110 dBc/Hz over the 1- to 10-kHz range would be required. Although this would be sufficient, better performance is desirable; however, the great bandwidth of such a synthesizer limits what we can do.

To get better performance, the DDS-driven PLL presented here operates at 10 times the frequency range required, or between 770 and 1050 MHz. This frequency range is then divided by 10 to realize a 20-dB improvement in the phase-noise performance of the synthesizer overall. Although, it could be argued that similar results can be obtained with a straight on-frequency

[1]Notes appear on page 9.

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DDS-driven PLL (see Part 1, Fig 5A), experience shows a 6- to 10-dB improvement in phase-noise performance over the conventional approach. This is attributed to advantages in phase-noise performance offered by today's L-band VCO technology, as improved by a tight PLL loop filter, applied to the wide-range coverage of the microwave synthesizer.

A block diagram of the synthesizer is shown in Fig 1. An 84-MHz PLXO (phase-locked crystal oscillator) is used as the reference clock for the DDS-driven PLL. The choice of 84 MHz is intentional: The PLXO output doubles as a fixed LO to convert the first IF of 75 MHz to a 9-MHz second IF, thus realizing a fully coherent system (full synthesis).[2]

Although the design of a PLXO can be the topic of an entire article,[3] the approach taken is to lock an 84-MHz crystal oscillator to a 10-MHz oven-controlled crystal oscillator (OCXO) that exhibits outstanding stability performance.[4] The locked 84-MHz crystal oscillator then uses its own high Q to provide good phase-noise performance while maintaining the 10-MHz oscillator's long-term stability.

In addition, by purposely locking the 84-MHz PLXO to a 10-MHz reference, we may easily compare the entire transceiver's calibration against the 10-MHz WWV signal. Automating this process is also possible. The receiver is simply programmed to come up exactly on 10 MHz to facilitate a calibration check every time the radio is turned on. The transceiver's PLXO (also called the master reference unit —MRU) is left running forever for near-perfect stability.

Notice again that the system presented here is fully coherent. That is, all frequency sources in the transceiver are locked to the 10-MHz OCXO reference and, consequently, to the WWV signal. This kind of design is more demanding than those using several individual crystal oscillators as LOs. Not only are incoherent systems more expensive to implement as individual LO sources, but they can also drift independently of each other, causing the radio to drift uncontrollably (Note 2).

Let's now analyze the DDS-driven PLL design for its spurious performance. The 84-MHz choice for the DDS reference works well. The Nyquist criterion explained in Part 1 of this article has been met (8×10.5 MHz = 84 MHz). The spurious analysis in Fig 2 shows that there are no direct, in-band problems in the DDS output range of 7.7 to 10.5 MHz.

The DDS output frequency is stepped in 1-Hz increments over the range. Control is provided to the synthesizer by a powerful microprocessor running at 32 MHz. To minimize EMI to the receiver, this frequency was chosen to be above the HF passband.

The output of the high-resolution variable-source DDS (the U1 DDS) goes through a low-pass filter of elliptical design. This filter is intended to reject any out-of-band spurious problems that may still exist despite the careful reference choice. The variable reference signal is further conditioned by Q1, Q2, U3A and B to produce fast-rise-time square waves to be presented to the reference input of PLL U4.

A tight loop filter controls an L-band VCO at G1. This VCO covers the wide range of 700 to 1200 MHz. We use only a portion of this range: 770 to 1050 MHz. Although several VCOs could have been used here to improve phase-noise performance, a single VCO with superior *Q* proved simple to implement while providing good performance. The output of the VCO (running at one hundred times the DDS output frequency) is stepped in 1-Hz increments by the DDS. A fixed-frequency divider (made of two divide-by-10 stages) in the PLL chip further conditions the waveform for phase comparison.

In this design, the two divide-by-10 devices have been cascaded in the loop, for a total division of 100, to get better phase-noise performance. A clean 77- to 105-MHz LO range is derived after the first divider. The synthesizer step resolution is therefore 10 Hz (for 1 Hz at the DDS) over the entire range.

When looking at Fig 1, several things are important to notice. First, in implementing an L-band DDS-driven PLL synthesizer of this caliber, the variable reference produced by the DDS and presented to the PLL implies the use of a PLL device capable of accepting such higher reference frequencies.[5] Second, most PLL chips on the market today, contain a built-in programmable divider (called the R divider) in the reference input, which is intended to make it easy for the designer to tailor specific division ratios using the same reference. This programmability usually allows for division by binary numbers such as 2, 4, 8,

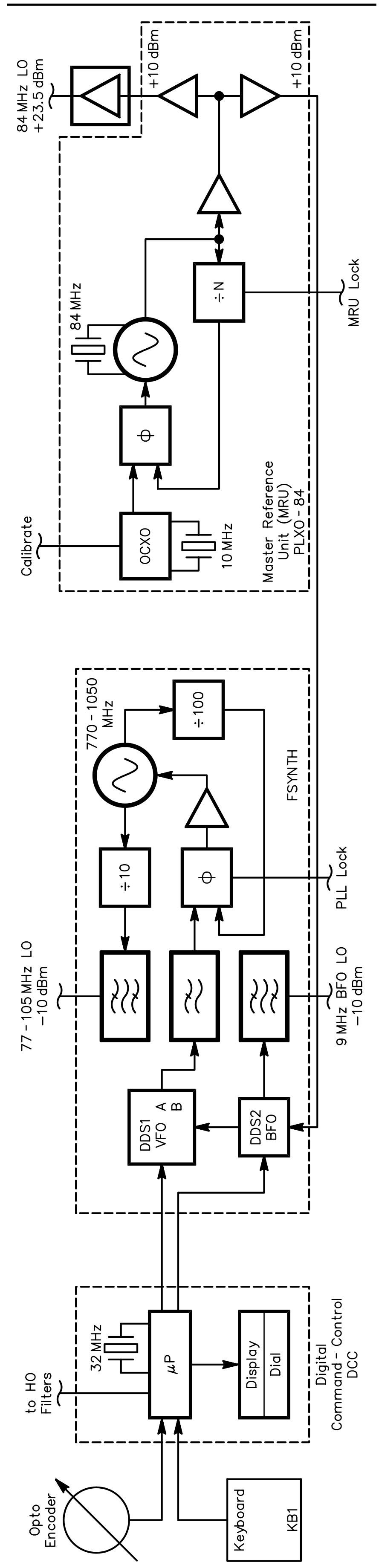

Fig 1—Block diagram of a microwave DDS-driven PLL synthesizer showing the PLXO master reference unit (MRU) and the command and control boards.

16 and so on. However, few of the PLL devices available on the market today allow for division by 1, which is needed in this DDS-driven PLL.

The above requirements restrict the choice of a PLL chip for this design. I used the SP-8855 PLL from Mitel. This choice proved worth the extra expense, since it satisfied both of the above conditions.

Circuit Description

The circuit for the microwave DDS-driven PLL and BFO synthesizer is shown in Figs 3 and 4. Both DDSs use the AD-9850 from Analog Devices. In this design, the word-clock information (WORD_CLK_1/2) and data (DATA_1, 2) are communicated serially to DDS1 and DDS2 from the command-and-control board via J1. The data are validated via the FQ_UD1/2 lines. The microprocessor control boards and circuits are not shown here because of the limited scope of this article, but they may appear in a future article. With over 8000 lines of code programmed in it, the command and control system manipulates the transceiver's over 290,000 synthesizer frequencies (including end-to-end split operation), as affected by the transceiver's various modes of operation, in less than 10 milliseconds. This is remarkable considering that for a worst-case scenario (split operation from say, 1.8 to 30 MHz), the microwave synthesizer has to travel and settle over 300 MHz in such a short amount of time.

The breadboard is shown in Fig 5. Several other functions such as IF shift (in both receive and transmit), receiver incremental tuning (RIT) and others are executed by the synthesizer under full control of the command and control system.

DDS1 (U1) is at the heart of the DDS-driven PLL. It outputs RF signals between 7.68 and 10.5 MHz with a resolution of 1 Hz. This range of frequencies is passed through a highly efficient low-pass filter, as shown. The filter has a corner frequency of 10.6 MHz and rejects the DDS out-of-band spurious products by at least 65 dB. The signals are further conditioned by Q1 and Q2 (Fig 4). The output of Q2 triggers gates U3A and B. These gates are fast 54S00s, which allow for sharp rise times. The conditioned digital 7.68-to 10.5-MHz signals are then terminated by R23 and R24 and finally presented to the reference port of the SP8855E PLL chip's phase detector. The chip allows for a divide-by-one setting using reference frequencies typically from 10 to 50 MHz. The special conditioning shown is used to actually lower the working frequency of the PLL from 10 to 7 MHz to accommodate this design to the internal configuration of the SP8855 PLL chip.

Coming out of the PLL chip are the charge pump signals (CPO and CPR) at pins 20 and 21. These signals are then presented to U5. The corner frequency for the loop filter of the PLL has been calculated at 10 kHz (other loop bandwidths have also been implemented). The loop gain is set at 1.5. The loop filter is made of U5 (OP27), a quiet choice. Additional filtering is done through a second-order filter made of R43, C38 and C71. The series trap, FB1 and C70, helps clean up other unwanted spurious responses. The calculations for the loop filter are shown in Fig 6.

The steering signal from the OP27 loop filter is presented to the single VCO, G1, that operates from 700 to 1200 MHz. The actual phase-noise performance of the G1 VCO is improved by the tight loop bandwidth of the PLL.

Again, despite the very wide coverage, this design makes use of a single VCO (instead of several) operating in concert with a quiet PLL at microwave frequencies. The system is further enhanced by a relatively clean VCO running in a tight and quiet loop. All this is done at 10 times the needed frequency range. Dividing the output by 10 yields a phase-noise improvement of 20 dB from the circuit.

DDS2 (U2, Fig 3) outputs a 9-MHz BFO signal. The exact frequency of the BFO DDS is dictated by mode selection in the transceiver. BFO frequency shifts required by the diverse modes are narrow enough to allow a band-pass crystal filter (FL1) to improve the spurious performance of the DDS.

As mentioned in Part 1, it may be argued that a DDS-driven PLL operating directly at 77 to 105 MHz could provide performance similar to the one presented here. The truth is that an advantage of 6 dB in phase-noise performance is gained by using a microwave DDS-driven PLL because of the reasons given above.

Finally, the output of the G1 VCO (Fig 4) is further conditioned by C42 and an attenuator pad made of R29, R30 and R31. The VCO output is then split and fed back to the PLL chip (U4) at pin 13. The PLL chip is set to divide by 100, so its output is lowered to the 7.68- to 10.5-MHz range. The L-band range is amplified by U6 and conditioned further through C47, R33, R34, R35 and C51. The RF signal is then divided by 10 in U7 (which has 5.5-GHz frequency performance) to provide an even-amplitude output signal over the entire range. A lower-frequency divider (1.5 GHz) was initially used with inferior performance at the higher frequencies (1 GHz and up).

The output of the divide-by-10 chip is further conditioned by a band-pass filter made of C53, C54, L4, C55, L5, C56, L6 and C57. Finally, the output

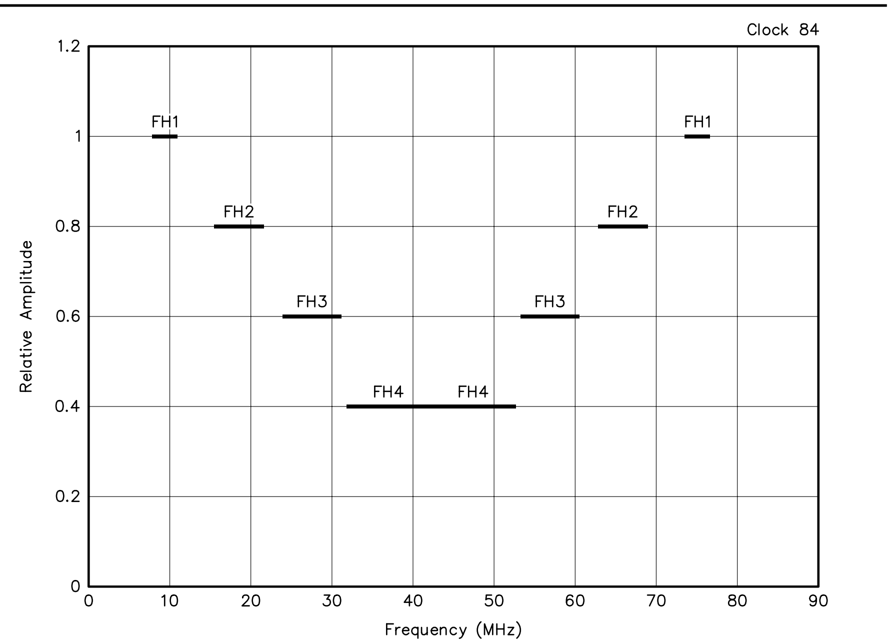

Fig 2—Spurious analysis of the DDS used in the DDS-driven PLL shows no in-band problems with the 84-MHz reference, for the DDS output from 7.7 to 10.5 MHz.

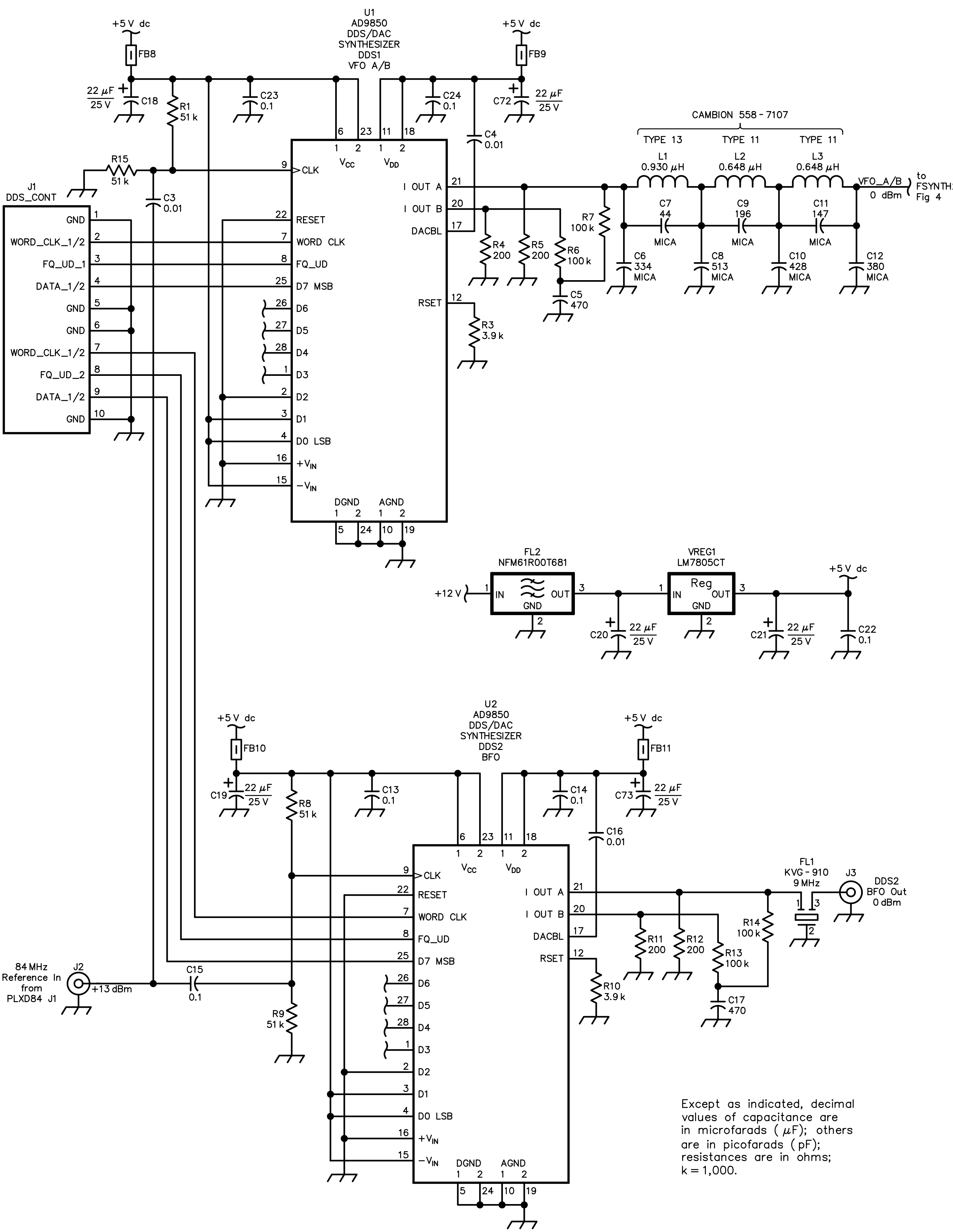

Fig 3—Schematic diagram of the microwave DDS-driven PLL FSYNTH1.

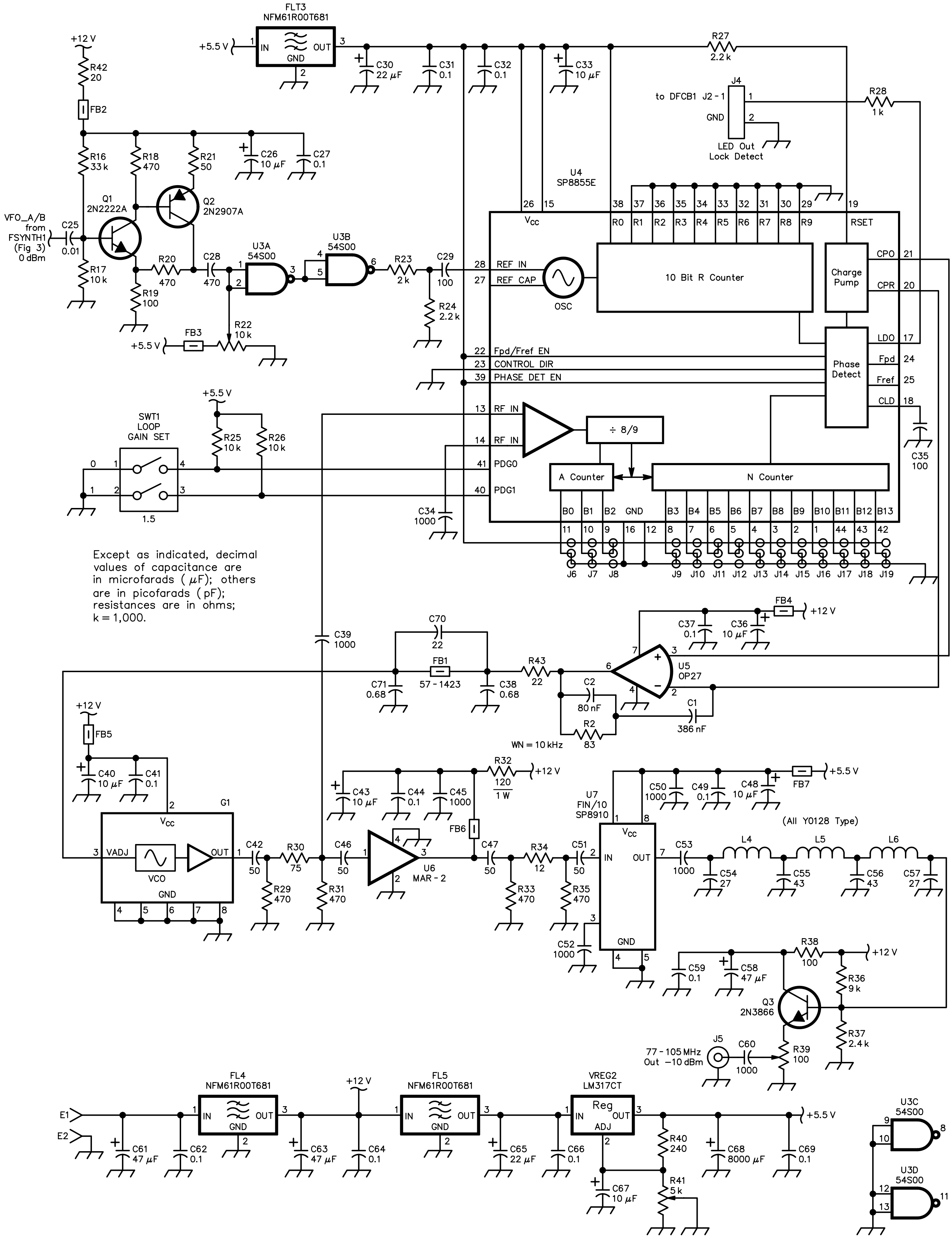

Fig 4—Schematic diagram of the microwave DDS-driven PLL FSYNTH2.

State-of-the-Art Synthesizer ICs

The Analog Devices AD9852 and AD9854 are among the fastest DDS chips around. (AD9850 is the workhorse of current designs.) The AD9852 boasts a 300-MHz maximum clock speed and a 48-bit phase accumulator. That results in a maximum tuning resolution of 1 µHz! The onboard reference-clock multiplier allows operation at top speed with a lower-frequency external reference. A 12-bit digital multiplier and a 14-bit PM register are included to facilitate modulation, as well as a single PSK input pin with programmable phase offset. Two 12-bit DACs and a comparator round out the chip's hardware complement. The comparator may be used to square the DDS DAC's synthesizer output for clock-generation applications or where AM spur elimination is sought. It may also be used in conjunction with the control DAC to allow PWM generation.

The AD9854 is only a little different from the AD9852. It provides two discrete synthesizer outputs for I/Q applications. It therefore has two 12-bit multipliers. The manufacturer claims excellent SFDR and tuning speed for both these devices. Additional detail may be found on AD's Web site: www.analogdevices.com.

QEX has learned that Analog Devices is close to releasing a DDS chip like those described above and having a 1-GHz clock speed. While details are not yet publicly available, our sources indicate the device will become available later this year.

Fractional-N synthesizer ICs have also shown performance increases and greater acceptance among designers. Outstanding among these is the line from Conexant, formerly Philsar. Their CX72302 supports operation at greater than 6 GHz with less than a 400 Hz tuning step size. Internal reference frequencies up to 25 MHz are allowed for extremely fast lock times. Other members of the Conexant family, such as the CX72300, are dual fractional-N devices that provide for both first and second LOs in many microwave radio architectures. More information is available at www.conexant.com. Also look for offerings from others, such as TI, National, Philips and so forth.—*Doug Smith, KF6DX*

Table 1—Loop Performance versus Loop Bandwidth

Loop Bandwidth	***Circuit Values****			*Phase-Noise Improvement†*
Center f (kHz)	*C1(nF)*	*C2 (nF)*	*R2 (Ω)*	
100	3.84	0.794	829.40	Nominal (minimum –120 dBc/1 kHz)
50	2.65	0.547	240.77	6 dB (minimum –126 dBc/1 kHz)
10**	**385.3**	**79.79**	**82.622**	**10 dB (minimum –130 dBc/1 kHz)**
5	1541.05	319.16	41.311	
2	9631.56	1994.76	16.52	
1	3852.62	7979	8.26	

*Use polystyrene or silver-mica capacitors and matched, film precision resistors
†Reference source must be better than –130 dBc/1 kHz
**Chosen for this design, loop-bandwidth choice is a compromise between phase-noise and lock-time performance.

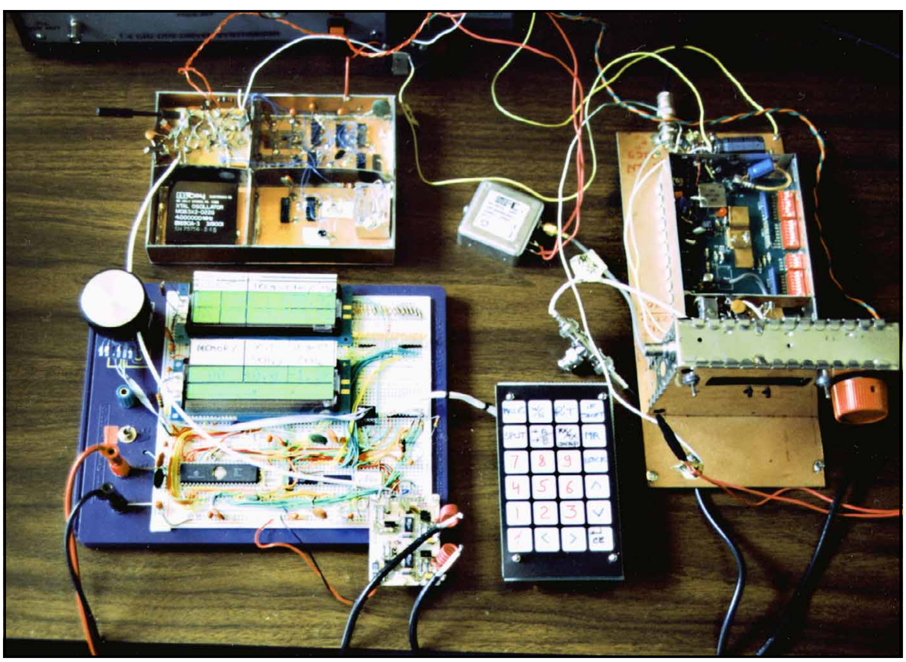

Fig 5—The breadboard version of the microwave DDS-driven PLL, showing the master reference unit and the command and control implementation. For the finished product, see the cover of this issue.

of the synthesizer from 76.8 to 105 MHz (in 10-Hz steps) is amplified to a –10 dBm level by Q3, a 2N3866.

Performance

The phase-noise performance of the microwave DDS-driven PLL meets the criteria set earlier. This performance is typical of any frequency within the range. The spurious performance of this synthesizer is better than –76 dBc anywhere from 76.8 to 105 MHz. The end-to-end lock time of the synthesizer is less then 10 milliseconds over the entire range, allowing for unconditional split operation over a 30-MHz delta frequency. The frequency resolution is 10 Hz anywhere in this range.

Conclusion

This series presented various state-of-the-art synthesizer techniques used by the HF receiver and transceiver industry. In addition, the article explained in ample detail the role of the

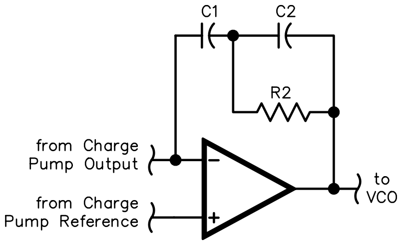

Fig 6—Microwave synthesizer's third-order loop-bandwidth calculations (see Table 1).

Third-Order loop filter with

$$f = 5\text{ kHz};\quad N = 100;$$

$$k_0 = 2\pi(10\text{ MHz}) = 62831853.1$$

$$\text{Phi} = \frac{6.3\text{ mA}}{2\pi} = 1.0027\text{ mA/radian} = 0.001\text{ A/radian}$$

$$\omega_n = 2\pi f = 31415.927\text{ Hz} = 31.416\text{ kHz}$$

$$\Phi_o = 45^\circ$$

$$T_3 = \frac{-\tan\Phi_o + \dfrac{1}{\cos\Phi_o}}{\omega_n} = 1.31848\times 10^{-5}$$

$$T_2 = \frac{1}{\omega_n{}^2 T_3} = 7.6847\times 10^{-5}$$

$$T_1 = \frac{k_0 k_\Phi}{N\omega_n{}^2}\left[\frac{1+\omega_n{}^2 T_2{}^2}{1+\omega_n{}^2 T_3{}^2}\right]^{\frac{1}{2}} = 1.54105\times 10^{-6}$$

$$C1 = T_1 = 1.54105\ \mu\text{F}$$

$$\frac{T_2}{R2} - C1 = \frac{T_3}{R2}\ ;\quad \frac{(T_2 - T_3)}{R2} = C1\ ;\quad \frac{(T_2 - T_3)}{C1} = R2$$

$$R2 = 41.311\ \Omega$$

$$C2 = \frac{T_3}{R2} = 3.19162\times 10^{-7} = 319.162\text{ nF}$$

DDS in present and future frequency-synthesis methods applied to HF radio design.

A novel microwave DDS-driven PLL design was introduced as a means of improving the performance of high-resolution, wide-band synthesis. In addition, new and novel ways of utilizing a DDS alone as a wide-band, high-resolution frequency synthesizer have been suggested. I envision that DDS-only synthesizers will become the rule—rather than the exception—in receiver and transceiver design in the very near future. I hope this article series stimulates creative thinking about future products as well as homebrew projects.

Acknowledgments

I would like to express my thanks to Richard Fesler and Silvio Cardero for reviewing this material. Additional thanks go to Constantin Popescu, KG6NK, and Randy Burcham, KD7KEQ, for their help in "bread-boarding," troubleshooting and laying out the boards for the prototype.

Notes

[1]C. Drentea, "Beyond Fractional-N, Part 1," *QEX*, Mar/Apr, 2001, pp 18-25.

[2]In a fully coherent radio, all LO sources in the synthesizer are derived from a single reference, the MRU. This avoids random frequency drift associated with multiple oscillators. While such design is more demanding than the "scattered-LO approach," more and more modern radios are pursuing it regardless of their class (for example, the Alinco DX-77).

[3]C. Drentea, "High Stability Local Oscillators for Microwave Receivers and other Applications," *ham radio*, November 1985, pp 29-39.

[4]Stability and phase-noise behavior of a 10-MHz crystal oscillator is better than at higher frequencies. Such an oscillator is the best choice as a frequency standard because of its superior phase-noise and spurious performance; those traits can be better controlled at this frequency.

[5]Most PLL chips on the market today only accept much lower reference frequencies.

Cornell Drentea was born in Bucharest, Romania. He studied electrical engineering in that country and the United States and has been at the forefront of developing state-of-the-art RF products for over 40 years. These include receivers, synthesizers, transmitters and transceivers at frequencies up to 100 GHz.

During his career, Cornell has worked for several companies, including Honeywell and Hughes Aircraft. He is currently a Senior Principal Engineer in the Transmitters, Receivers and Exciters Group of Raytheon Electronic Systems, where he leads design and development of ultra wide-band receivers, synthesizers and ultra high frequency digital signal processors (DSP).

Cornell has published over 60 technical articles and papers in various trade and ham magazines. He is also the Author of Radio Communications Receivers. Cornell has been a ham for over 40 years and holds an extra-class Amateur Radio license. He lives in Tucson, Arizona.

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Distributed Processing Goes Galactic

Are we the sole sentient species in the cosmos or might there be others? Take a look at how you can help find the answer.

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

[Editor's note: This paper first appeared in the Proceedings of the Trenton Computer Festival 2000*, May 6, 2000.]*

If there is indeed an interstellar Internet, might we someday log on? What are the protocols for cosmic communication? Now, for the first time in human history, we have the technology to ask—and perhaps begin to answer—these questions. In this paper, we explore the strengths and weaknesses of "SETI@home," the most ambitious distributed-computing experiment on this planet. We learn how thousands of Amateur Radio telescopes are forming a global net to snare that elusive fish in the cosmic pond. In addition, we explore how the lessons learned from the SETI @home experience can be brought to bear on the problem of massive data collection and analysis. I believe it is the world's radio amateurs and computer hobbyists who will ultimately bring in signals from the stars.

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SETI 101: An Introduction

SETI, the electromagnetic Search for Extra-Terrestrial Intelligence, is a relatively young science with a colorful history that seeks to detect direct radio evidence of other technological civilizations in the cosmos. For 40 years, its dominant paradigm has been the use of giant radio telescopes, having sensitive microwave receivers and powerful computers, to scan nearby stars for signals of intelligent origin. Once funded through NASA, SETI research in the United States lost its government support eight years ago, and now continues as a private venture conducted by various grass-roots, nonprofit organizations.

Giant radio telescopes (such as the 1000-foot spherical reflector of the Arecibo Observatory in Puerto Rico) achieve part of their sensitivity by directing an extremely narrow beam on the heavens. Such instruments view perhaps one millionth of the sky at a given time, reducing the received background noise, hence improving the signal-to-noise ratio of any detected radio artifact by a factor of a million relative to an omnidirectional (isotropic) antenna. If you have such

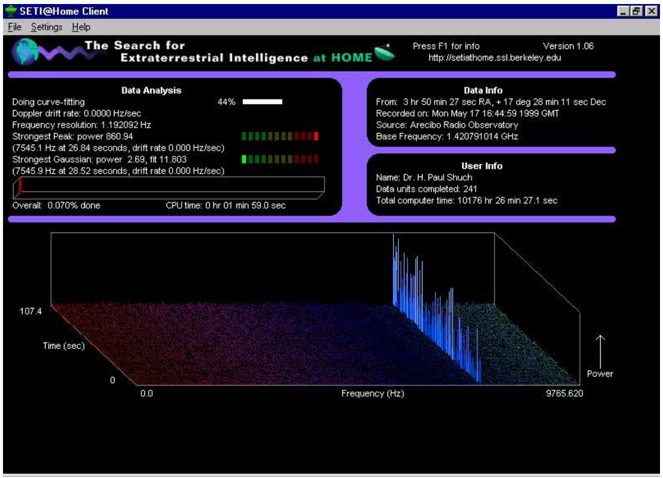

Fig 1—Strong, coherent signals such as this one quicken the pulse of many SETI@home project participants. Unfortunately, all so far have been generated not by ETI, but by terrestrial interference, or by the Wizards of Arecibo as they inject test signals to verify the proper operation of their equipment. (Source: **www.setileague.org/photos/homehit.jpg**)

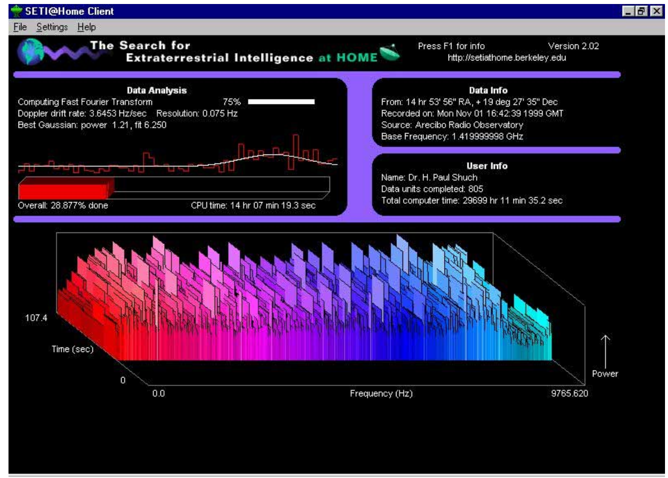

Fig 2—The SERENDIP receiver at Arecibo has a 2.5-MHz instantaneous bandwidth. For SETI@home processing, its output is parsed out into 256 subbands, each 9765 Hz wide. Notice how the amplitude of the noise rolls off at both the top and the bottom frequency ends of this analysis spectrogram. The curve seems to follow the frequency response of an audio band-pass filter optimized to the desired subchannel bandwidth. (Source: **www.setileague.org/photos/home_bpf.jpg**)

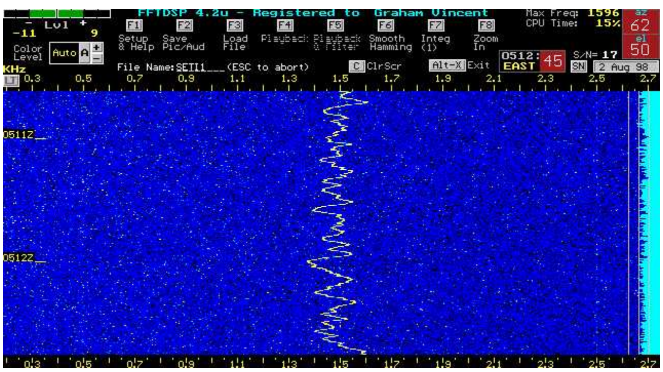

Fig 3—Graham Vincent, SETI League volunteer coordinator for New Zealand, received this intriguing signal on 2 August 1998, at a frequency of 1281.919 MHz USB (in the 23-cm Amateur Radio band). The appearance of the signal is similar to a class of anomalies detected by the SETI Institute's Project Phoenix targeted search. Dubbed "wigglers" by the SETI Institute's Dr. Jill Tarter, they have always proved to be cases of radio frequency interference. Graham's signal is no exception. It turned out to be computer RFI, generated within the very computer he was using to run his signal-analysis software. England co-coordinator Ken Chattenton, who has had similar experiences, recommends that if a signal is strong enough to be audible, one should turn off the computer and see if it goes away. (Source: **www.setileague.org/photos/wiggler.jpg**)

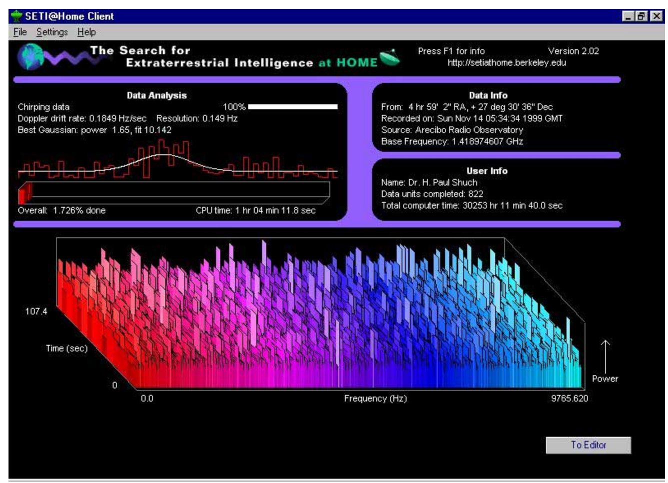

Fig 4—Here's a signal with a weak Gaussian fit, which is not evident from viewing just the 3D spectrogram (bottom window). The SETI@home client divides the 9765-Hz-wide data block into thousands of very narrow bins. The amplitude of the signal in each individual bin is analyzed over time, and the bin with the best fit to the antenna's expected drift-scan time series is displayed as a ragged trace in the data Analysis (upper left-hand) window. The smooth trace represents an ideal Gaussian curve (normal distribution) corresponding to the pattern of the Arecibo antenna. The two curves are statistically compared. The closer the fit, the more credence is given a candidate signal. Of course, the Gaussian test is only one of many hurdles a signal must pass before it is considered to be of extra-terrestrial origin. (Source: **www.setileague.org/photos/homegaus.jpg**)

an antenna hooked up to the right kind of receiver tuned to precisely the right frequency, at exactly the instant when "The Call" comes in, there is about a 99.9999% chance it will be pointed the wrong direction.

Since we don't know what exactly that right frequency is, the problem of SETI's success is complicated further by our need to tune our receivers systematically across a wide spectral range. If a SETI receiver is to achieve reasonable sensitivity, its desired reception bandwidth is, of necessity, quite narrow. This is because radio noise and natural interference phenomena are broadband, while one of the hallmarks of intelligently generated emissions is high spectral coherence, resulting in narrow bandwidth (on the order of a fraction of a hertz.) Unfortunately, that narrow signal may fall anywhere within several gigahertz of potential spectral real estate.

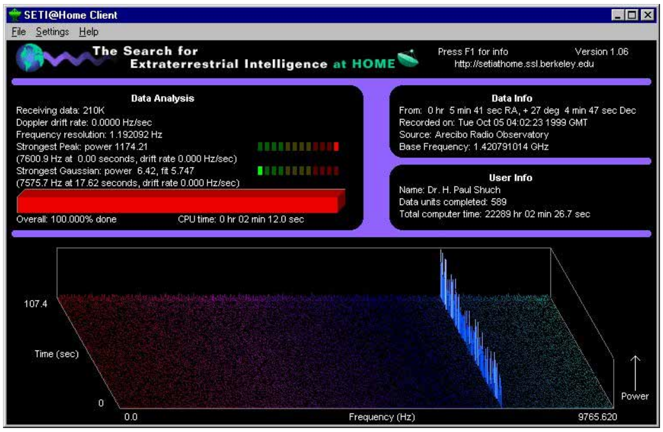

Fig 5—SETI@home's over two million users continue to see occasional anomalies such as this one, observed by the author in December 1999. Members sometimes call or e-mail The SETI League, requesting that we check such signals (most of which turn out to be terrestrial interference). Unfortunately, we can do nothing from here to analyze these because all verification is performed by the SETI@home team at Berkeley, California. Be sure to upload your analysis files to them, and rest assured that they will indeed follow-up on all interesting candidate signals and inform you if yours is The One. (Source: www.setileague.org/photos/homehit3.jpg)

Fig 6—The typical computer takes tens of hours to fully analyze a single SETI@home data block. Occasionally, strong, wideband terrestrial interference obliterates any useful information. When that happens, the SETI@home client determines that no further analysis of that data block is possible, quickly terminates analysis of that particular file and requests another one for analysis. SETI@home gave up on this file after just five minutes. (Source: www.setileague.org/photos/home_qrm.jpg)

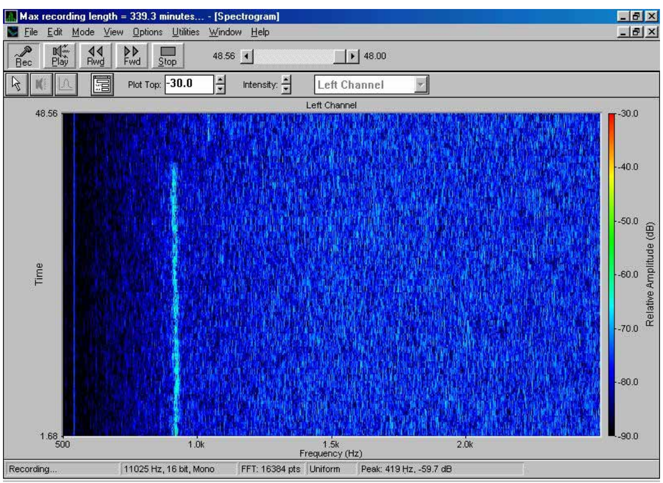

Fig 7—EME (moonbounce) contests provide Project Argus participants with an opportunity to detect weak amateur microwave signals reflected off the lunar surface. This unusually strong 1296.015-MHz EME echo from the 30-foot dish of Jay Leibmann, K5JL, was received at Argus station FL11LH during the 30 October 1999 ARRL EME contest. (Source: www.setileague.org/photos/k5jl-cq2.jpg)

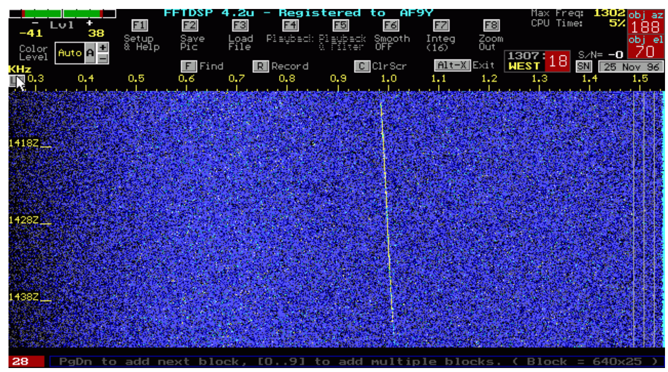

Fig 8—This CW signal from the Mars Global Surveyor was received by SETI League member Mike Cook on 25 November 1996, while the spacecraft was about 5 million kilometers from Earth. The satellite's 1.3-W beacon transmitter—feeding an omnidirectional antenna—provided SETI enthusiasts with an excellent dry run to verify the operation of their receivers and DSP software. Several other SETI League members also recovered the signal using Mike's FFTDSP shareware program. (Source: www.setileague.org/photos/mgs14.gif)

If we wish to find an artificial signal occupying a bandwidth of, say, 1 Hz, and if to locate it we must scan, for example, 10 GHz of spectrum, then we have a temporal problem analogous to the previously stated spatial problem. If we are pointed in exactly the right direction at the instant when The Call comes in, there is something like a 99.99999999% chance we'll be tuned to the wrong frequency!

Problems associated with the spatial dimension of SETI are partly overcome by constructing phased arrays of a great many antennas, operating so as to look in many directions at once. To address the spectral challenge, SETI engineers concentrate much of their efforts toward developing elaborate multichannel spectrum analyzers (MCSAs) capable of monitoring many millions of narrow channels simultaneously. Despite these technological advances, SETI critics rightly point out that after 40 years of SETI, we have yet to detect a single confirmed signal of intelligent, extraterrestrial origin.

The SETI@home Experience

In those 40 years of searching, SETI proponents counter, we have not only failed to scratch the surface, we haven't even felt the itch. Our massive antennas and multichannel spectrum analyzers generate more data than we can ever hope to analyze, even using Earth's most powerful supercomputers. Digital signal-processing efforts, necessary to separate the cosmic wheat from the galactic chaff, depend upon an ability to crunch numbers at an ever-increasing rate. Right now we are generating more

SETI data than we can ever hope to analyze. It has been argued that if we are to be limited at all, computing power is the place to be limited. Since computing power seems to double every year or so, we need merely wait until our computer power is up to the task: in just a few more years or decades—or centuries.

One group of "SETIzens," lead by Professor Woody Sullivan at the University of Washington and Dan Werthimer, David Anderson and David Gedye at Berkeley, got tired of waiting. In 1999, they launched SETI@home, planet Earth's most ambitious distributed-processing experiment. They have harnessed the idle processing power of three million personal computers around the world; in so doing, they created the world's most powerful super-computer. Tied to a Project SERENDIP microwave receiver at the Arecibo Observatory, the SETI@home network has crunched more data in a few short months than has been analyzed by all the world's prior SETI efforts combined.

After preprocessing on site, SERENDIP cranks out data at a rate greater than a megabyte per second. That is just over 100 CD-ROMs per day of data generated by one receiver alone: truly an example of drinking from the fire hose. Until now, most of that data simply would not have been analyzed.

SETI@home's power derives from clever software that parses this massive data stream into bite-sized chunks for Internet distribution and off-line data analysis. This is done by first filtering and dividing the receiver's 2.5-MHz bandwidth into 256 individual audio channels, each a manageable (and easily digitized) 9756.6 Hz wide. A 341-kB digital file can store 107 seconds worth of data from just one of these audio channels. At 56 kbps V.90 dial-up connect speeds, it takes about a minute and a half to transfer such a packet to a SETI@home user via the Internet. Then a modest computer running the SETI@home analysis software can thoroughly analyze it off-line. On a typical Pentium III-class PC, this analysis takes roughly ten hours, after which the user uploads a results file to the SETI@home server at Berkeley, downloads another packet and starts all over again. Working together is certainly working!

Today, nearly two million home computers are devouring data from the world's largest radio telescope, terabytes at a time. Still, while the screen saver churns away in the background, the appetite for involvement is not sated. "I'm no rocket scientist," I hear you saying, "but I want to do more than wait for my Pentium to claim the prize. Where can I go from here? The software is fully capable of discovering that elusive needle. Only, where do we find the haystack?"

SETI@home has some impressive strengths offset by one significant weakness. Its nearly three million home computers are crunching data from a single source: one antenna that "sees" only a tiny fraction of the sky at any time. To avoid missing The Call,

What Sort of Signal Processing Does SETI@home Do?

SETI@home data are taken from the Arecibo telescope using a very sensitive, cryogenically cooled receiver (45 K). The receiver's 2.5-MHz bandwidth is centered on 1420 MHz. That is the 21-cm hydrogen line where many researchers suggest it is most likely interstellar transmissions would be lurking. Arecibo's 305-meter dish is fixed with respect to the Earth; the SETI@home survey covers only about 28% of the sky. Over several years, each section of sky is examined several times. The system's beamwidth is about 0.1°! Provision is made for the feedpoint to track the rotation of the Earth so that accurate position information is available, but only over short time frames, say 12-24 seconds.

Signals are quadrature down-converted to baseband and filtered using a pair of 256-tap FIR filters. The resulting 2.5 MHz of spectrum is broken into 256 sub-bands using a 2048-point FFT and 256 eight-point inverse FFTs. From these, 107-second work units are built; those are the data sent to participating client computers around the world.

SETI@home software parses the data in 15 octaves of frequency and over discrete time spans of 0.8 ms-13.4 s. Because intelligent extraterrestrial signals are likely to originate on a body that is rotating, the data are examined for Doppler shifts that change or "chirp." Chirp transforms are applied to the data and FFTs are used to analyze them. FFT lengths range from 8 to 131,072. Power spectra are computed and peak-searching algorithms note any peaks greater than 22 times the mean power. Those are marked for further analysis by the SETI@home team. Gaussian beam fitting is also computed at every frequency, chirp rate and resolution. Chunks of data meeting certain requirements are also noted for further examination.

Most of the signals noted this way turn out to be terrestrial RFI. Many RFI sources have been heard and characterized before, so the software deletes them and has another look at the data.

It is perhaps interesting that many SETI techniques are applicable to two-way Amateur Radio communications, especially EME or moon-bounce. The concepts of Doppler "dechirping" and long-time integration of signals are now being deployed there. Other, more esoteric DSP methods, such as homomorphic deconvolution, may find use. In addition to cleaning up EME signals, a benefit of such deconvolution is that it gives you a picture of the reflecting surface—in this case, the Moon—and data about its motion. That information would constitute valuable scientific data that could be used by astronomers at high levels of research. It would also bolster our status as experimenters at the leading edge of technology.

Signals from beyond the solar system that passed all the tests were first detected in 1967 by Anthony Hewish and Jocelyn Bell at Cambridge University; they caused the first big SETI stir because the signals' periods were so precise. We now understand that those signals are produced by rapidly spinning neutron stars (pulsars). Neutron stars are the remnants of old, giant stars that collapsed under their own weight and exploded quite violently to become novas and supernovas. So while those coherent signals contain information, the information is apparently not generated by intelligent beings.

Thus far, many candidate signals have been examined, but none has met the criteria for containing intelligence. In all our inductions about what extraterrestrials (ETs) would do, we must wonder whether they might decide that announcing their presence is not a smart idea. It may be that an advanced civilization would not try to contact us deliberately via radio signals. But if they did, should we respond?

For more technical details, visit **www.setiathome.ssl.berkeley.edu/sciencepaper.html**—*Doug Smith, KF6DX, QEX Editor.*

we really need about a million such radio telescopes, coordinated to listen in all directions at once. Nevertheless, at a cost of perhaps a $100,000,000 apiece, we'd exceed the Gross Planetary Product. Fortunately, there is another way.

Project ARGUS and the Amateur Radio Astronomer

Launched in 1996, Project Argus is an amateur-run all-sky survey that attempts to accomplish something NASA SETI never contemplated: see in all directions at once. This major initiative of the membership-supported, nonprofit SETI League seeks to harness the power of 5000 small radio telescopes worldwide, in a coordinated search of all 4π steradians of space. Its Amateur Radio telescopes are typically built around discarded satellite TV antennas for a few hundred to a few thousand US dollars. They achieve sensitivities on the order of 10^{-23} watts per square meter, roughly equivalent to the best research-grade radio telescopes of the late 1970s. As personal computers and DSP software become more powerful, this two-decade gap between professional and amateur capabilities is beginning to narrow.

One argument for the validity of Project Argus is the example set by amateur optical astronomers in their discovery of numerous comets, supernovae and other highly intermittent astrophysical phenomena. These events are not typically discovered by the world's great observatories, but rather by dedicated amateur astronomers. Allen Hale and David Levy both use 14-inch Schmidt-Cassegrain telescopes. Tom Bopp doesn't even own a telescope, but codiscovered Comet Hale-Bopp with one he borrowed from his astronomy club. Yuji Hyakutake discovered the comet that bears his name with a pair of high-power field binoculars! The late Gene Shoemaker was a geologist by trade but a longtime and avid amateur sky-gazer.

Regrettably, the analogy breaks down when one considers equipment availability. In most cities of the world, an aspiring comet-hunter can walk into a local optical shop, write a check for $1000 or so and walk out with a telescope that would have turned Galileo green with envy. Amateur Radio astronomers aren't quite so fortunate. You can't walk into your local RadioShack store and buy a radio telescope—at least not yet. The SETI League is trying to change that by designing the hardware (licensed for commercial manufacture) and software (distributed as shareware via the Internet) to turn a surplus 3- to 5-meter TVRO dish and a home computer into a credible research instrument. About 100 radio amateurs, microwave hobbyists, electronics experimenters and computer hackers around the world have already succeeded in putting their Argus stations on the air. Hundreds more stations are now under construction, and the dream of all-sky coverage (whereby no direction on the sky shall evade our gaze) comes closer to reality every year. Construction details of a Project Argus radio telescope appear in the SETI League Technical Manual. It's available in hardcopy for a modest contribution or may be downloaded free from the Web at **www.setileague.org**.

Global and Galactic: The ARGUS@home Concept

Current Project Argus instruments each scan about 22 kHz of frequency spectrum at a time—a small fraction of the 2.5-MHz instantaneous bandwidth of the SERNENDIP receiver at Arecibo. They typically break that spectrum down into 8192 simultaneous channels, each about 2.5 Hz wide. One such instrument generates on the order of 44 kB per second of data. This is a small fraction of the data gathered by the SETI@home experiment at Arecibo. On the other hand, the existing 100 Argus stations, collectively, already approach the data output of the SETI@home receiver. By the time Argus reaches full strength, its combined network of 5000 Amateur Radio telescopes will collectively generate as much data as about a hundred Arecibos!

The SETI@home packet your PC is processing came from the world's largest radio dish. So did everybody else's. That means three million PCs are being serviced by a single data source. It is a powerful source to be sure; but with lotteries all over the world, why buy all our tickets for a single drawing? Remember that Arecibo achieves its sensitivity by scanning a slim slice of the celestial sphere. No software in the world is going to find photons that didn't hit the fan, no matter how many computers are running it.

Perhaps that's where the eyes of Argus can really shine. Imagine a global network of thousands of Amateur Radio telescopes, scanning the entire sky in real time. Now imagine something akin to SETI@home software that will let you scan that data via the Internet. Only instead of archival data recorded weeks ago, we're talking about live data that your computer can capture in real time. Therefore, you need not wait for the evening news to hear the winning numbers.

ARGUS@home won't happen overnight, any more than SETI@home did. In addition to the multitude of small radio telescopes required, we still need to come up with a SETI@home-compatible data block format and a way for Project Argus software to parse out the gathered data for Internet distribution. Then there's the challenge of collecting and correlating all those processed packets. The SETI@home experiment has already solved many of these problems; it remains for The SETI League to adapt their solutions to amateur practice. We hope that by the time SETI@home drinks the Arecibo well dry, we will have risen to these technical challenges.

Conclusion

Project Argus went online just five years ago with only five small Amateur Radio telescopes. Today we're running about a hundred. It's going to take us a few more years before the Argus network grows to truly global proportions. Until then, there's always Arecibo. The distributed-computing concept pioneered by SETI@home is very adept at finding needles. The global network of Argus telescopes will be ideal for finding haystacks. It seems to me that it's a marriage made in heaven. □□

Making a Sound Card Work for You

It's not that difficult—if you have the right information. Share the author's practical experience while developing a program to measure polar antenna patterns.

By Bob Freeth, G4HFQ

This article describes how to access and control a sound card under *Windows* using *Visual Basic*. It is based on the practical experiences of the author while developing a program to measure the polar radiation pattern of a beam antenna. It gives a brief history of how the program came about, together with the principles of its operation.

The structure of the sound-card mixer and the basics of how to access and control it are described. Some pitfalls to be aware of are noted if you want to do it yourself. References to useful sample programs are given to put you on the fast track to making a sound card work for you.

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The Beginning

About 10 years ago, a friend, Bill Sykes, G2HCG, acquired a transceiver that had the ability to connect to a computer using a dedicated interface. One of the things the transceiver could do was provide an S-meter reading to the computer. Bill, being a man much interested in beam antenna design and performance, thought of taking a series of S-meter readings whilst the antenna was rotating to plot the polar diagram of the beam. Could it be done? Bill knew little about computers, but a lot about antennas; I was in the opposite state, and so the project began.

To cut a long story short, the result was a very useful polar diagram. Several drawbacks kept it from being made generally available to the ham community: It worked fine only for that particular model of transceiver. Each transceiver's S-meter needed to be individually calibrated, and a specialized transceiver-to-computer interface was required. In addition, there were not as many PCs in use as there are today. So, Bill had a very useful tool all to himself to better understand the characteristics of his many and various designs.

The Next 10 Years

Over the next 10 years, PCs and software developed at a dramatic pace, and more people had access to a machine with a sound card. SSTV and weather-satellite programs using sound cards became readily available.

Multimedia software became commonplace.

This set Bill thinking again: If there were a reasonable correlation between the RF level input to the AF level output, anyone with a half-decent receiver and standard PC could plot polar diagrams. A few tests confirmed the theory and I was, once more, "prodded" into action.

Principles of Operation

To put the remainder of this article into a real-life context, the following extract from the program's user guide describes the principles of operation:

> "Plotting a polar diagram involves recording a received signal strength at known intervals, in synchronism with the rotation speed of the antenna. To do this you require a signal source (from either a transmitter or signal generator) and a receiver. The audio output of the receiver is connected to the line-in or microphone jack of a sound card in a computer.
>
> Reception of a plain unmodulated carrier as a beat note in SSB or CW mode gives a good correlation between RF input and AF output, as long as the AGC circuits are inoperative and there is no overloading. Modulated tones in AM or FM cannot be used, particularly FM, which is inherently designed to maintain audio level regardless of the RF input level. The program relies on the linearity of the receiving audio system for accuracy of plot and the measurement of gain. Whilst the linearity of the average sound card is generally quite good, the linearity of the receiver depends on how it is operated.
>
> To measure the transmitting station's polar diagram, the receiving station's antenna remains stationary and the transmitting station's antenna rotates. To measure the receiving station's polar diagram, the transmitting station's antenna remains stationary and the receiving station's antenna rotates. Unfortunately, every rotator moves at a different speed, dependent on make and condition, and different runs may change in speed as the rotator warms up. The program has facilities to compensate for these differences.
>
> The program takes a given number of readings per plot. To allow for different rotator speeds the time taken for these readings is variable under user control. Precise synchronization of the program's beginning and end of plotting with the rotator's time is difficult. To cater for these differences we have the ability to 'trim off' unwanted readings at the start and end of the plot. You can optionally have the program automatically start plotting when a carrier is detected and stop plotting when carrier is dropped. This mode of operation is particularly useful when plotting the transmitting station's antenna, because you do not know when the station operator starts and stops his rotator.
>
> Gain is calculated by measuring the half-power beamwidth of the antenna and converting the reading to decibels with respect to a dipole reference (dBd). As well as other factors, the gain of an antenna is principally a function of horizontal and vertical beamwidth. The program is not aware of the full details of the antenna design, for example, whether it is a stacked array, quad or simple Yagi, its element spacing and so on and is simply reading the level of received signal of a given polarity, it can only provide an indication of gain…"

Research Undertaken

Having never previously had any need to access a sound card by program, the first problem was finding out how to do it. This proved to be a very time-consuming process and an "idiot's guide" to programming sound cards has yet to be found!

Fortunately, being a subscriber to the Microsoft Developer Network (MSDN) and using *Visual Studio Enterprise Edition*, help was at hand. MSDN is an excellent library of information about Microsoft developer products and contains practically anything you want to know—if only you can find it! Most of the routines implemented for sound-card access were based on example code from the MSDN library.

The other research source used was the Internet, though very little of practical use was found. This may have been because of ineffective search criteria or not knowing the best places to look. It is sufficient to say that only one article surfaced that went a little way to help with sound-card programming. Using other people's compiled code carries a degree of risk unless you also have the source code, so that ruled out using several routines that might have been useful.

Various Internet newsgroups were tried but with zero results. I also tried talking to other programmers; but, being mostly involved in hard-core commercial applications, they were blissfully unaware of how to access a sound card.

What was Needed from the Sound Card?

Among the basic needs was to be able to choose which sound card input to use. Sound cards usually have at least two input sockets accessible at the rear of the PC: line-in and microphone. Sometimes others are available, variously called auxiliary, phone and so on, which are generally pins directly mounted on the card or the computer's motherboard. Receivers often have a fixed level audio output designed for a phone patch or tape recorder, as well as headphone and external-speaker outputs. To give some flexibility as to which of the receiver's various audio outputs best matched the inputs of the sound card, a method was needed to choose, under program control, which one to use for collecting data.

We also needed to read the instantaneous level of the AF signal from the receiver, whichever input was chosen; we needed to control the level of that signal as "seen" by the program. Finally, we needed to cater for differing peak input levels by controlling the sound card's amplification level overall. The actual waveform of the received signal was of no interest, only the level.

Essential Reference Material

Gone are the days of directly controlling the sound card using assembler, I/O addresses and IRQs. To do things properly, you do it by talking to the *Windows* mixer services. So how is it done with *Visual Basic*?

The first thing to realize is that there are no standard *Visual Basic* routines to enable you to access sound cards, so don't waste time looking for them. You use calls to the Windows mixer services found in *winmm.dll*, which are documented in the Windows Multimedia Platform Software Development Kit (SDK). Unfortunately, even there, the meanings of the return codes from calling these routines are not easily found.

The most comprehensive definition found of the structures required is defined in the *C* language. These, together with the meanings of the return codes and the multiplicity of different devices and controls, can be found in mmsystem.h. This is the *C* language "include" file for Multimedia APIs. If you have access to this file, I recommend you print it out—all 60 pages—and use it as your prime source of reference for control types, error codes and so forth.

If you have no access to mmsystem.h, all is not lost; there are two essential sample programs on the Microsoft Web site in the MSDN Online area. The first is article number Q187673, "AUDIOLVL.EXE—Monitor Input and Output Audio Levels" (**support.microsoft.com/support/kb/articles/Q187/6/73.asp**), the second is Q178456, "VOLUME.EXE—Set Volume Control Levels Using *Visual Basic*" (**support.microsoft.com/support/kb/articles/Q178/4/56.asp**). Without something like these two samples, you will not get very far in comprehending what to do programming-wise.

Another very useful diagnostic tool also available from the same source will help you understand how your particular mixer is configured. It is a sample program (written in *C*) called *Mixapp.exe*. This utility displays everything about your mixer, showing how the various components are linked together, the names of the controls and the component types, among other things. It was indispensable in visualizing the interconnection of individual mixer components and in debugging the program.

Talk to the Mixer

To get anywhere, you must talk to the mixer, a device for controlling audio lines. The standard software mixer supplied by Microsoft is called the Volume Control and is found in **Start|Programs|Accessories|Multimedia**. You can also start it by double-clicking the Volume Control icon that usually sits in the System Tray.

In a mixer (see Fig 1), there are two kinds of lines: *source* lines and *destination* lines. A number of source lines can be attached to, or associated with, a destination. If the destination is grouping a number of input signals into the mixer's Wave In destination, it is usually called Record Control. If the destination is grouping a number of sources to an external output, it is usually called Play Control. The precise names displayed by the mixer application for Record Control and Play Control are decided by the programmer who wrote the sound-card driver. It is typical to have several input signal source lines attached to more than one destination. For example, Fig 1 shows Line In appearing on both Record Control (the Speakers destination) and on Play Control (the Wave-Input) destination.

Each source line and destination line has one or more *controls* attached to it (not shown in Fig 1 to improve clarity). A control may be a volume slider, a peak meter, a mute check box and so on. Controls are the items you need to access to affect their state or read/change their values.

I found the mixer particularly confusing at the start and was erroneously relating sources with *input signals* and destinations with *output signals*. Avoid making the same mistake. Study the construction of your own mixer until you can draw its schematic from memory. You must also notice the differences between the Microsoft-defined *component types* that are standard in all sound-card-driver software and the actual names assigned to the various components by the programmer who wrote the driver. One particular manufacturer's mixer had me stumped until I noticed that although the line was *named* Line In, it was actually *defined* as an auxiliary input component type—little wonder it was difficult to find!

In a mixer, audio output signals go to a destination component type of Speakers, which is typically shown onscreen as Play Control. Audio input signals normally go to two destinations: a component type of Wave In, which is shown as Record Control, as well as to the Speakers destination previously mentioned. The Wave In destination can be used to mix various

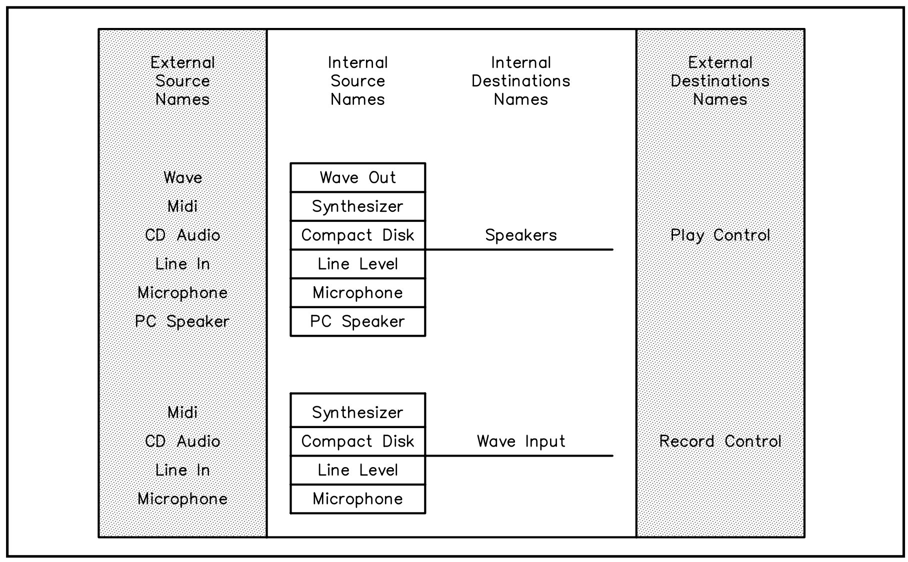


Fig 1—The naming and grouping of a set of mixer controls for an AWE64G sound card. Internal names are only 'visible' by the program; external names are shown on the mixer application screens.

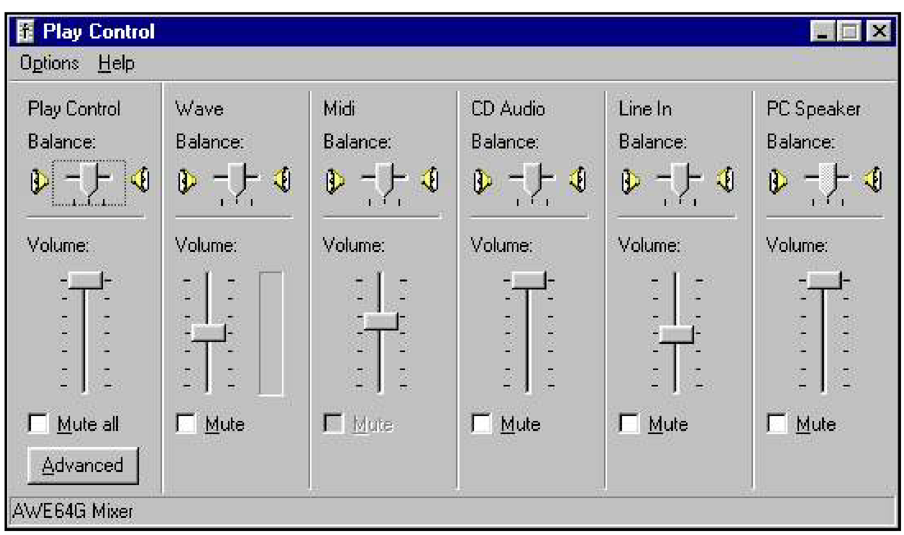


Fig 2—The first screen you see when you start the Windows Mixer application. Usually this is the set of lines and controls for the Speakers destination. Note that the first item called Play Control is actually the destination line itself and also has controls associated with it.

inputs for the purposes of signal analysis or recording .WAV files.

When you start the Windows Mixer application, the first displayed screen is usually called Play Control (see Fig 2). This is the internal destination Speakers. Each item shown on the screen, except for the first (at left), represents a source line attached to that destination. The first item is for the destination itself; remember that destinations can also have controls. In this mixer, the destination's own controls are: a volume slider; **Mute All** check box; **Stereo Balance** and more controls accessed by clicking the **Advanced** button.

If you click **Options|Properties |Recording** and click **OK**, you will see the windows title change to Record Control (see Fig 3) and a different set of lines and controls are displayed. Again, the first item is for the destination, in this instance Wave Input. The other items are for the attached sources.

In case your own mixer looks different from Fig 2 and Fig 3, remember that the actual names on both of these screens are chosen by the programmer who wrote the sound card's device driver. Those shown are for a Creative Labs SoundBlaster AWE64 Gold. Unfortunately, not all programmers have the same view as to which controls should be attached to the various lines and, as already mentioned they have different opinions on the precise internal component type to use!

The *Mixapp.exe* Utility

Another useful utility from MSDN is *Mixapp.exe*. Fig 4 shows the initial screen for the previously mentioned sound card.

This shows that mixer device AWE64G has two destinations: DST Speakers and DST Wave Input. The sources, SRC, are each listed underneath the destinations to which they are attached. The internal Microsoft component type for each line is listed under Component; refer to mmsystem.h to see what those are. The internal Line ID and flag settings are shown together with each line's unique ID within the mixer. Ctrls is the number of controls attached to each device, and Conns is the number of connections to each destination—six in the case of Speakers. The Name column shows the name that the programmer assigned to each of the lines. This one screen shows you most of what you need to know about your mixer.

Fig 5 shows the two controls attached to destination Wave Input. Fig 6 shows further details of the Peak Meter control attached to the same destination.

Mixapp was used many times when debugging the program. The ability to see the structure of the mixer, the internal and external names, the control IDs and the current settings of the controls was an enormous help.

Mixer and Wave-In API Functions and Data Structures

To get the sound card to do what you want, you call various Windows multimedia routines that perform the lower-level work on your behalf. They

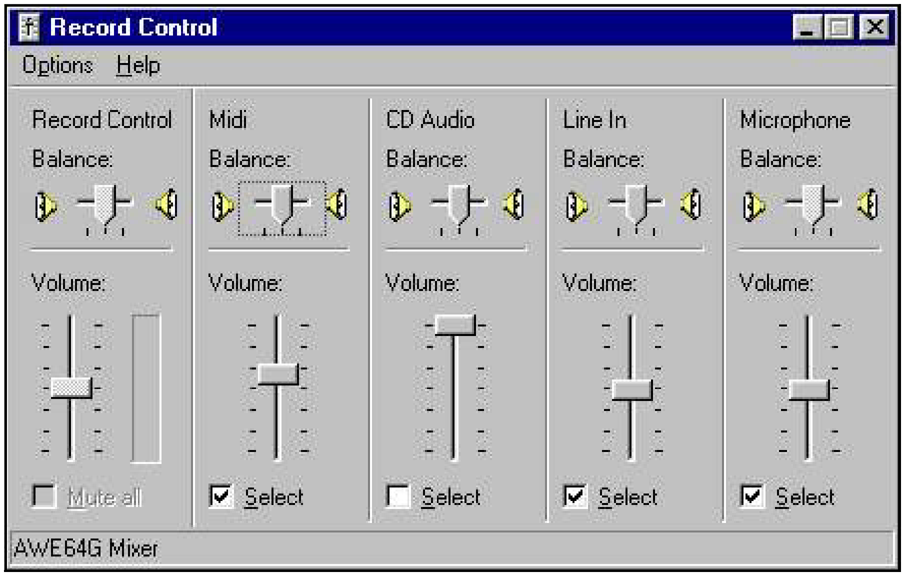
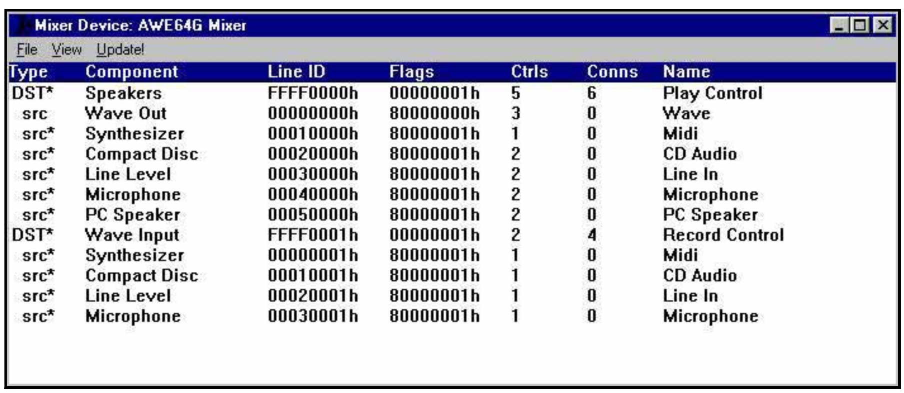


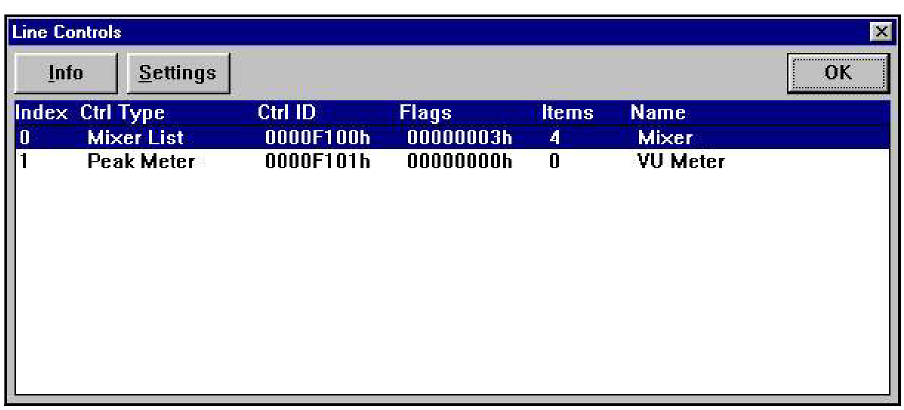
Fig 3—The screen you see after choosing 'options', 'properties', 'recording' and 'OK'. This is the set of lines and controls for the Wave In destination. Some mixers do not allow more than one source line to be enabled at a time!


Mixer Device: AWE64G Mixer

File View Update!

Type	Component	Line ID	Flags	Ctrls	Conns	Name
DST*	Speakers	FFFF0000h	00000001h	5	6	Play Control
src	Wave Out	00000000h	80000000h	3	0	Wave
src*	Synthesizer	00010000h	80000001h	1	0	Midi
src*	Compact Disc	00020000h	80000001h	2	0	CD Audio
src*	Line Level	00030000h	80000001h	2	0	Line In
src*	Microphone	00040000h	80000001h	2	0	Microphone
src*	PC Speaker	00050000h	80000001h	2	0	PC Speaker
DST*	Wave Input	FFFF0001h	00000001h	2	4	Record Control
src*	Synthesizer	00000001h	80000001h	1	0	Midi
src*	Compact Disc	00010001h	80000001h	1	0	CD Audio
src*	Line Level	00020001h	80000001h	1	0	Line In
src*	Microphone	00030001h	80000001h	1	0	Microphone

Fig 4—The initial screen of the sample C application *Mixapp.exe*. Note the grouping of the various sources (src) to the destinations (DST).


Line Controls

Info Settings OK

Index	Ctrl Type	Ctrl ID	Flags	Items	Name
0	Mixer List	0000F100h	00000003h	4	Mixer
1	Peak Meter	0000F101h	00000000h	0	VU Meter

Fig 5—These are the controls attached to destination Wave Input. This screen is displayed by double-clicking on the DST Wave Input line shown in Fig 4.

Table 1

The Mixer functions control the mixer itself, and the Wave-In functions control waveform audio input devices such as the line-in and microphone inputs. The Data Structures enable you to pass information to and receive information from these interfaces.

Mixer API functions

mixerOpen	Opens a mixer device
mixerClose	Closes a mixer device
mixerGetNumDevs	returns the number of mixers installed
mixerGetID	returns the ID of an individual mixer
mixerGetDevCaps	returns the mixer's capabilities
mixerGetLineInfo	returns information about a line
mixerGetLineControls	returns information about a line control
mixerGetControlDetails	returns information about a control e.g. present peak level
mixerSetControlDetails	changes attributes of a control e.g. set the volume level

Wave In API functions

waveInGetDevCaps	returns the capability of a waveform-input device
waveInGetNumDevs	returns the number of waveform-audio devices present
waveInOpen	opens a wave-in device for recording
waveInClose	closes a wave-in device
waveInPrepareHeader	prepares a buffer for waveform-audio input
waveInUnprepare	cleans up preparation performed by waveInPrepareHeader
waveInStart	starts recording
waveInReset	stops recording and resets the current position to zero
waveInStop	stops recording
waveInAddBuffer	sends an input buffer to a given waveform-audio input device

Data Structures

MIXERCAPS	contains information about the capabilities of the mixer
MIXERLINE	describes the state and metrics of an audio line
MIXERLINECONTROLS	contains information about the controls of an audio line
MIXERCONTROL	contains the state and metrics of a single control
MIXERCONTROLDETAILS	a header containing the control ID and a pointer to further detailed information about the control
WAVEINCAPS	describes the capabilities of a waveform audio input
WAVEHDR	defines the header used to identify a waveform audio buffer
WAVEFORMAT	defines the format of waveform audio data

Table 2

```
Declare Function mixerOpen Lib "winmm.dll" (phmx as Long, _
                    ByVal uMxid As Long, _
                    ByVal dwCallback As Long, _
                    ByVal dwInstance As Long, _
                    ByVal fdwOpen As Long) As Long
```

The code to call the function could be:

```
Dim hmixer As Long              ' mixer handle
Dim rc As Long                  ' return code
Const DEVICEID =0               ' device ID of the 1st mixer

' Open the mixer specified by DEVICEID
rc = mixerOpen(hmixer, DEVICEID, 0, 0, 0)       ' open the first mixer
If rc <> 0 Then                 ' if the open has failed
  MsgBox "Couldn't open the mixer, rc='+ Str(rc)
End If
```

Table 3—Approach to search for a particular control in a mixer structure

1. Open the desired mixer.
2. Find out how many destinations are in the mixer.
3. For each destination line:
   A. If the destination contains the wanted control, save the ID of the control for later use and stop searching; if not:
   B. Find out how many source lines are on the destination.
4. For each source line:
   A. If the source line contains the wanted control, save the ID of the control for later use and stop searching; if not:
   B. Look at the next source line.
5. We have come to the end of the source lines without success.
6. Look at the next destination if there are more to do.
7. We have come to the end of the destinations without success.

Table 4

```
Function GetVolumeControl(ByVal hmixer As Long, _
  ByVal componentType As Long, _
  ByVal ctrlType As Long, _
  ByRef mxc As MIXERCONTROL) As Boolean

′ This function attempts to obtain a mixer control to enable user control
′ of the input volume.
′ First we find wave-in, then use mixerGetLineInto to index up each of
′ the source lines until we find a match for the requested line - either
′ linein or mic. Then we use mixerGetLineControls to get the required attributes of the
′ volume control

  Dim mxlc As MIXERLINECONTROLS    ′ mixer line control structure
  Dim mxl As MIXERLINE             ′ mixer line structure
  Dim hMem As Long                 ′ memory address
  Dim rc As Long                   ′ return code
  Dim WaveInConnections As Long    ′ number of connections to destination wave-in
  Dim I As Long                    ′ index to sources on wave-in destination
  Dim NameOfControl As String      ′ name of control being surveyed

  mxl.cbStruct = Len(mxl)          ′ initialise length of mixer line structure
  ′ first, find the wave-in destination
  mxl.dwComponentType = MIXERLINE_COMPONENTTYPE_DST_WAVEIN
  rc = mixerGetLineInfo(hmixer, mxl, MIXER_GETLINEINFOF_COMPONENTTYPE)
  If (MMSYSERR_NOERROR <> rc) Then
    GetVolumeControl = False       ′ failed to get wave-in destination
    Exit Function
  End If

  ′ now save number of connections on this destination
  WaveInConnections = mxl.cConnections

  ′ loop through sources on wave-in to find a matching component type
  For I = 0 To WaveInConnections - 1
    mxl.dwSource = I               ′ set the source identifier
    rc = mixerGetLineInfo(hmixer, mxl, MIXER_GETLINEINFOF_SOURCE)  ′ get source
    If (MMSYSERR_NOERROR = rc) Then       ′ if we got a source line
      ′ first, look for a component type match
      If mxl.dwComponentType = componentType Then
          Exit For                 ′ found a match
      End If
      ′ however, if we are looking for line-in see if there is a name match
      ′ because some mixers use a component type of auxiliary for line in
      If componentType = MIXERLINE_COMPONENTTYPE_SRC_LINE Then
        StringCopy mxl.szName, NameOfControl  ′ get the control name
        If InStr(UCase(NameOfControl), "LINE") <> 0 Then
          Exit For             ′ found line in
        End If
      End If
    End If
Next
```

are shown in Table 1. Thankfully, there are comparatively few API (Application Program Interface) functions to understand, though some of them have multiple functions.

How the Functions are Defined and Used

Each routine that you wish to use must be declared. For example, code to open the mixer using the multimedia DLL winmm.dll appears in Table 2.

If all this looks rather daunting, do not be put off. Fortunately, absolutely all of the hard work in declaring the required functions has already been done for you, together with guidance on using them, in the excellent example programs on MSDN. The example containing the most comprehensive set of functions, complete with descriptions of the parameters required is in article Q187673, "AUDIOLVL.EXE - Monitor Input and Output Audio Levels" found at **support.microsoft.com/support/kb/articles/Q187/6/73.asp**

The examples also provide a level of "insulation" from using the calls directly in your own code. "Wrapper" procedures are provided which take care of numerous pieces of initialization required prior to making calls. The problem of *Visual Basic*'s habit of doing things to copies of data structures rather than to the real ones is also resolved.

```
   ' check we have not dropped through and found nothing
   If I = WaveInConnections Then
      GetVolumeControl = False          ' show not found
      Exit Function
   End If

   If (MMSYSERR_NOERROR = rc) Then
      mxlc.cbStruct = Len(mxlc)          ' set length
      mxlc.dwLineID = mxl.dwLineID ' set the lineID
      mxlc.dwControl = ctrlType          ' set the control type wanted
      mxlc.cControls = 1                 ' show only 1
      mxlc.cbmxctrl = Len(mxc)           ' set length

      ' Allocate a buffer for the control
      hMem = GlobalAlloc(&H40, Len(mxc))
      mxlc.pamxctrl = GlobalLock(hMem)
      mxc.cbStruct = Len(mxc)            ' set length

      ' Get the control
      rc = mixerGetLineControls(hmixer, _
      mxlc, _
        MIXER_GETLINECONTROLSF_ONEBYTYPE)

      If (MMSYSERR_NOERROR = rc) Then
          GetVolumeControl = True        ' show found

          ' Copy the control into the destination structure
          CopyStructFromPtr mxc, mxlc.pamxctrl, Len(mxc)
      Else
          GetVolumeControl = False       ' show not found
      End If
      GlobalFree (hMem)                  ' free buffer
      Exit Function
   End If

   GetVolumeControl = False              ' show not found
End Function

' To use of this function within the body of the program proper:
' search for a volume control on line-in
   ok = GetVolumeControl(hmixer, _
      MIXERLINE_COMPONENTTYPE_SRC_LINE, _
      MIXERCONTROL_CONTROLTYPE_VOLUME, _
      micCtrl)
   If (ok = True) Then                   ' if we found line-in then........
```

Some Practical Examples

Because it required a good understanding of the mixer, the task found most satisfying to get working successfully was to find a way to traverse its structure when searching for a particular control. The actual code is too long to be included here but the general approach eventually used appears in Table 3.

If at the end of a search as described above, you have found the control you are looking for, you then use mixer GetControlDetails or mixer SetControlDetails to read or write values as appropriate.

The particular use of the above process was to scan the whole of the mixer structure, find every peak-meter control and display them all. The user could then choose which one he wanted to use for collecting readings of audio signal level.

Another situation, this time illustrated with part of the real code used (Table 4), required finding the volume control for either the line-in or the microphone on the Wave In destination. See if you can spot the rather crude but effective method used to find Line In defined as an Auxiliary. Lines starting with an apostrophe are comments.

What Can You Do With The Controls?

So, what sorts of things can you do with the various controls? That depends on the control type. There are numerous types of controls: on/off switches, peak meters, mute indicators, mono and stereo indicators, faders, volume controls, bass and treble controls and so on.

Once you have opened the mixer, traversed the required structure and found the particular control you need, it is a relatively straightforward process to either read the current values using mixer GetControlDetails or to set new values using mixer SetControlDetails. With these two calls, you can read or set volume levels using the volume slider, read the current level of a line using the peak meter (if one is attached to that line), enable or disable various lines that have mute switches and so on.

Conclusion

Accessing and controlling a sound card is not as difficult as you might have thought. The hardest part is in understanding the structure of the mixer. I hope that this description is sufficiently clear to encourage you. Dig a little deeper and give it a go!

It is essential that you get some code that already works to help you understand how to make things happen. To this end, the MSDN sample programs are great because they get you up and running in a matter of hours rather than weeks or months.

The most important tip for those writing software to access sound cards is: From the outset, cater to the variability in construction of a mixer. There are many different implementations to consider when writing software that will work on a broad range of different manufacturers' equipment and drivers. You cannot simply look at your own mixer construction and presume that others will be built the same way.

On the other hand, if your only interest is in doing something for yourself, things are much easier. Just look at your own mixer setup and have a lot of fun!

Fig 7 shows the current look of the project and illustrates what you can do by reading audio levels. This is a screen shot of the polar diagram for a 10-element 2-m Yagi designed and built by G2HCG. If you would like to know more about the program, point your browser to **www.bob.freeth.dial.pipex.com/polarplot** and read the current user guide.

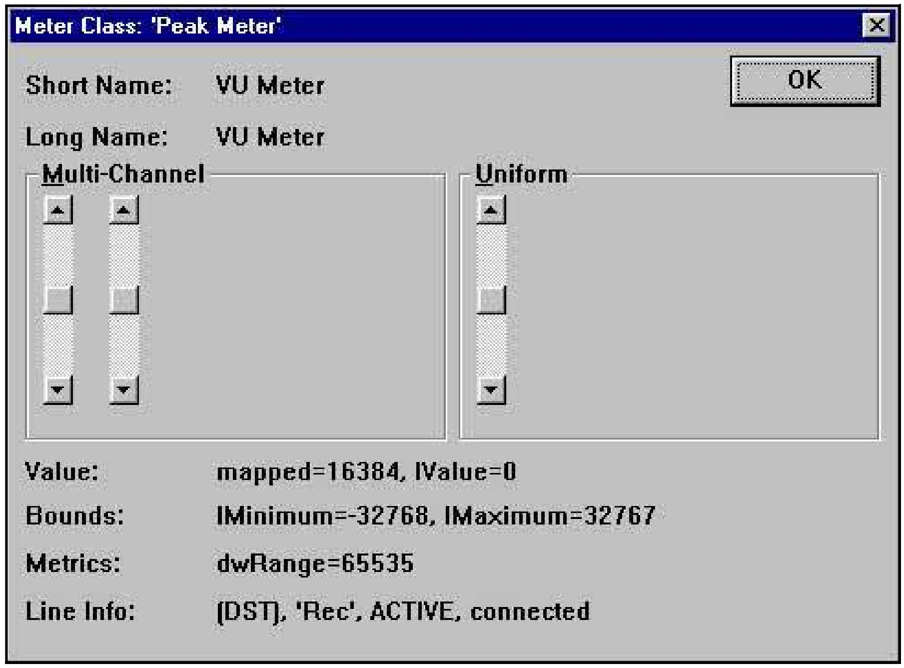


Fig 6—Details of the Peak Meter. This screen is displayed by double-clicking on the Peak Meter line shown in Fig 5. Note that this peak meter can return negative as well as positive numbers in the range –32768 to 32767.

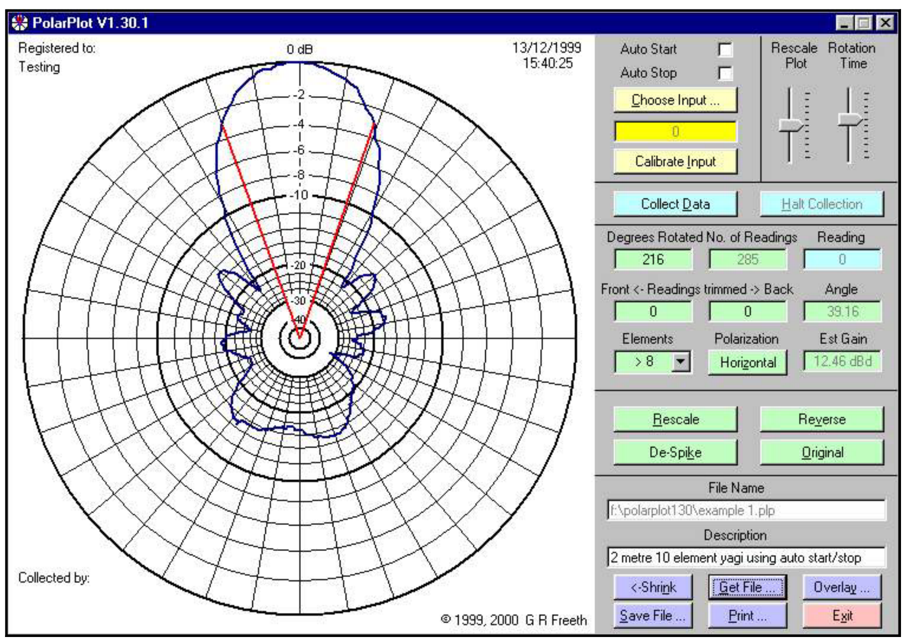


Fig 7—A screenshot of the finished project showing the polar diagram of a 2 meter 10 element Yagi designed and built by G2HCG.

Bob Freeth was born in 1943, the son of an Army bandmaster. Before age 12, he lived in England, Malta, Germany, Malaya and Borneo. Next, he lived five years at boarding school in Sussex, England. He spent the first five years of working life as an articled clerk with a firm of chartered accountants in the city. He left accountancy in 1966 and took up a computing career working for a manufacturing company on IBM mainframes in London. For the next 10 years, he specialized in the Systems Programming field. That involved installing, maintaining and modifying operating systems, programming mainly in assembler.

He got married and moved to his present location on south coast of England, and spent 15 years working for an insurance company in the computer-services division. His primary roles were the selection, implementation and management of data-center hardware/software resources and the provision of consulting services within the company. For four of these years, he managed the data center serving 2000 users at head office and 70 sales branches.

He then left the IBM mainframe environment and became an independent PC consultant. After nine years, Bob is now sliding into semiretirement. He enjoys getting his hands "dirty" again after starting development of PC software for the Amateur Radio market.

Laying Out Azimuth Lines

Which way is north? GPS alone can't tell you. Come tour the worlds of geology, navigational astronomy, geometry and surveying, as we search for true *north!*

By Walter Schulz, VQ9TD/K3OQF

Recently, I retired and moved to a rural location in northeastern Pennsylvania. My new ham-radio site consists of three acres of wooded land on Scrub Mountain. The mountaintop is approximately 1300 feet above sea level. I decided to install three types of directional antennas: a two-wire bi-directional Beverage system, a three-band Yagi on a 70-foot tower and a six-foot dish pointed at Intelsat K (over the mid-Atlantic) to receive Deutsche Welle television signals.

Radio-site layout requires accurate aiming of directional antenna systems. This means finding true north and the site's exact geographical position. This article discusses how to accomplish these installation tasks.

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Finding True North: with a Compass

Magnetic Declination

One would think that aiming antennas at a particular DX target area—using true north as a reference—should not be too difficult. At first, I thought a magnetic pocket compass would be sufficient; however, I found a pocket compass alone was inadequate. It is far superior to employ an open-sighting compass (surveyor's compass) that is sturdily mounted on a level tripod. Even with this arrangement, though, compass-reading errors can occur that may involve several minutes of arc—which is significant. In addition, this method is obsolete.

Magnetic determination of true north is complicated by another factor: magnetic declination. Compass needles point to the north *magnetic* pole, located at about 76° N 101° W, which is in northern Canada and far from the true geographic North Pole.[1] *Magnetic declination* must be considered to correct magnetic readings to true north.[2]

To find your magnetic declination, it is first necessary to have your exact geographical position. GPS units cost about $100. They are undoubtedly the most efficient way of obtaining position information. I employ a Trimble Ensign model.[3] The Ensign gave me my coordinates as 40° 56' 33" N and 75° 39' 38" W. I used two programs, found on the Internet, to compute my

[1]Notes appear on page 30.

magnetic declination. The first program may be found at **www .geolab .nrcan.gc.ca/geomag/e_ cgrf.html**, which is maintained by the Canadian government. The second is at **www .ngdc.noaa.gov/cgi-bin/seg /gmag/ fldsnth1.pl**, and it is maintained by NOAA and the National Geophysical Data Center (NGDC). This latter program is perhaps too verbose for most amateurs and I recommend the first site's program. In addition, freeware for magnetic-field calculations is available from **geomag.usgs.gov/ Freeware/geomagix.htm**. The National Geomagnetic Information Center, US Geological Survey maintains a GEOMAG site at **geomag.usgs.gov/**.

Magnetic declination may also be found from US government quadrangle (7.5') maps that cover your area. These maps are available from the US Government Printing Office and may be found locally, in many instances. Look for them at your county surveying/engineering offices or at one of the federal government's

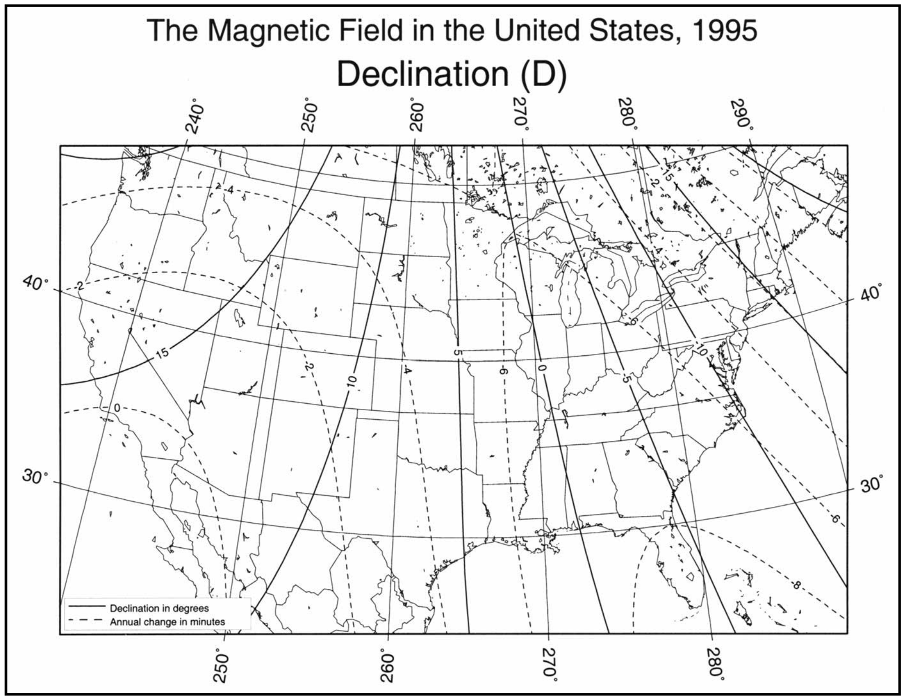

Fig 1—Magnetic declination in the United States, 1990. (Courtesy of National Geomagnetic Information Center.)

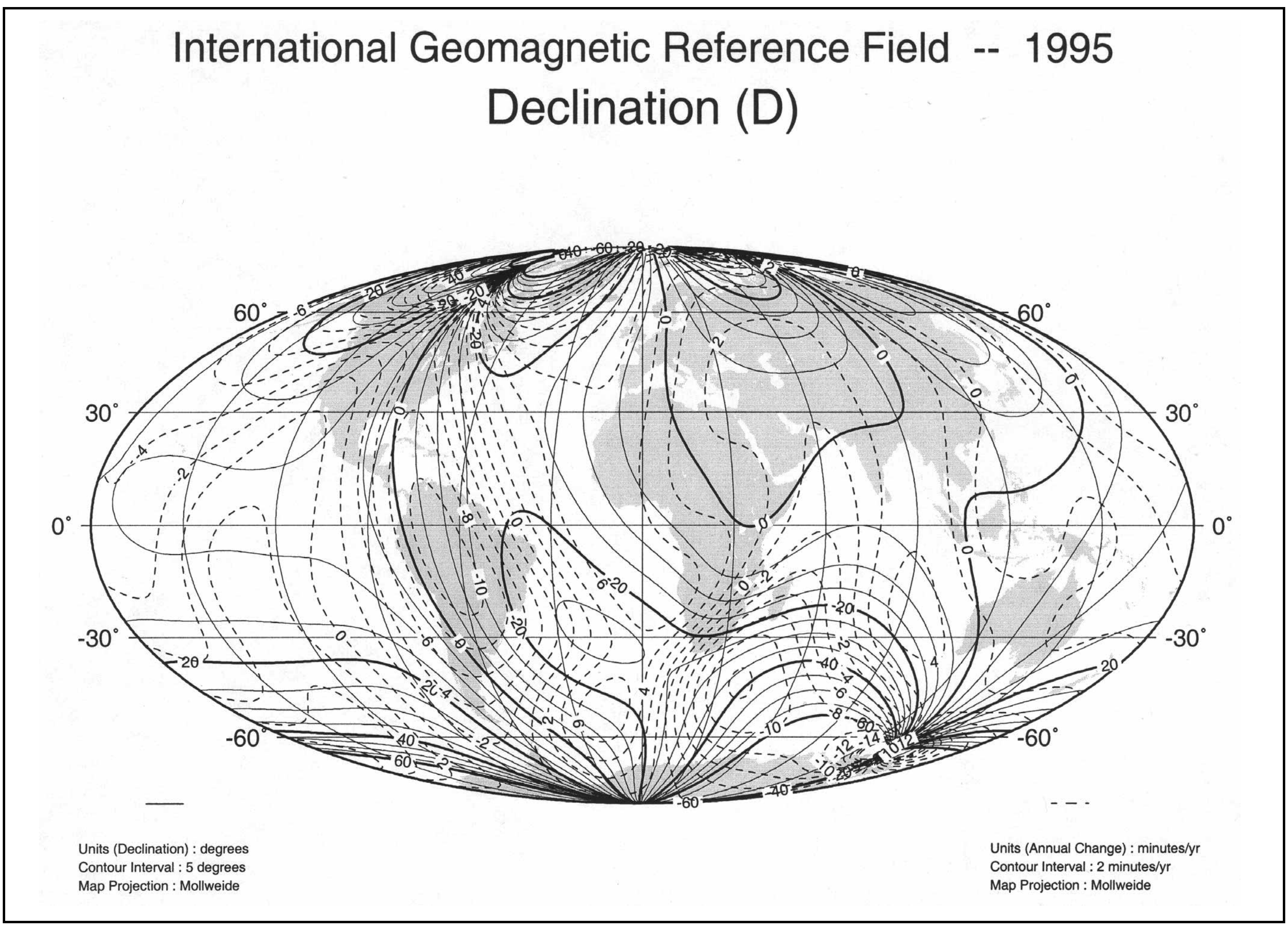

Fig 2—International Geomagnetic Reference Field, 1995 Declination (D). (Courtesy of National Geomagnetic Information Center.)

authorized sales agents. Magnetic-declination values are printed on each map.

You can get a rough estimate of your magnetic declination from an *isogonic* chart of the US, such as that shown in Fig 1. Fig 2 shows an isogonic chart of the entire world.

Using the Canadian program, I found my magnetic declination to be 11° 52' W. That means my magnetic bearing to true north is less than 360° by the 12° declination, or about 348°.

Magnetic Variation

Variation is another type of compass error. Variations may be secular, daily or irregular, along with miscellaneous alterations in declination. Variation can be caused by nearby ferrous objects, such as automobiles, railroad tracks, gas pipelines, culverts, steel radio towers and wire fences.

Armed with the above information, I set out to find true north, but this proved an arduous task at my location. In fact, readings seemed to contain a consistent error. The error was probably due to some flaw in the compass or a large iron deposit nearby. The source did not matter; I could not rely on an accurate compass reading! Therefore, I decided to use astronomical observations and a GPS to obtain true azimuth angles.

Geographical and Geocentric Coordinates

The Earth is not really a smooth sphere, but an oblate spheroid.[4, 5, 6, 7] It bulges at the equator with the north and south poles slightly flattened. The equatorial diameter is approximately 7939 miles, while the polar diameter is about 7900 miles (see Fig 3). The difference influences some very important calculations. That is because the geographical position defined by

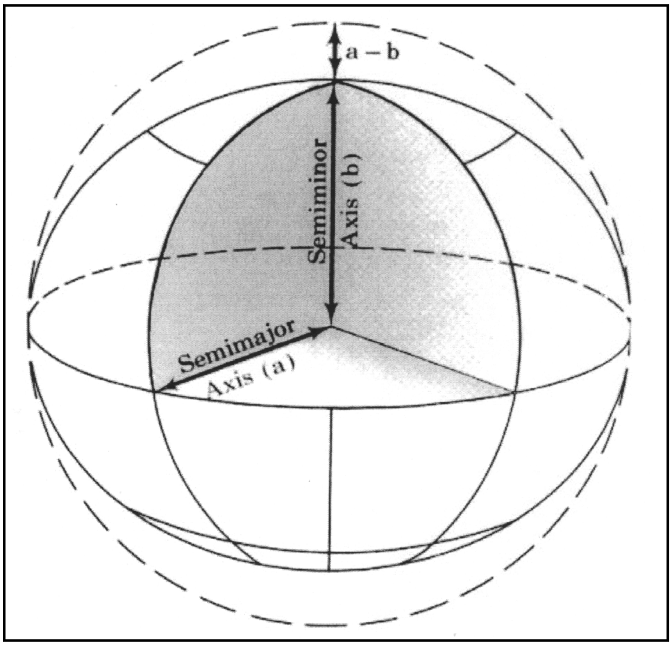

Fig 3—Oblate spheroid or ellipsoid of revolution (from Bowditch).

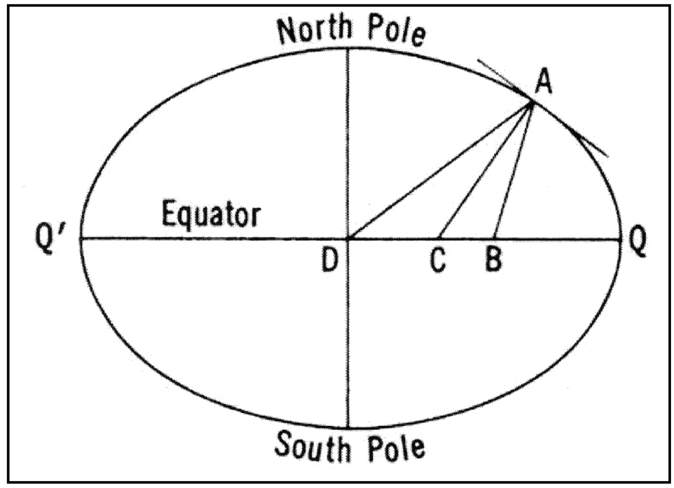

Fig 4—Three kinds of latitude at point A (from Bowditch).

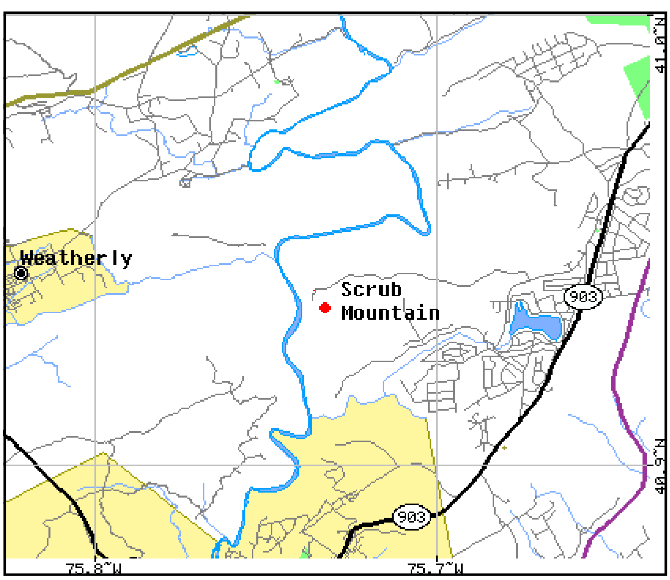

Fig 5—Tiger map of Scrub Mountain vicinity, near Jim Thorpe, Pennsylvania. (Source: USGS GNIS Map Server.)

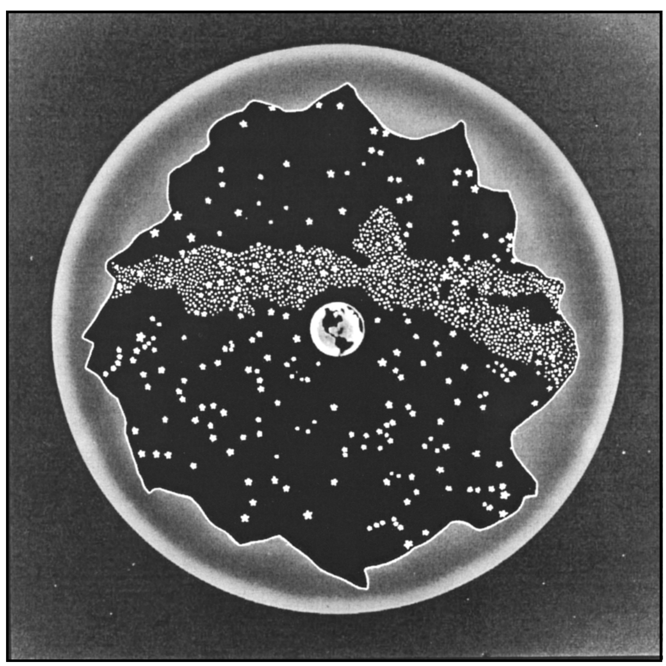

Fig 6—The earth at the center of the celestial sphere (from Bowditch).

latitude does not necessarily coincide with geocentric latitude.

Geographic latitude is the angle formed by the equatorial plane and an imaginary plumb line placed at some position on the Earth's surface. See Fig 4. Geocentric latitude can be defined as the angle between the equatorial plane and a line from the Earth's center to a position on its surface. Again, the plumb line is placed at right angles to the geoidal horizon. There are only two places on the Earth's surface where the imaginary plumb line will coincide with the geocentric and geographic latitudes: the north and south poles.

The difference in distance between the geoidal and celestial horizons is not constant but varies with geographical latitude. The ellipsoidal shape of the Earth must be accounted for during astronomical calculations, especially when calculating the direction of a high-gain parabolic dish to point at a geostationary satellite. Ignoring the ellipsoidal shape results in parallax error. Although the errors are small, purists would say they are significant. From my own point of view and experience, they should be considered for best results, but it is not necessary.

Planning a Site Layout

Begin a site layout with your geographical position. A GPS is the easiest way to obtain those data. Take a reading from the proposed location of your antenna. This location will become the starting point for all future calculations. Your property deed will show survey boundaries and property corner markers. First, go out and find the markers. Then determine the distance from each of the markers to the antenna location. The GPS may again be used for this.

Another way to get good position information is to bring up the US Census Bureau's page at **www.census.gov/cgi-bin/gazetteer**. A gazetteer may be used to identify your location on the Tiger Map Server. Next, select the Tiger Map of your area (see Fig 5). Note that downloading time from the Tiger Map server can be lengthy, depending on your connection speed.

You can also get good position information from quadrangle maps by interpolation. Mr. Charles Elam, a well-known land surveyor, sells the Geo-Ruler, for this very purpose. The Geo-Ruler[8] is available from Mr. Elam's surveying company.

Celestial Coordinates, Ephemeris and the Navigational Triangle

To understand how accurate azimuth lines are obtained, and thus antenna elements placed in their proper orientation, some knowledge of celestial navigation and the *navigational triangle*[9, 10, 11] is necessary. In particular, navigational astronomy is helpful in illustrating fundamental concepts that make GPS possible. It is not my intent to provide an in-depth study of spherical trigonometry, which is certainly beyond the scope of this article, but to give a brief overview that lets you grasp the concepts. Should you want to know more, consult the references at the end of this article.

From early human history, mankind has used the sun and stars to discover something about the nature of his surroundings. We found, through the study of celestial bodies, that we could tell time and make measurements that provide us with knowledge of the Earth and heavens. From our position, it looks as if we're at the center of the universe and objects in the sky seem to rotate around us. Today we know that is not true, but it is a very simple concept that is easy to understand (see Fig 6).

Latitude is relatively easy to determine from astronomy because the point in the sky about which celestial objects seem to rotate is either true north or true south. Latitude is defined with respect to the rotation of the Earth alone. Longitude is defined with respect to the positions of the Sun and other heavenly bodies in the context of time. Imagine a line drawn from the Sun to the Earth's surface. That point on the Earth is called the subsolar point or the geographical position of the sun. As the Earth rotates, the Sun's geographical position moves westward. The questions are how do we define the Sun's location on the *celestial* sphere, and how does this relate to geographical positioning?

To start, we assign rectilinear coordinates to the map of the celestial sphere, just as we do to the surface of the Earth. That coordinate known as latitude on Earth is projected upward

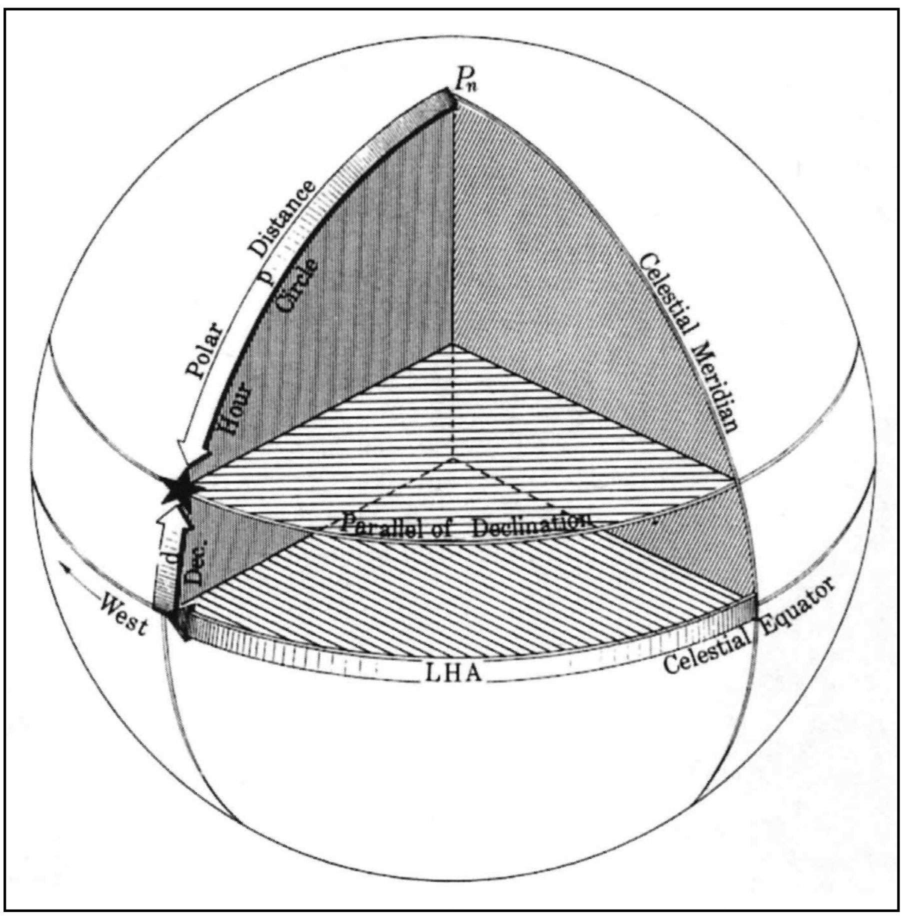

Fig 7—The celestial system of coordinates, showing measurements of declination, polar distance and local hour angle (from Bowditch).

from the surface and is known as *declination* in celestial coordinates. The equivalent of longitude here on Earth is known as *right ascension* in celestial coordinates. Therefore, the map of the celestial sphere is just like a globe, except that we are looking at the globe from the inside rather than the outside (see Fig 7).

Both the US Naval Observatory and the Royal Observatory (UK) publish an ephemeris and nautical almanac every year. The almanacs contain tables of the declination and right ascension of the Sun (and other astronomical objects) at specific times. Therefore, when we observe the Sun at a specific time, we may determine its sub-solar position on the Earth's surface. We know its declination directly corresponds to latitude and that its right ascension is somehow related—by time—to longitude. Fortunately, the almanacs also publish lists of the Greenwich Hour Angle (GHA) for their predictions. The GHA is measured in time, westward from the Royal Observatory in Greenwich (pronounced "gren-itch"), whose longitude is defined to be 0° or 360°. With an accurate timepiece and a measurement of the Sun's position, we can directly find our longitude.

When we know our geographical position and the Sun's geographical position, a great-circle arc between the two points may be drawn. The length of this arc is the distance between the two points. If we add a point at the North Pole, a spherical triangle may be drawn connecting the three points. This is the well-known navigational triangle (see Fig 8). It is intuitively obvious that when we know the Sun's azimuth angle, we can work backward to find true north! With the advent of computer algorithms, one can calculate an ephemeris and the Sun's azimuth for any given time. Two programs are available to the public at reasonable cost: (1) *Interactive Computer Ephemeris*[12] (*ICE*) and (2) *Multiyear Interactive Computer Almanac 1990-2005*, on CD-ROM.[13] Both were developed at the US Naval Observatory. I recommend the *Interactive Computer Ephemeris*. It also calculates great-circle headings (azimuths) toward the geographical positions of a number of celestial bodies, including the Sun. It will do so for any time of observation. ICE is available as freeware at **www.seds.org/billa/ice/ice.html**. It is offered in a .ZIP file for the years 1902-2049. Unfortunately, the software is no longer supported by the US Naval Observatory.

Professor Roelofs and the Local Hour Angle Method

In WW2, many technological advancements were made in applied sciences. WW2 prompted new methods of rebuilding a Europe in ruins. Land surveyors employed astronomical methods to determine control points and reference azimuths for surveys. Dr R. Roelofs of the University of Delft, Holland, wrote a book[14] describing these methods. The Roelofs Local Hour Angle method (LHA, see Fig 8) created a solution to finding and fixing directions on a long leg of a reservoir boundary survey. He used the LHA method 500 times to fix and control azimuths in the survey. Later, as Dr Roelofs' book became more widely read in the US, Drs Elgin, Knowles and Senne[15] in Rolla, Missouri, and Dr R. B. Buckner[16] at Ohio State University wrote further about the LHA method in their own books. These men improved on Dr. Roelofs' methods and made them popular among surveyors.

To make an accurate measurement of the position of the Sun in the sky, one must use a telescope that is accurately aligned with respect to the Earth's surface. That is, it must be level and have means of precisely reading the angle in which it is pointed. For surveying work, such an instrument is called a *theodolite*.

Theodolites

The origins of the theodolite date to 16th century English mathematician Leonard Digges. He developed such an instrument to measure horizontal and vertical angles. Today, it has become the basic surveying instrument. It consists of a telescope with a cross-hair reticule, mounted on a swivel

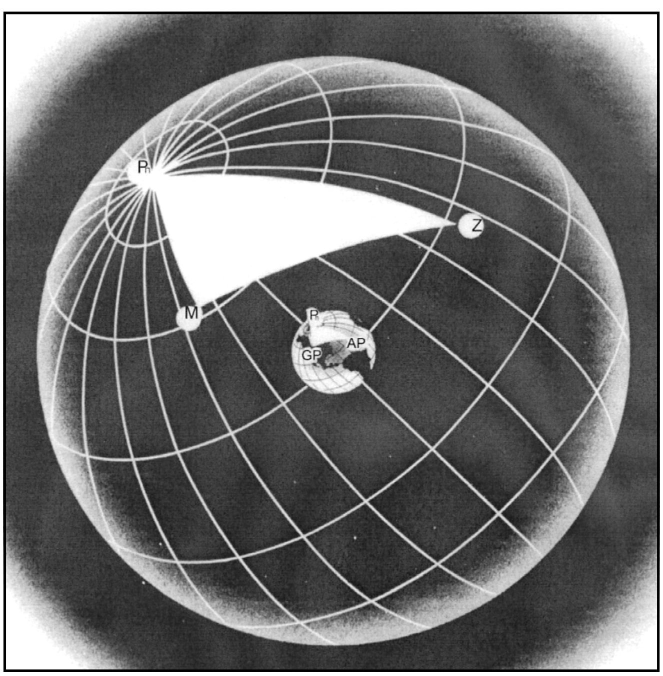

Fig 8—The navigational triangle in perspective, the earth at its center inside the celestial sphere (from Bowditch).

Fig 9—A DT-100-series digital theodolite. (Source: Topcon Corporation of America.)

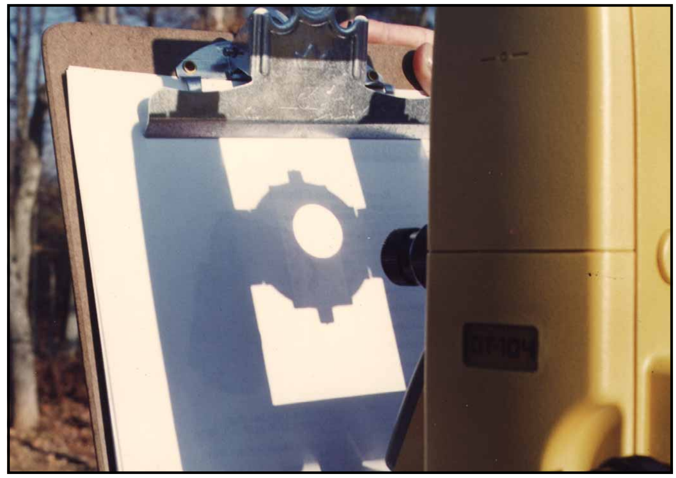

Fig 10—Joe Berens, W3BYX uses a theodolite to project the sun's image on a sheet of paper.

atop a tripod. A theodolite is normally equipped with two scales—altitude and azimuth—that are used to read angles. The usual technique begins with leveling the instrument by means of spirit levels in the horizontal plane.

A modern theodolite is the Topcon model DT-104, an electronic type that I use to make azimuth measurements (see Fig 9). Fig10 shows a theodolite on its tripod.

The importance of properly leveling a theodolite cannot be overstated. To obtain precise azimuths, the instrument must be perpendicular to the force of gravity at one's location, not necessarily parallel to the Earth's surface there. Charles Elam conducted a number of lectures on the LHA method. At one of them,[17] he described a procedure that uses a fictitious star to measure the error in the leveling of a theodolite.

When sighting on the Sun with a theodolite, observations at elevation angles above 45° are not very useful. That means that our measurements of the Sun are going to be made in the early morning or late evening hours. To safely view the Sun with a telescope requires a great deal of care.

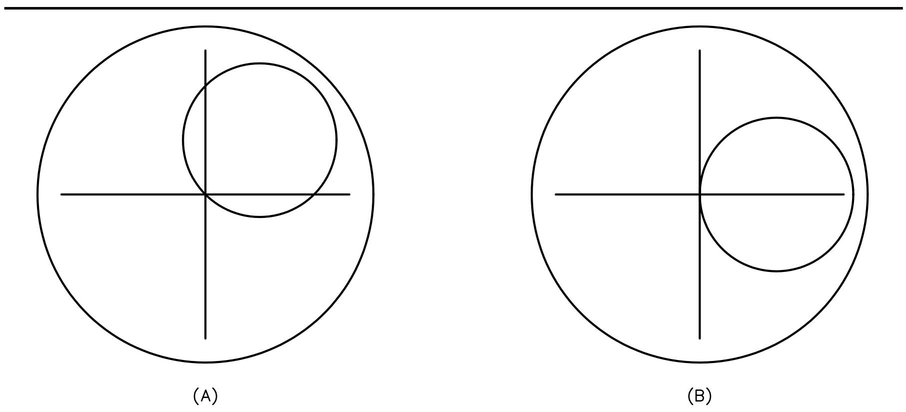

Fig 11—View of the theodolite telescope. At (A) the sun shot is beginning; the sun is moving, left to right, across the vertical wire. At (B) the sun's left limb is coincident with the wire. Mark the time at this instant for the sun's GHA and declination.

Viewing the Sun

The radiation coming from our Sun contains a broad spectrum of energy. It has been found that its ultraviolet radiation is most harmful to the human eye. Therefore, *never look at the Sun without the protection of a solar filter*. The retina of the human eye is *very* light sensitive. Direct viewing of the Sun without protection may result in retinal burns and *permanent blindness*. Heat absorption is the main cause and the destruction takes place without pain; the actual visual degradation may not occur for several hours after exposure.[18]

One excellent way to avoid viewing the Sun directly is to project its image onto a piece of paper (see Fig10). This is a nice way to view the Sun, but it is not very useful for finding angles. Heat from sunlight can degrade the accuracy of a theodolite and even ruin its components. A solar filter is necessary at the objective. This is the only way to obtain the sharpest reading. I use a Thousand Oaks Type 2+ solar filter.[19, 20] The filter is glass with a special metal coating on one side. The plus in the part number indicates the addition of a steel alloy to the metal coating to improve its durability. The filter shows the Sun as a yellow-orange light against a black sky. This is perfect for measuring angles using the Sun's limb. Make sure the filter is fitted securely to the telescope, lest it fall off during observation with disastrous results!

Solar observations are best made in early morning or late evening when the Sun's elevation is low (see Fig 11). The theodolite telescope is set up so that the Sun's path through the sky will take it across the vertical crosshair in the eyepiece. The instrument remains stationary during measurement and *does not* *track the Sun in the sky*. First, the observation is taken when the trailing limb of the Sun transits the vertical crosshair. At that instant, the time is recorded. Next, an ephemeris is used to determine the Sun's geographical position at the exact time of the measurement. That is found through interpolation of ephemeris data given as either Greenwich Hour Angle or right ascension and declination.

As I said before, one point in a navigational triangle is the observer's position. The second point is on a great circle from the observer's position toward the Sun, and the third is at the pole. We can solve this triangle to determine the exact azimuth of the theodolite during our measurement above, and thus we find the offset to true north.

Time

An accurate measurement calls for accuracy in the exact time at which it was made. During the 1960s, major changes in how we measure time took place. Atomic clocks, based on fine transitions between energy states of subatomic particles, were found extremely accurate. These new standards were based on properties of certain atoms such as cesium, hydrogen, mercury and rubidium. They are so accurate that we no longer rely on the Earth's rotation to keep time.

Before these clocks were invented, though, time measurement was accomplished by astronomical observation. Since 1964, universal coordinated time or UTC has been regulated by atomic clocks. Irregularities in the orbit and rotation of the Earth are compensated by a correction factor called DUT1. Therefore, UT1, the time used in azimuth calculations, is equal to UTC+DUT1. As of January 1, 1972, continuous correction of the UTC rate is no longer made for slowing of the rotation of the Earth. Instead, leap seconds are added or subtracted when necessary at the end of a UTC month, preferably June 30 or December 31. The last minute of the month in which the correction is made therefore has either 59 or 61 seconds!

Listening to WWV or WWVH, one can hear the DUT1 correction superimposed on the other modulation. The DUT1 correction is encoded using double ticks during the first 15 seconds of each minute. Doubling of the first through seventh ticks indicates a positive correction, while doubling of the ninth through fifteenth ticks indicates a negative correction. The eighth-tick is never doubled. For example, were the first, second and third pulses doubled, you would infer a correction of "plus 0.3 s." Leap-second adjustments ensure that the DUT1 correction cannot exceed 0.7 s.

Reducing the Data

Remember that the solar diameter in the sky is about 32' of arc. Its semidiameter of 16' must be accounted for in calculations to find the exact center of the Sun and thus, the precise solution for its azimuth angle. Mssrs Elgin, Knowles and Senne in their articles and books recommend calculating the actual semidiameter, which changes with the seasons and the elevation of the Sun above the horizon. The latter effect is because of differential refraction in the Earth's atmosphere of the Sun's image. I recommend using the *ICE* program and taking the Sun's semidiameter from that.

Table 1 illustrates this LHA method using an example.[21, 22] LHA is the difference in longitude between the Sun's geographical position on the surface of the Earth and the position of the observer (see Figs 12 and 13).

Conclusion

The impetus for this article about the LHA method stems from the heavy reliance seen today on computer-aided antenna-design programs. Many amateurs rave about how well these programs design and optimize antennas, and that is good; however, without due

Table 1—Solution of an Example Azimuth Problem Employing the LHA Method

Date: 17 October 1999
Target: The Sun
City/Township: Penn Forest, Jim Thorpe, Carbon County
Time of Observation: 13h 47m 31s UTC
Latitude: 40° 56' 49" N
Latitude in Decimal Degrees: 40.95° N
Longitude: 75° 39' 01" W
Longitude in Decimal Degrees: 75.65° W
Sun's GHA: 30° 31' 24"
Sun's GHA in Decimal: 30.52°
Sun's Declination: 9° 13' 00" S
Sun's Declination in Decimal: 9.22°
LHA, Difference in Longitude: 75.65°–30.52° = 45.13°

Ice Program Output

Celestial Navigation Data for 1999 Oct 17 at 13 47 31 UT
Delta T = 63.5 seconds
For Assumed Location: Longitude W 75° 23.7' (75.6503°), Latitude N 40° 34.1' (40.9469°)

	Almanac Data				*Altitude Corrections*			
Object	*GHA*	*Declination*	*Altitude*	*Azimuth*	*Refraction*	*Semidiameter*	*Parallax in Altitude*	*Sum*
SUN	30° 31.3'	s 9° 13.0'	+25° 17.5'	129.6°	–2.1'	16.1'	0.1'	14.1'
ARIES	232° 30.0'							

thought to site construction, one can rapidly depart from the carefully designed performance. In my opinion, all that optimization work goes for naught without a careful layout. Should one not carefully consider proper site-construction practices, deployment of antenna systems designed for high forward gain and front-to-back ratio may prove futile. Frequently, antenna systems do not perform as expected because of inaccurate placement of elements near the Earth's surface.

It seems to me incredible that so much emphasis is placed on computer simulation and accuracy, while that same accuracy is not demanded in the construction and alignment of arrays. My search of amateur and professional literature found only three pages in one *ARRL Antenna Book*[23] where it is even mentioned. I certainly do not pretend that this article gives all the answers, but it should serve as a springboard for further discussion. I hope those of you with more knowledge than I will add to the information presented here.

Acknowledgments

I am heavily indebted to Mr Charles Elam and Mr Wayne Reed, KA1BBC, for reviewing this article, as well as former *Communications Quarterly* Editor Terry Littlefield for her encouragement to write about the subject. I would also like to thank Knowles & Senne Inc, Surveying Consultants, and Mr Gary Parise of Topcon Corporation for their help in this endeavor. "This article is dedicated to the memory of Mr Norris Sapp, Master Radio Electronics Officer, whose key fell silent in March 2000, his last ship was the *SS Green Bay*."

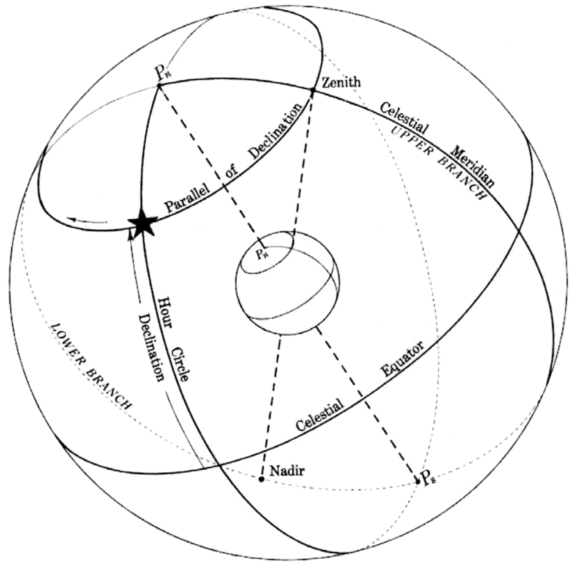

Fig 12—Elements of the celestial sphere. The celestial equator is the primary great circle. A point on the celestial sphere can be located by its declination and hour angle (after Bowditch).

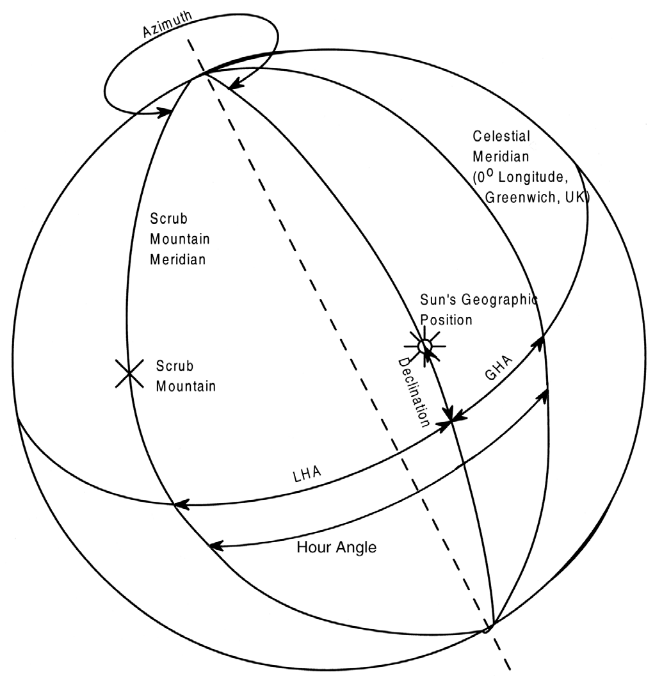

Fig 13—Local Hour Angle defined from the Sun's geographic position over the Earth's surface to Scrub Mountain location.

Notes

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[2]N. Bowditch, *The American Practical Navigator,* Defense Dept, National Imagery and Mapping Agency (NIMA; **pollux.nss.nima.mil/**) pp 81-111, 1995 edition. Hardcopies are available through the US Government Printing Office bookstores (**bookstore.gpo.gov/**), many retail bookstores and marine-supply stores. The entire text is available as PDF files on line at **http://pollux.nss.nima.mil/pubs/pubs_j_apn_sections.html?rid=100**.

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[7]V. G. Szebehely, *Adventures In Celestial Mechanics*, (Austin, Texas: University of Texas Press, 1993) pp 124-130.

[8]*GeoRuler*, Elamco, PO Box 301, Lincolnton, GA 30817; $10.

[9]J. H. Clough-Smith, *An Introduction to Spherical Trigonometry*, (Glasgow, UK: Brown, Son and Ferguson Ltd, Nautical Publishers, 1978) pp 1-17.

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[11]*Nautical Almanac* (Washington, DC: US Naval Observatory; **www.usno.navy.mil/**), available through the Government Printing Office, see Note 2.

[12]Although no longer supported by the US Naval Observatory, *ICE* is in the public domain and available for download from **seds.lpl.arizona.edu/nineplanets/ice/ice.html**. For more information, visit **aa.usno.navy.mil/AA/software/docs/ice.html**.

[13]*Multiyear Interactive Computer Almanac* (MICA), *V 1.5*, US Naval Observatory, Astronomical Applications Department, Washington, DC; **aa.usno.navy.mil/**. MICA is available through Willmann-Bell Inc, see Note 6.

[14]R. Roelofs, *Astronomy Applied to Land Surveying*, (Amsterdam: Uitgeverij Argus, 1950) pp 31-49, 104-106, 116-119 and 211-214.

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[17]C. Elam, Survey Directions: Solar, LHA, HP-41C Lecture Series, 1985; also, personal telephone interview.

[18]R. Chou, "Solar Filter Safety," *Sky & Telescope*, June 1999, pp 36-40, Sky Publishing Corp, Cambridge, Massachusetts.

[19]A. M. MacRobert, "Solar Filters: Which is Best?," *Sky & Telescope*, July 1999. This article is available on the Sky & Telescope Web site **www.skypub.com/resources/testreports/accessories/9907tp_filters.html**.

[20]Thousand Oaks Optical, PO Box 4813, Thousand Oaks, CA 91359; tel 800-996-9111; **www.thousandoaksoptical.com**

[21]P. K. Seidelmann, *Explanatory Statement to the Astronomical Almanac*, (Washington, DC: US Naval Observatory, 1992) pp 39-74. This publication is available from University Science Books, PO Box 605, Herndon, VA 20172; tel 703-661-1572, fax 703-661-1501; e-mail **univscibks@igc.org**; **www.uscibooks.com/seid.htm**.

[22]HP-41C Navigation Pack/Perpetual Almanac, pp 28-29 and 35-36, Hewlett-Packard.

[23]*The ARRL Antenna Book* 13th edition (Newington, Connecticut: ARRL, 1974) pp 323-329. Unfortunately, this information does not appear in the current *ARRL Antenna Book* edition. It has been moved to *The ARRL Operating Manual.* Look for the chapter on "Antenna Orientation." A method first described in *QST*'s "Technical Correspondence" (July 1994, p 83) column does appear in recent *ARRL Handbooks;* look in the index for "North Shadow." In addition, there is a local-noon calculator on the Web at **www.cmpsolv.com/los/sunset.html**.—*Ed)*

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4. *Topcon Digital Theodolite Instruction Manual, DT-100 Series*, Topcon America Corp, 37 W Century Rd, Paramus, NJ 07652; tel 800-223-1130, 201-261-9450, fax 201-387-2710; **www.topcon.com**.

Walter was first licensed in 1960. He is presently active on 20 meter CW and RTTY. After High School he went to Navy Radioman "A" School at San Diego, served at Naval Communication Station Guam (NPN) during 1963, and was one of the operators at the club station, KG6AAY. Later serving on the USS Hancock CVA-19 (NWLD) stationed on Dixie and Yankee Station. Walter studied Electronic Engineering Technology, majoring in Communication Engineering Technology and Aeronautical and Navigation Engineering at Capitol Radio Engineering Institute in Washington, DC. He holds a First Class Radiotelegraph License with sea and radar endorsements, General Radiotelephone Operator License and Amateur Extra Class License

After the Navy, he worked in television microwave-transmission facilities for Bell of Pennsylvania. Walter later worked with large-office information systems at Bell Atlantic Corporation. Retiring from Bell Atlantic, he became involved with Preposition Squad Two stationed at Diego Garcia Atoll in the Indian Ocean. He was a communications officer onboard military chartered ships involved with Desert Shield/Storm, Operation Restore Hope and Operation Vigilant Warrior in the Central Arabian Gulf. Walter retired from the United States Merchant Marine in 1997 and is enjoying the "Switzerland" of Pennsylvania, the town of Jim Thorpe where he resides presently.

On the Nature of the Source of Power in Class-B and -C RF Amplifiers

A popular author wants to set the record straight about conjugate matching.

By Walter Maxwell, W2DU
ARRL Technical Advisor

In this article, we will discuss the nature of the source of power in class-B and -C RF amplifiers. As were my 1970s *QST* articles, "Another Look At Reflections," and my book *Reflections—Transmission Lines and Antennas,* that clarified misconceptions concerning SWR and reflected power, this article is concerned with clarifying misconceptions prevalent among amateurs and professional electrical engineers concerning the operation of RF power amplifiers. In attempting to resolve the unfortunate and protracted controversy concerning the conjugate matching theorem in relation to these amplifiers, discussions with many people revealed an alarming number of misconceptions concerning the complex relationships of voltages and currents that occur in the development of the source of power in these amplifiers, especially in relation to the coupling to their loads. At the core of the controversy are amateurs and engineers alike, who assert that some of the teachings in "Reflections" are fundamentally incorrect. Therefore, it is important that the focus of this article is to highlight and clarify those misconceptions.

Before discussing amplifier operation, however, two synonymous terms that play a vital role in amplifier operation need clarification, because they are widely misinterpreted in discussions relating to the source of power delivered to a load. These terms are *maximum available power* and *all available power*.

Maximum available power, or all available power, is simply the power available for delivery from the source to the load whenever the source is matched to the load. In class-B and -C amplifiers it is the power delivered when the loading is adjusted for peak output *at any given level of drive desired*. It is *not* the absolute maximum power that can be obtained by over driving, or using excess plate voltage or plate current, as many amateurs and engineers alike have been misled to believe.

Turning now to the discussion of amplifier operation, one misconception is that class-C amplifiers cannot support circuit analysis using general network theorems because of the

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nonlinearity[1] of the amplifier operation. In clarifying this misconception we will show that, although the input circuit of the π-network tank circuit in class-C amplifiers is nonlinear, the output circuit to the load is indeed linear, due to energy storage in the tank. Consequently, the linear relationship between voltage and current appearing at the output of the tank circuit does indeed support the application of theorems that require circuits to be linear for their application to be valid.

Another misconception concerns the relationship between the output and load resistances of these amplifiers. Because of wild speculations without verification by *valid* measurements, many people believe incorrectly that the output resistance is much greater than the load resistance, and thus proclaim that a conjugate match cannot be obtained between the amplifier and the load. However, when a *linear* source of power is delivering *all* of its available power to the load, there is a conjugate match by axiomatic definition, as explained in the following paragraph. An example from Terman is used in clarifying the misconception concerning the relationship between output and load resistances, shown below in "Analysis of the Class-C Amplifier." In addition, data obtained from my own measurements, shown below in "Measuring the Output Resistance of the RF Power Amplifier," prove that after the amplifier has been adjusted to deliver all of its available power at any given drive level, the output and load impedances of the amplifier *are equal* and thus are conjugates of each other.[2] My measurements have been confirmed by Tom Rauch, W8JI, using an identical measurement procedure. Tom is an RF-power-amplifier engineer with Ameritron.

Now to explain two axioms of the conjugate matching theorem that are commonly overlooked, which has resulted in widespread confusion concerning its use. We know that when a load impedance differs from its source impedance, a matching device is required to allow delivery of all the available power from the source to the load. In this condition, we say the load is matched to the source. The term "matched" has been used universally for many decades, and in those earlier days, the term was used alone. However, when all the power available from the source is delivered to the load, the matching occurs because the source and load impedances are conjugates of each other. Consequently, during the last 50 years, the term conjugate match gradually came into use synonymously with "match" to describe the term more accurately. In other words, "match" (used in this context) and "conjugate match" are often used interchangeably with no difference in meaning. Unfortunately, misinterpretation and misunderstanding of conjugate in the newer term has created confusion for many people when a routine impedance match is referred to as a conjugate match. To clarify the confusion, the following two axioms, which follow from the maximum power-transfer theorem, accurately define a conjugate match:

Axiom 1: There is a conjugate match whenever all of the available power from a source or network is being delivered to the load.

Axiom 2: There is a conjugate match if the delivery of power decreases whenever the impedance of either the source or load is changed in either direction.

We now return to clarify the misconception concerning output and load resistances. The term source resistance, R_S, of an RF power amplifier, as is often misused (and confused with R_P) in referring to the source of RF power delivered by class-B and -C amplifiers, reveals still another prevalent misconception. This misconception is that the *entire source of power* in these classes of vacuum-tube amplifiers is a *dissipative* resistance. In clarifying this misconception, we will use the example by Terman to demonstrate that the source of RF output power in a class-C amplifier is the combination of *two* resistances; a *nondissipative* resistance (related to the characteristics of the *effective* load line) and a *dissipative* plate resistance R_{PD}. R_{PD} is not plate resistance R_P, as determined from the well-known expression $R_P = \Delta E_P / \Delta I_P$. From this expression it is evident that R_P is the result of a small change in plate current due only to a change in plate voltage, which is *not* the source of power in RF power amplifiers as is claimed by many who have misinterpreted the expression. The source of power is actually derived by a large change in plate current resulting from a change in *grid voltage*. This phenomenon will be discussed in more detail later.

One portion of the *nondissipative* resistance is the reciprocal of the total conductance from both plate and power supply to the input of the π-network tank circuit. At that point in the typical amateur, π-net class-B and -C amplifier, the load is the tank input. The source is the combination of two parallel conductive paths to the tank: (1) the blocking capacitor in series with the active device, the tube(s)[3] and (2) the same blocking capacitor in series with the RF choke and the voltage of the power supply. These two conductance paths are paralleled at the input of the tank, operating at different, but overlapping times throughout the cycle. The other portion of the nondissipative resistance is related to the operating load line, which will be discussed below.

Plate resistance R_{PD} *is* dissipative, whose value is determined by the power P_D dissipated as heat by the plate divided by the square of the average dc plate current I_{dc}, the current measured by the dc plate ammeter. Notice in Terman's statement #3 below, that dissipated power P_D is the product of the instantaneous plate-to-cathode voltage and the instantaneous plate current. We know that energy is transferred from the plate circuit of the amplifier to the π-network by periodic pulses of plate current that flow during the conducting portion of the RF cycle. Knowledge of the nondissipative portion of the source resistance will allow you to understand why class-B and -C amplifiers can deliver all of their available power into a conjugately matched load with efficiencies greater than 50%. This concept is important, because the ability of these amplifiers to be conjugately matched has been incorrectly disputed due to three erroneous assumptions that have caused many amateurs and engineers to be misled.

Erroneous Assumptions

The principal reason that many people have been misled is that they have incorrectly estimated the amount of the source resistance in the amplifier that is dissipative. This incorrect assumption led them to believe that half the power is dissipated in the source resistance, and thus, as in the classical generator, a conjugate match would limit the efficiency to 50%. However, this is not true, because, as noted above, the source of the power delivered to the π-network tank circuit is nondissipative, except for the dissipative plate resistance R_{PD}. Because dissipative plate resistance R_{PD} is generally less than the load resistance R_L, more power is delivered to the load resistance than that dissipated in the dissipative plate resistance, thus allowing efficiencies greater than 50%. The lower dissipative plate resistance occurs because plate current is allowed to flow only

[1]Notes appear on page 44.

when the plate voltage is at the minimum of its sinusoidal swing, as explained in Terman's statement #7 below. Repeating Terman's statement #3 for emphasis, *dissipated power* P_D *is the product of instantaneous plate-to-cathode voltage and instantaneous plate current*. (Keep in mind that plate current is zero except during the short conduction time, considerably less than 180°.)

The second erroneous assumption is that the conjugate matching and maximum power-transfer theorems don't apply to class-B and -C RF amplifiers, because the operation of these amplifiers is nonlinear (see Note 1). This assumption is also incorrect because they have failed to appreciate the isolating action of the vitally important π-network tank circuit. The π-network tank is not simply an impedance transformer, as many believe, but is also an energy-storage device. The energy-storage capacity of the tank isolates the pulsed nonlinear mode at the input from the smoothed linear mode at the output that delivers the nearly perfect sine waves. This widely overlooked and misunderstood concept will be discussed in depth later.

A third erroneous assumption concerns the misuse of the role source resistance R_S plays in the delivery of power to the tank circuit. Because some say the value of R_S is as much as five times greater than load resistance R_L (a condition that violates the conjugate matching theorem), some people assert that no conjugate match is possible in systems where the source is an RF power amplifier. However, to obtain R_S they erroneously used the expression $R_P = \Delta E_P / \Delta I_P$, where R_P *is* greater than R_L. The reason the expression was used erroneously is that—in this expression—I_P varies only with variation of plate voltage, not grid voltage, as explained earlier. Because the change in plate current due to a change in plate voltage is small compared to the change in plate current due to a change in grid voltage, R_P and the erroneous 'R_S' are much greater than R_L. The crucial point here is that the source of power is derived from the much larger change in plate current due to the change in grid voltage, while the effect of the change in plate current due to the change in plate voltage is insignificant in relation to the output impedance of the amplifier. Consequently, as we proceed we will learn that both R_P and 'R_S', as perceived by some, are totally irrelevant to conjugate matching the output impedance of the amplifier to the impedance of its load, and thus impose no impediment to the conjugate match.

Analysis of the Class-C Amplifier

The following discussion of the class-C amplifier, which reveals why the portion of the source resistance related to the characteristics of the load line is nondissipative, is based on statements appearing in Terman's *Radio Engineers Handbook*, 1943 edition, p 445, and on Terman's example of class-C amplifier design data appearing on p 449. Because the arguments presented in Terman's statements are vital to understanding the concept under discussion, I quote them here for convenience (parentheses and emphasis mine):

1. The average of the pulses of current flowing to an electrode represents the direct current drawn by that electrode.
2. The power input to the plate electrode of the tube at any instant is the product of plate-supply voltage and instantaneous plate current.
3. The corresponding power (P_D) lost at the plate is the product of instantaneous plate-cathode voltage and instantaneous plate current.
4. The difference between the two quantities obtained from items 2 and 3 represents the useful output, at the moment.
5. The average input, output and loss are obtained by averaging the instantaneous powers.
6. The efficiency is the ratio of average output to average input and is commonly of the order of 60-80%.
7. The efficiency is high in a class-C amplifier *because current is permitted to flow only when most of the plate-supply voltage is used as voltage drop across the tuned load circuit* R_L*, and only a small fraction is wasted as voltage drop (across* R_{PD}*) at the plate electrode of the tube.*

Based on these statements, the discussion and the data in Terman's example that follow explain why the amplifier can deliver power with efficiencies greater than 50% while conjugately matched to its load, a condition that is widely disputed because of the incorrect assumptions concerning class-B and -C amplifier operation, as noted above. The terminology and data in the example are Terman's, but I have added one calculation to Terman's data to emphasize a parameter that is vital to understanding how a conjugate match can exist when the efficiency is greater than 50%. That parameter is dissipative plate resistance R_{PD}. (As stated earlier, dissipative resistance R_{PD} should not be confused with plate resistance R_P of amplifiers operating in class A, derived from the expression $R_P = \Delta E_P/\Delta I_P$.)

It is evident from Terman that the power supplied to the amplifier by the dc power supply goes to *only two places,* the RF power delivered to load resistance R_L at the input of the π-network, and the power dissipated as heat in dissipative plate resistance R_{PD}. (Again, this is not plate resistance R_P, which is *totally irrelevant* to obtaining a conjugate match at the output of class-B and -C amplifiers.) In other words, the output power equals the dc-input power minus the power dissipated in resistance R_{PD}. We will now show why this two-way division of power occurs. First, we calculate the value of R_{PD} from Terman's data, as seen in Eq 9 of the example below. It is evident that when the dc-input power minus the power dissipated in R_{PD} equals the power delivered to resistance R_L at the input of the π-network, there can be no significant dissipative resistance in the amplifier other than R_{PD}. The antenna effect from the tank circuit is so insignificant that dissipation due to radiation can be disregarded. If there were any significant dissipative resistance in addition to R_{PD}, the power delivered to the load plus the power dissipated in R_{PD} would be less than the dc-input power, due to the power that would be dissipated in the additional resistance. This is an impossibility, confirmed by the data in Terman's example, which is in accordance with the law of conservation of energy. Therefore, we shall observe that the example confirms the total power taken from the power supply goes only to (1) the RF power delivered to the load R_L and (2) to the power dissipated as heat in R_{PD}, thus proving there is no significant dissipative resistance in the class-C amplifier other than R_{PD}.

Data from Terman's example on p 449 of *Radio Engineers Handbook*:

$$E_b = \textit{dc Source Voltage} = 1000\text{ V} \quad \text{(Eq 1)}$$

$$E_{min} = E_b - E_L = 1000 - 850 = 150\text{ V} \quad \text{(Eq 2)}$$

See Terman, Figs 76A and 76B.

$$I_{dc} = \textit{dc Plate Current} = 75.1\text{ mA} = 0.0751\text{ A} \quad \text{(Eq 3)}$$

$$E_L = E_b - E_{min} = 1000 - 150 = 850\text{ V} = \textit{Peak Fundamental ac Plate Voltage} \quad \text{(Eq 4)}$$

$$I_1 = \textit{Peak Fundamental ac Plate Current} = 132.7\text{ mA} = 0.1327\text{ A} \quad \text{(Eq 5)}$$

$$P_{IN} = E_b \times I_{dc} = \textit{dc Input Power} = 1000 \times 0.0751 = 75.1 \text{ W} \quad \text{(Eq 6)}$$

$$P_{OUT} = \frac{(E_b - E_{min})I_1}{2} = \frac{E_L I_1}{2} = \textit{Output Power Delivered to } R_L$$
$$= \frac{(1000-150)0.1327}{2} = 56.4 \text{ W} \quad \text{(Eq 7)}$$

$$P_D = P_{IN} - P_{OUT} = \textit{Power Dissipated in Dissipative Plate Resistance } R_{PD} = 18.7 \text{ W} \quad \text{(Eq 8)}$$

$$R_{PD} \frac{P_D}{{I_{dc}}^2} = \frac{18.7 \text{ W}}{0.0751^2 \text{ A}} = \textit{Dissipative Plate Resistance } R_{PD} = 3315.6\ \Omega \quad \text{(Eq 9)}$$

$$R_L = \frac{E_b - E_{min}}{I_1} = \frac{E_L}{I_1} = \textit{Load Resistance} = \frac{850}{0.1327} = 6405\ \Omega \quad \text{(Eq 10)}$$

(6400 Ω in Terman)

$$\textit{Plate Efficiency} = P_{OUT} \times \frac{100}{P_{IN}} = 56.4 \times \frac{100}{75.1} = 75.1\% \quad \text{(Eq 11)}$$

Notice that in Eq 10, R_L is determined simply by the ratio of the fundamental RF ac voltage E_L divided by the fundamental RF ac current I_1, and therefore does not involve dissipation of any power. Thus R_L is a nondissipative resistance.

Referring to the data in the example, observe again from Eq 10 that load resistance R_L at the input of the π-network tank circuit is determined by the ratio E_L / I_1. This is the Terman equation which, prior to the more-precise Chaffee Fourier analysis, was used universally to determine the approximate value of the optimum load resistance R_L. (When the Chaffee analysis is used to determine R_L from a selected load line the value of plate current I_1 is more precise than that obtained when using Terman's equation, consequently requiring fewer empirical adjustments of the amplifier's parameters to obtain the optimum value of R_L.) Load resistance R_L is proportional to the slope of the operating load line that allows all of the available integrated energy contained in the plate-current pulses to be transferred into the π-network tank circuit. (For additional information concerning the load line, see below.) Therefore, the π-network must be designed to provide the equivalent optimum resistance R_L looking into the input for whatever load terminates the output. The current pulses flowing into the network deliver bursts of electrical energy to the network periodically, in the same manner as the spring-loaded escapement mechanism in the pendulum clock delivers mechanical energy periodically to the swing of the pendulum. In a similar manner, after each plate-current pulse enters the π-network tank circuit, the flywheel effect of the resonant tank circuit stores the electromagnetic energy delivered by the current pulse, and thus maintains a continuous sinusoidal flow of current throughout the tank, in the same manner as the pendulum swings continuously and periodically after each thrust from the escapement mechanism. The continuous swing of the pendulum results from the inertia of the weight at the end of the pendulum, due to the energy stored in the weight. The path inscribed by the motion of the pendulum is a sine wave, the same as at the output of the amplifier. We will continue the discussion of the flywheel effect in the tank circuit with a more in-depth examination later.

Let us now consider the dissipative plate resistance R_{PD}, which provides the evidence that the dc input power to the class-C amplifier goes only to the load R_L and to dissipation as heat in R_{PD}. With this evidence, we will show how a conjugate match can exist at the output of the π-network with efficiencies greater than 50%. In accordance with the conjugate matching theorem and the maximum power-transfer theorem, it is well understood that a conjugate match exists whenever all available power from a linear source is being delivered to the load. Further, by definition, R_L is the load resistance at the tank input determined by the characteristics of the load line that permits delivery of all the available power from the source into the tank. This is why R_L is called the *optimum* load resistance. Thus, from the data in Terman's example, which shows that after accounting for the power dissipated in R_{PD}, all the power remaining is the available power, which is delivered to R_L and thence to the load at the output of the π-network. *Therefore, because all* available *power is being delivered to the load, we have a conjugate match by definition.* In a following section we will show how efficiencies greater than 50% are achieved in class-C amplifiers operating into the conjugate match.

Examining the Operating Load Line

The details of the somewhat trial-and-error method of establishing the operating load line are beyond the scope of this article. However, once established, the load line represents the nondissipative load resistance R_L appearing at the input of the π-network tank circuit. The slope of the load line is proportional to the ratio of the continuous fundamental RF voltage and current. When the network is terminated with the correct output load resistance (a resistance equal to the network output resistance as explained below), the network transforms the output load resistance up to resistance R_L at the network input. Once established (and proven by measurements of network output impedance), the slope of the operating load remains constant with changes in output power resulting from changes in drive levels. Consequently, because R_L represents the slope of the load line, both the fundamental RF voltage-current ratio appearing along the load line and the network output impedance remain constant whatever the power level of the integrated current pulses enter the network. It should be clearly understood that, because the operating load line, and the optimum resistance R_L it represents, are established solely by the *ratio* of the RF voltage and current, the load line and R_L are nondissipative. As explained earlier, the entire dissipation to heat occurs only in the dissipative plate resistance R_{PD}.

When using the Terman equation to determine load resistance R_L, an approximate load line and average plate current are first estimated from the tube characteristic curves. The corresponding value of R_L is used as a trial value and the output power and efficiency are determined in a trial run. However, several trial runs with different load adjustments are necessary to converge toward the optimum value of R_L that will yield the desired conditions for operation, simply because the first estimation of average plate current is rarely the optimum value.

When the Chaffee analysis is used to determine R_L in establishing the load line, the average value of plate current I_1 during the conduction period is obtained by first plotting the load line on a graph of constant plate current characteristics of the tube. The load line is then marked off in several increments corresponding to successive angles of conduction of plate current. The plate current at each conduction angle is then found at the intersection of the load line and the constant-current curve. The plate voltage at each conduction angle is also found on the plate voltage line directly below the above stated intersection. The averages of plate current and voltage are then determined using the trapezoidal rule. Load resistance R_L is then determined by dividing the average fundamental RF plate voltage by the average fundamental RF plate current, the Terman equation. Thus, the Chaffee method saves time compared to using Terman's equation alone, because the initial value of average plate current is closer to the optimum value than that estimated for use in the Terman equation.

Calculation of Efficiency Greater than 50%

To show how efficiencies greater than 50% are obtained while the amplifier is conjugately matched, we will dissect the data in the Terman example to discover that load resistance R_L is greater than dissipative plate resistance R_{PD}, thus allowing more power to be delivered to the load than that dissipated in R_{PD}. Referring again to Terman's example in (Eq 10), his calculation of load resistance R_L is 6400 Ω. From (Eq 9) we find R_{PD} is 3315.6 Ω by dividing 18.7 W dissipated in R_{PD} by the square of 75.1 mA dc plate current I_{dc} flowing through R_{PD}. Correspondingly, (Eq 7) shows the power delivered to R_L is 56.4 W, and from (Eq 8), power P_D dissipated in R_{PD} is 18.7 W. With 56.4 W delivered to R_L and 18.7 W dissipated in R_{PD} we have accounted for the total input power, 71.5 W, shown in Eq 6. The relative power distribution is 75.1% delivered to R_L, and 24.9% dissipated in R_{PD}. Earlier we showed that after accounting for the power dissipated in R_{PD}, all the remaining available power is delivered to the load R_L. Thus, this distribution of power clearly demonstrates why a class-C amplifier can deliver more than 50% of its input power to the load, because its load resistance R_L (6400 Ω) is greater than its dissipative plate resistance R_{PD} (3315.6 Ω). These calculations are in accord with with Terman's statement #7 that "efficiency is high in the class-C amplifier, because current is permitted to flow only when most of the plate-supply voltage is used as voltage drop across the tuned load circuit R_L, and only a small fraction is wasted across R_{PD} at the plate electrode of the tube." None is dissipated in the non-dissipative resistance related to the characteristics of the load line. As stated earlier, the nondissipative portion of the source resistance is the reciprocal of the total conductance from both the plate of the tube and the power supply to the input of the π-network tank circuit. It should be noticed however, that we are considering only the power delivered to the tank; we are not concerned here with inherent loss in the tank that results in some decrease in the power delivered at the output of the tank.

Evidence of Conjugate Match

The example has proven that a conjugate match exists, because all the available power has been delivered to the resistive load R_L, and thence to the load terminating the π network, in accordance with conjugate matching axioms 1 and 2 recited above. The example has also shown that more power has been delivered to the load than was dissipated, because 54.6 W were delivered and only 18.7 W were dissipated. Thus, contrary to the opinion of many who fail to understand this concept, we have shown that conjugate matching to a class-C RF amplifier does not limit its efficiency to 50%. The same reasoning applies to amplifiers operating in class B (see Note 2).

So now you ask, "Do we have a conjugate match during SSB operation?" The answer is yes, but it begs an additional question: Does the output impedance of the amplifier remain constant with SSB modulation, or does it change during the variations of drive and output power corresponding to the voice modulation? My measurements, described below, show that the output impedance does not change significantly with voice modulation. This is because, for a given load resistance R_{LOAD}, the operating load line related to the load resistance R_L appearing at the input of the tank circuit, and the output resistance R_{OUT}, are established during the tuning and loading procedure when the loading is adjusted to deliver maximum available power. During this procedure, maximum available power is that power delivered to the load with the drive level set to obtain the desired output power at the full modulation level. After the load line has been established in this manner, it remains constant for all values of drive. I have made extensive measurements, which show that once the operating load line is established during this routine procedure, it remains constant during swings of grid voltage during SSB modulation, as long as the plate supply voltage remains constant.

So now we ask, "Is the conjugate match of such importance that we should be concerned about it?" Yes it is, if we are to understand why antenna tuners perform their intended task of matching the complex impedance appearing at the input of a transmission line that is Z_0 mismatched to an antenna, while also establishing a conjugate match that overrides the Z_0 mismatch at the antenna. The principles of conjugate matching are fundamental to the matching function performed by the antenna tuner, and are indeed fundamental to all impedance matching obtained with any impedance matching device that allows delivery of all available power from its source!

The Vital Role of Energy Storage in the Tank Circuit: Providing Linear Operation at the Output

We now return to conduct a close examination of the vitally important flywheel effect of the tank circuit. The energy storage (Q) in the tank produces the flywheel effect that isolates the nonlinear pulsed energy entering the tank at the input from the smoothed energy delivered at the output. Because of this isolation, the energy delivered at the output is a smooth sine wave, with linear voltage/current characteristics that support the theorems generally restricted to linear operation. We know that the widely varying voltage/current relationship at the tank input results in widely varying impedances, which precludes the possibility of a conjugate match at the input of the tank circuit. However, the energy stored in the tank provides constant impedance at the output that supports both the conjugate matching and the maximum power-transfer theorems.

The acceptance by many engineers and amateurs of the notion that the output of the RF tank is nonlinear is a reason some readers will have difficulty in appreciating that the output of the RF tank circuit is linear and can thus support the conjugate match. Valid analogies between different disciplines are often helpful in clarifying difficulties in appreciating certain aspects of a particular discipline. Fortunately, energy storage in the mechanical discipline has a valid and rigorous analogous relationship with energy storage in LC circuitry, which makes it appropriate to draw upon a mechanical example to clarify the effect of energy storage in the RF tank circuit. (A further convincing analogy involving water appears later, in which the origin of the term tank circuit is revealed.)

The smoothing action of the RF energy stored in the tank circuit is rigorously analogous to the smoothing action of the energy stored in the flywheel in the automobile engine. In the automobile engine, the flywheel smoothes the pulses of energy delivered to the crankshaft by the thrust of the pistons. As in the tank circuit of the amplifier, the automobile flywheel is an energy storage device, and the smoothing of the energy pulses from the pistons is achieved by the energy stored in the flywheel. In effect, the flywheel delivers the energy to the transmission. The energy storage capacity required of the flywheel to deliver smooth energy to the transmission is determined by the number of piston pulses per revolution of the crankshaft. With more pistons, less storage capacity is required to achieve a specified level of smoothness in the energy delivered by the flywheel. The storage capacity of the flywheel is determined by its moment of inertia, and the storage capacity of the tank circuit in the RF amplifier is determined by its Q.

As stated earlier, the tank circuit in the RF amplifier receives two overlapping pulses of energy per cycle. If the effect of the overlapping pulses were considered a single pulse, we would have a condition that is somewhat analogous to an engine having only one cylinder. If we were to assume that the piston in the one-cylinder engine delivers one thrust of energy per revolution, it is evident that a large amount of energy storage is required to enable the crankshaft to deliver a smooth output during the entire rotation of the crankshaft. In this case, a very heavy flywheel is required to deliver a smooth output. This is also the case with the RF tank circuit, which requires a Q of 10 to 12 to yield a smooth sine-wave output with an acceptable minimum of harmonic ripple. Because the tank receives only one pulse of energy per cycle, it must store many times the amount of energy it passes through, to provide a continuous sine-wave output when supplied only with pulses of energy at the input. Thus the energy-storage capacity of the tank provides for the smoothed linear output to the load circuit, despite the nonlinear pulsed input, which, for the purpose of analysis, allows the pulsed source and tank to be replaced with an equivalent Thévenin generator whose output impedance equals R_{LOAD}. Although no conjugate match exists at the input of the π-network tank (because of the large variation of impedance in the current pulses) the isolation derived from the flywheel action of the tank thus allows a conjugate match to exist between the output of the tank and its load, a concept which will become clear as we continue.

To further clarify the action and the effect of the energy storage in the tank that achieves a linear voltage/current relationship at the output of the tank, parts of the following discussion are paraphrased from correspondence with Dr. John Fakan, KB8MU. It should be emphasized here that a conjugate match can exist between the output of the RF power amplifier and its load because of the linear voltage/current relationship at the output of the tank resulting from the energy storage in the tank.

The tank circuit stores the energy by passing it back and forth cyclically between the L and C components, and passes only a fraction of that energy to the tank's load on each cycle. Because the amount passed to the load is such a small fraction of the total stored in the tank, and because even that small amount is restored during the cycle, the tank can be considered an active source. Because it can be considered as an active source, we have no need for interest in what is going on ahead of it in the overall system (as far as what the downstream devices see).

Consider that when designing to get energy from a power supply our concern is only with the characteristics seen at the power supply terminals. Our design does not depend on whether the line-feed to the supply is single-phase or three-phase, 60 Hz or 400 Hz, or even if the power factor is unity or some other value. These things don't matter once you know what shows up at the output connections of the supply. For our purposes, the actual source of energy is the connection at the output of the supply, and the characteristics at that point will be determined by the components in the filter circuitry.

As a source of sinusoidal energy, our RF amplifiers are no different. The source of this energy that will be passed on to our antenna system is the tank circuit. The load connected to the output port of the amplifier can only see the voltage swings and the impedance presented by the tank components. A properly designed tank (of any type) will not pass so much energy on each cycle that the relationship between its terminal voltage and current is affected enough to cause nonlinearity. Sometime during the cycle even that small amount of energy will be replaced, thus maintaining its operating levels.

Because this "new" source happens to present a linear impedance to its load (the first connection in our antenna system) we need have no concern about nonlinear processes occurring at points upstream of the tank circuit. Once we have a linear active source in the cascade and we do nothing downstream to subsequently cause nonlinearities, we can take advantage of those theorems and ideas that depend on the linearity of the network.

My teachings in "Reflections" depend on the linear nature of the energy transfer from the amplifier's output port right on through to the last antenna element. Because the energy to this network is supplied by a linear source (the tank circuit) everything in my teachings can be assured of sound scientific basis. Objections by others, based on nonlinearities ahead of the tank are simply not applicable.

The energy pulses supplied to the tank must be sufficient to "refill" the tank's energy store on each cycle. The connection where that energy transfer occurs is at the input to the tank. As stated earlier, at that point in the typical amateur π-network class-C amplifier, the load is the tank input. The source is the combination of two processes: (1) the blocking capacitor in series with the active device (tube) and (2) the same blocking capacitor in series with the RF choke and the voltage of the power supply. These two sources are in parallel at the connection, but operate at different, overlapping times through the cycle.

The load resistance R_L appearing at the input to the tank is determined by the value required to accommodate the energy transfer required per cycle to make up for that being transferred by the tank to its load. Because of the lack of linear or even simple square-wave characteristics of the active device, the designs in this region have always been very empirical. Actual experience and a good seat-of-the-pants feel for the significance of active-device data sheets have been the main tools for the design of tank circuits. The amount of energy delivered via the action of the active device (the tank) is dependent on things like drive, feedback, supply voltages and so on. They all can play a role in providing for the right amount of energy transfer to allow the tank to function as a linear active source.

If the tank does not receive enough energy to sustain the power level it has established with its load, its output will decrease accordingly. Malfunctioning of the upstream energy "bucket brigade" can result in linear operation at a lower level or in nonlinear operation, depending on how well the tank design can handle the changes. The important point is that once conditions allow the tank to operate as a linear active source, everything else downstream of the tank is linear and follows the conjugate-match theorem and other linear-system theories. Changing the conditions at the input to the π-network (for example: changing drive, feedback and so on) affect the performance of the tank as a linear source of RF energy. If the tank is supplied with energy pulses having a different integrated average energy than that being supplied to the tank's load, the tank's output characteristics will change to fit the available energy. It will do this by changing its output impedance to whatever value is required so that, at the new conjugate matching point, the voltage-current product will equal the energy rate available. It has no other choice because the conjugate-matching theorem requires a change in output impedance if it is to remain a linear source. Consequently, the load impedance at the output must change accordingly to retain the conjugate match. If the changes are extreme, it may not be able to accommodate the required impedance change in a linear manner, so wave-shape distortion could occur, for example, flat topping.

The important point is that the design and operation of the circuitry providing energy to the tank circuit involves a number of issues having to do with protection of the active device, stability, efficiency and such, as well as the amount of energy transferred to the tank during each cycle. It doesn't matter that the wave shape of the energy pulse is ugly and would be difficult to characterize mathematically, because the tank circuit doesn't care.

It is positively uncanny how easy it is for some people to simply ignore experimental evidence staring them right in the eye when one of their pet understandings is in jeopardy. Many people concede that amateur class-C amplifiers typically operate at greater than 50% efficiency. They will also agree that it is common to tune for a power peak. They will then wiggle and squirm to avoid agreeing that the tuning process is simply matching the output and load impedances to a common conjugate. Their reason is that the internal "resistance" precludes the higher than 50% efficiency. The fact that there are two independent definitions for the word resistance doesn't seem to matter. They are completely ignoring definition #2 of the IEEE definitions of resistance, the nondissipative resistance, that is, the real part of the impedance.

RF power amplifiers are necessarily designed to match load impedances at or near the characteristic impedance of common coaxial transmission lines. No other design value would make sense. The conjugate-match theorem is simple and absolute: When the energy being transferred across any linearly behaving connection cannot be further increased simply by changing the impedance of either source or load, a conjugate match exists. That is commonly the operating condition for amateur class-B and -C RF amplifiers. From the tank circuit forward, the behavior is linear, because the voltage and current at the output of the tank are continuous and sinusoidal due to the energy storing (flywheel), smoothing action of the tank. There really is no wiggle room for debate.

Origin of the Term "Tank Circuit"

Let me digress for a moment to say that it is customary for an author of an article such as this to have his writing peer reviewed to uncover possible errors that may have escaped him. Because of the protracted and unfortunate controversy brought about by those who claim that a conjugate match cannot exist in an RF system powered by an RF power amplifier, my engineering credibility as an author has been questioned. Therefore, because this article is primarily concerned with presenting a convincing argument that a conjugate match does indeed exist in RF power amplifiers, I have attempted to make sure it contains no conceptual or substantive errors, or invalid statements. Consequently, I requested several professional RF engineers with unquestionable credentials and expertise to review and critique it. All reviewers but one found my presentation accurate. This dissenting reviewer flatly rejected the concept that a π-network tank circuit isolates its pulsed input from the output, and therefore he maintains that the output circuit of the π-network cannot support or sustain linear operation, and no conjugate match. Unfortunately, during the nine years of the controversy I discovered that opposition to the application of linear theorems to *any* aspect of RF power amplifier operation is prevalent in the confusion of *many* otherwise intelligent and capable engineers. It therefore occurred to me that others also might have similar difficulty in accepting the concept of energy storage in the tank circuit providing isolation between the input and output of the tank that allows linear theorems to be valid at the output. I have already presented two examples of the storage of mechanical energy that illustrate the smoothing function of energy storage, which are precisely analogous to energy storage in the tank circuit of the RF power amplifier. In addition, a valid water analogy where the operative word is "tank" in the literal sense might further clarify the issue. I also believe you'll find it interesting to learn how the term *tank* originated as an active description of the LC circuit used in the output coupling of all discontinuous RF power amplifiers.

Legend has it that in the early days of RF amplifier development the water-tank analogy was applied for the very purpose of explaining the energy-storage function of the LC output circuit. It goes like this. A water tank is filled to a specific depth that causes a corresponding pressure applied on the bottom. A hole is made in the bottom with a size that allows one gallon per minute to flow with the specific applied pressure. Water is added at the top of the tank at the rate of one gallon per minute, thus maintaining the original level in the tank as the water flows smoothly out from the bottom. Let's now consider how the water is added at the top. It can be added in spurts, but the water flowing out from the bottom will still flow smoothly without ever knowing the nature of the spurts added at the top. The spurts can be added at a rate of one gallon dumped in every minute, a half gallon twice during the minute, one-thirtieth of a gallon thirty times per minute and so on, you get the picture. As long as enough water is added to maintain the level, thus maintaining the same constant pressure at the bottom, the water will continue to flow smoothly from the bottom at the rate of one gallon per minute, regardless of how the water is added. It is the *energy stored in the tank* that isolates the intermittent additions of water at the tank top from the continuous flow at the bottom. If the tank is filled to a greater depth, pressure at the bottom is increased, resulting in an increased rate of flow of water at the bottom, in direct proportion to the increase in pressure.

It should be appreciated that the *fluid impedance* at the output of the tank is the ratio of the pressure to the flow rate. This sets the rate at which the energy contained in the water flows from the bottom of the water tank. The tank output is established solely by the size of the hole and the height of the water above the hole. The same energy rate can exist with a tall tank and a small outlet hole (high output impedance), or shorter tanks with appropriate larger holes (low output impedance). However, the impedance and linearity of the *input* to the tank is irrelevant as long as it results in maintaining a constant water level. Thus, the action at the bottom is linear although the action at the top is not.

The same is true in the *tank* circuit of the RF amplifier. The impedance at the output of the RF tank is the ratio of the voltage to current at which power is being delivered to the tank's load. The voltage and current appearing at the output of the π-network tank circuit are analogous to the water pressure at the bottom of the tank (voltage), and the rate of flow of the water (current) out of the tank. As in the water tank, the shape of the current pulses entering the π-network tank has no effect on the smooth sinusoidal voltage and current appearing at the output. If the average integrated energy of the current pulses entering the tank increases, the voltage and current at the output will increase in a linear relationship. Thus, it is shown that the output of a properly designed RF tank circuit is linear, and the theorems associated with linear circuits are applicable.

Measuring the Output Resistance of the RF Power Amplifier

I have developed a test setup and procedure based on the standard IEEE load-variation method for measuring the source, or output resistance R_{OUT} of networks, which are described below. Measurements made with this setup and procedure show that output resistance R_{OUT} equals the load resistance R_{LOAD} when the amplifier is initially adjusted to deliver all of its available power to that load, thus proving the existence of a conjugate match. However, before proceeding further it will be helpful to obtain an appropriate perspective by reviewing some background concerning the issue.

There has never been a problem in determining the correct value of load resistance R_L appearing at the input of the π-network tank circuit of the RF amplifier. Resistance R_L is routinely calculated using either the Terman equation or the more precise Chaffee analysis to determine the slope and other characteristics of the operating load line, as mentioned earlier. After the network has been adjusted to deliver its intended power into its terminating load, resistance R_L appearing at the input of the network is easily and routinely measured with appropriate impedance-measuring equipment with the amplifier powered down.

However, determining the *output* resistance R_{OUT} appearing at the output of the π-network has been daunting. Wild speculations (sans measurements) concerning the output resistance abound because of the misunderstandings and incorrect assumptions concerning the actions of the tank circuit as described above. The misconception that a conjugate match cannot exist at the output of RF power amplifiers has precluded logical reasoning, that when the amplifier is delivering all its available power there is a conjugate match by definition. Consequently, it has been considered unthinkable that the output source resistance could possibly be equal to the load resistance.

I am not aware of any writings in the professional literature that discuss the measurement of R_{OUT}. A probable reason for this lack of discussion is that knowledgeable people know that R_{OUT} must equal the load resistance when all the available power is being delivered, and it would therefore be redundant to state it. Because of the controversy surrounding conjugate matching and amplifiers, it is now appropriate to describe the test setup and procedure that does yield the correct value of source resistance R_{OUT}, *the value that equals the load resistance*. Consequently, using the standard IEEE load-variation procedure described below, it will be seen that the data resulting from my measurements (also shown below) prove two things: (1) source resistance is *not* what some previous authors were measuring, and (2) my measurements prove the existence of the conjugate match at the output of RF power amplifiers. The data obtained from my measurements have been verified by Tom Rauch.

The test setup I developed for measuring the output resistance R_{OUT} of RF power amplifiers is arranged to use the load-variation method of measurement, based on the IEEE expression for measuring the output resistance of networks. The IEEE expression is $R_{OUT} = \Delta E / \Delta I$, where ΔE and ΔI represent the corresponding changes in load voltage and load current, respectively, with a small change in load resistance R_{LOAD} terminating the network. In the measurements described below, all values of R_{LOAD}, (*R1* and *R2*) are pure resistances, $R + j0$. In these measurements, the output load resistance R_{OUT} (*R1*) terminating the π-network is selected and the param-

Table 1—Using Standard IEEE Small-Load-Variation Method to Measure Network Output Source Resistance

Load Resistance (Ω)	*Load Voltage (V)*	*Load Current (A)*	*Output Resistance (Ω)*	*Measured Power Out (W)*
51.2	75.9	1.482	52.7	112.5
44.6	70.6	1.583		111.6
	Δ =5.3	Δ = 0.101		
51.2	76.9	1.502	51.2	115.5
44.6	71.6	1.605		114.9
	Δ = 5.3	Δ = 0.1034		
51.2	69.75	1.36	49.4	94.9
46.4	66.29	1.43		94.8
	Δ = 3.46	Δ = 0.70		
51.2	62.5	1.22	51.7	76.25
46.4	59.4	1.28		76.0
	Δ = 3.1	Δ = 0.60		
51.2	77.8	1.519	47.8	118.2
46.4	74.1	1.597		118.3
	Δ = 3.7	Δ = 0.078		
51.2	77.5	1.514	47.4	117.3
47.75	74.9	1.569		117.5
	Δ = 2.6	Δ = 0.0549		
		Average 50.3 Ω		

eters of the amplifier are then adjusted to obtain delivery of all the available power into that load at a given drive level. Then by varying the load resistance a small amount (to *R2*), around a 10% change from *R1*, and then measuring the difference in load voltage and current, the output resistance is obtained by substituting the differential voltage and current values in the IEEE expression for R_{OUT} shown above.

The equipment used in the measurements consisted of two vacuum-tube transceivers using two parallel 6146 tubes and a π-network tank circuit in the RF power amplifier. They are a Heathkit HW-100 and a Kenwood TS-830S. A Hewlett-Packard HP-4815A RF Vector Impedance Meter modified for digital readout, along with ESI 250-DA universal impedance bridge, for measuring RF and dc resistances of noninductive load resistors *R1* and *R2*. An HP-8405A Vector Voltmeter modified for digital readout for measuring voltages appearing across load resistors *R1* and *R2*, and an HP-410B RF Voltmeter with an HP-455A Coaxial Adapter, also modified for digital readout to indicate load voltage. The RF vector impedance meter was used to confirm that the load resistors contained zero reactance. The experiments were conducted at 4.0 MHz.

Procedure

The π-network output of the amplifier is initially terminated with *R1*, then tuned and loaded to deliver a specific maximum available output power with a given level of grid drive. The load voltage E_1 is measured with load *R1*, then the load is changed to *R2* and load voltage E_2 is measured. Load currents, I_1 and I_2, are then determined by calculation of $I = E/R$, using the measured values of *R* and *E*. Finally, as stated above, $R_{OUT} = \Delta E / \Delta I$, as shown in the data resulting from the measurements shown in Table 1.

The amplifier was tuned and loaded with its drive level set to deliver maximum available power of ≈110 W. All adjustments remain undisturbed thereafter. The data in Table 1 were obtained using the Heathkit HW-100. Result: Load resistance R_{LOAD} when adjusted for maximum power out $R1$ = 51.2 Ω. Average measured source resistance R_{OUT} = 50.3 Ω (See Fig 1 and Table 2).

The reason for the variation, or scatter in measured output resistance and output power seen in the data above was found to be a short-term sag in output power between the measurement of *R1* and *R2*. This problem was solved by changing the switching between *R1* and *R2* from manual to coaxial relay, thus reducing the time delay, and by waiting until the sag in power output bottomed out. However, the worst-case difference between *R1* load of 51.2 Ω and the *R2* value that yielded R_{OUT} of 47.35 Ω is a mismatch of only 1.081:1, with a reflection coefficient ρ of 0.039, for a conjugate mismatch loss of only 0.0066 dB.

After many more measurements similar to those above, except for adjusting the π-network for delivery of maximum available power prior to each measurement, I found that with load *R1* now at 51.0 Ω, the measured values of R_{OUT} varied randomly within 11 Ω on either side of the 51.0-Ω load with each measurement, that is, from 40 Ω to 62 Ω. However, I discovered the variation results from the very small slope of the power output curve near the peak. Using only the analog output-power meter to observe the point at which the power was maximum, I found it impossible to find the true peak in output power where R_{OUT} equals *R1* of 51 Ω, because the characteristics of the matching curve near its peak appear to be more like a round-top hill than a peak. Evidently, the adjustment for maximum output requires a method of indicating that provides better resolution than that obtained with the analog output-power meter alone. On the other hand, the mismatch between 51 and 40 Ω, and between 51 and 62 Ω, is only 1.28:1, for a voltage reflection coefficient ρ of 0.12, which results in a conjugate mismatch loss of only 0.066 dB at the maximum 11-Ω difference from

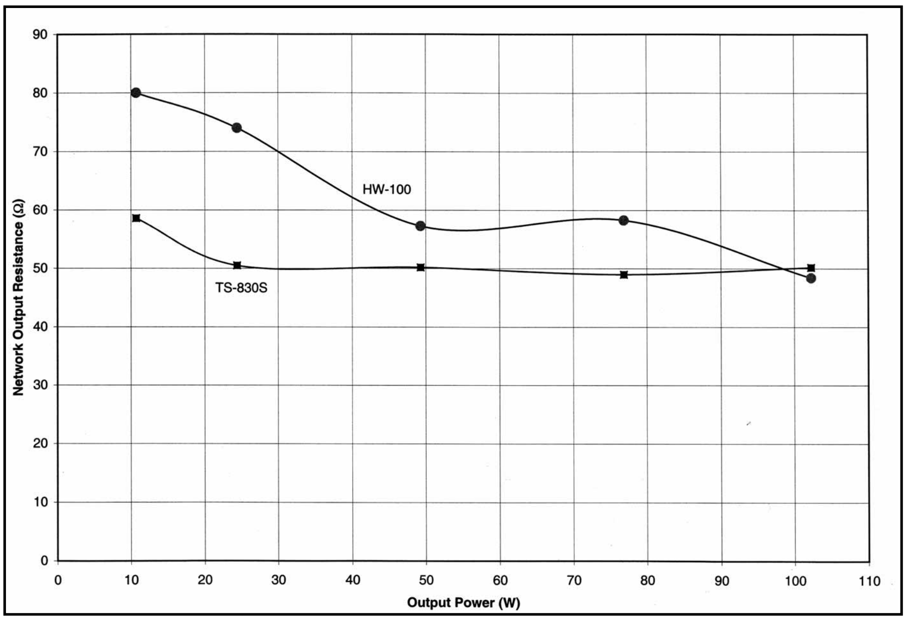

Fig 1—Measured network output resistance of a Heathkit HW-100 and a Kenwood TS-830S versus output power at 4.0 MHz, power level set by drive level (see Tables 2 and 3).

Table 2—Measured Network Output Resistance versus Output Power, HW-100 Transceiver

Also see Table 3 and Fig 1

Output Power (W)	*Network Output Resistance R_{OUT} (Ω)*	*Plate Voltage E_P (V)*	*Plate Current I_P, (mA)*
100.0	48.4	800	270
75.0	58.3	810	240
50.0	57.3	820	190
25.0	74.0	840	140
12.5	80.0	860	110
0.0	NA	890	70

Notice the increase in network output resistance with increase in plate voltage, due to poor power supply regulation as plate current decreases with decrease in output power.

51 Ω. Thus, it is evident that during routine tuning and loading adjustments using analog meters to indicate peak power output, the actual output resistance of the network during the measurements will seldom be *exactly* the value of the load resistance, but the consequence of the small difference is insignificant.

The next step in the procedure yielded the increase in resolution of the data required to find the exact point on the output curve where the output is maximum and R_{OUT} equals the load *R1*. After the maximum output that could be obtained by observing the indication on the analog output-power meter, R_{OUT} was measured at that point. The π-network was then re-adjusted in very small increments, measuring R_{OUT} after each re-adjustment, until R_{OUT} became equal to 51.0 Ω. The increments were so small that although the difference in output power at each increment was indiscernible on the analog power meter, it was clearly indicated by the digital voltmeter. Thus, it is shown that the output or source resistance R_{OUT} of an RF power amplifier is equal to the resistance of the load when the maximum available power of the source is being delivered to the load. It is therefore also evident that a conjugate match exists when the conditions just stated are present.

In addition to measuring R_{OUT} with the load resistance of 51.0 Ω, R_{OUT} was also measured with load resistances of 25 and 16.7 Ω. Using the same technique as described above, R_{OUT} measured 25 Ω when the load resistance was 25 Ω, and—consistent with the measurement pattern already developed—R_{OUT} measured 16 Ω when the load resistance was 16.7 Ω. These measurements indicate that, when the loading is initially adjusted to deliver maximum available power to any value of load resistance within the matching capability of the π-network, R_{OUT} equals the load resistance.

There is more. So far, we have only considered the condition in which the amplifier is delivering its maximum available power in the CW mode. However, we would also like to know what happens to output resistance R_{OUT} during SSB modulation after the amplifier is first tuned and loaded to deliver maximum available power with a specific drive level. The measurement data appearing in Table 2 was obtained using the HW-100 transceiver. These data show that, except for the two caveats stated below, once the operating load line and resistance R_{OUT} are established at tuning and loading, R_{OUT} remains substantially constant over the entire range of drive and output power encountered during SSB modulation. After setting the drive level for the π-network to deliver maximum available power of 100 W, and leaving all adjustments except for the level of drive undisturbed thereafter, R_{OUT} was measured at power levels decreasing from 100 W to 12.5 W. (The changes in output power were obtained by changing the drive level.) This range of power levels, as shown in the Table 2, corresponds to approximately the same range of output power prevailing during SSB modulation.

However, the two caveats are necessary to explain the changes in R_{OUT} that appear to conflict with the statement above that R_{OUT} is substantially constant. First, and most important, due to imperfect regulation of plate voltage E_P, the increase in E_P as the plate current and output power decrease, causes a small change in the slope of the load line, resulting in an increase in R_{OUT} that would not occur with perfect regulation of E_P. Second, the scatter in the R_{OUT} data results from the measurements being taken prior to the improved setup and method of taking later measurements that yields the more precise data.

As shown in Table 2, the maximum value of R_{OUT} is 80 Ω, which appears at the minimum output-power level. The conjugate mismatch between the 80 Ω output source resistance and the 51-Ω load resistance is 1.569, establishing a voltage-reflection coefficient $\rho = 0.2214$, a power-reflection coefficient $\rho^2 = 0.0490$, and thus a transmission coefficient $(1 - \rho^2) = 0.951$, which translates to a conjugate mismatch loss of 0.218 dB. This small amount of loss is insignificant when considering that the purpose of the experiment was to establish proof that a conjugate match exists throughout the range of output power during SSB modulation.

However, the picture is even more optimistic when using the same measurement procedure with the Kenwood TS-830S transceiver. It can be seen from Table 3 and Fig 1 below that the variation in output resistance with this transceiver is much less than with the HW-100 over the entire range of drive and output power. The reason for the nearly constant output resistance with the TS-830S is better plate-voltage regulation of the power supply. Especially notice that, in the region of constant output resistance of the network with changes in power level from 102.3 to approximately 25 W, the constant output resistance indicates a constant slope of the load line with changes in drive and power level. This point has

Table 3—Measuring Output Resistance of RF Amplifier of TS-830S Transceiver at 4.0 MHz at Various Power Levels Determined by Drive Level with All Other Adjustments Undisturbed

$R_{OUT} = \Delta E / \Delta I$ See Fig 1 and Table 2

Nominal Power Out (W)	R_{LOAD} *(Ω)*	E_{LOAD} *(V)*	I_{LOAD} *(A)*	R_{OUT} *(Ω)*	*Measured Power Out (W)*	E_P *(V)*	I_P *(mA)*	*Measured Power In (W)*
100	51.2	72.381	1.414	50.2	102.3	790	265	209
	43.5	66.548	1.530		101.8			
		ΔE = 5.833	ΔI = 0.161					
75	51.2	62.738	1.225	49.0	76.9	790	240	190
	45.3	57.738	1.327		76.6			
		ΔE = 5.000	ΔI = 0.102					
50	51.2	50.238	0.981	50.2	49.3	810	195	158
	43.5	46.190	1.062		49.1			
		ΔE = 4.048	ΔI = 0.081					
25	51.2	35.357	0.691	50.5	24.4	810	165	134
	43.5	32.500	0.747		24.3			
		ΔE = 2.857	ΔI = 0.057					
20	51.2	32.857	0.642					
	43.5	30.119	0.692	54.1	21.1	820	135	111
		ΔE = 2.738	ΔI = 0.051					
10	51.2	23.453	0.458	58.6	10.7	825	110	91
	43.5	21.429	0.493		10.6			
		ΔE = 2.024	ΔI = 0.035					

been in dispute in the absence of appropriate measurements. It is also important to observe that at the 10-W output level, where the output resistance has increased somewhat due to imperfect voltage regulation (plate voltage increase with decrease in power), the mismatch between the 51.2-Ω load resistance and the 58.6-Ω output resistance is 58.6 / 51.2 = 1.145:1. This small mismatch results in a voltage-reflection coefficient ρ = 0.0674, a power-reflection coefficient ρ^2 = 0.00454 and thus a transmission coefficient $(1 - \rho^2)$ = 0.9955 (99.55%), which translates to an insignificant conjugate mismatch loss of only 0.0198 dB with minimum speech level referenced to zero loss at maximum speech level.

Except for the lowest levels of speech, which would result in output power less than 10 W during SSB modulation, the data in Tables 2 and 3 and Fig 1 show that the output resistance of the RF amplifier remains sufficiently constant over the normal range of speech levels, ensuring a realistic conjugate match during SSB modulation. Additionally, extrapolation of the data extending the range of output power below 10 W clearly indicates that further increase in output resistance during the lowest practical levels of speech transmission is insignificant relative to load mismatch. Consequently, a realistic conjugate match exists over the entire range of speech levels.

Before concluding this subject, another way of explaining the action occurring in the amplifier during SSB modulation is that as the average integrated energy of the current pulses entering the tank changes with speech modulation, both the RF voltage and current at the output of the tank change linearly in proportion to the modulation level. Consequently, the ratio of output voltage to output current remains constant during speech modulation. Since the output resistance R_{OUT} is determined by the ratio of current to voltage, the output resistance also remains constant and equal to the load resistance during modulation. Thus all the available power from the source is delivered to its load at all levels of speech, satisfying the condition required for a conjugate match at the junction of the source and load during all levels of SSB modulation.

Justifying the IEEE Method for Determining Network Output Resistance

Some authors have claimed that the load variation method cannot determine the correct output resistance when the network is the output load circuit of an RF power amplifier. Their reasoning is that the nonsymmetrical π-network circuit used in the amplifier does not behave like a perfect impedance transformer (perfect integral mathematical input/output ratio). Therefore, they claim even the small change in load resistance used in the IEEE method does not transform linearly through the network, and thus the input impedance of the network does not change directly with changes in load resistance. Their claim is that the imperfect transformer action of the network corrupts the results of the measurements of network output resistance. However, as I will explain using the data from my measurements appearing in Tables 4 and 5, and in Fig 2, it will become evident that that is unfounded.

Recall that in the measurements described above to determine the network output resistance, the change in load resistance is small, around ±10% from the matched load resistance. The change in load must be small to avoid a significant change in the normal operation of the amplifier that would distort the results if the change in load resistance were relatively large. For example, a change in load that would result in a significant change in plate current would change the slope of the load line and the output resistance of the network. On the other hand, for examination and comparison, the data appearing in Table 4 and Fig 2 show the change in network output resistance (and magnitude of impedance) that results from somewhat larger changes in load resistance. Notice that the magnitude of the output impedance changes linearly, but in indirect proportion to the load resistance. However, as seen in Fig 2, the output magnitude remains close to the load resistance when the load resistances are close to the value at which the network was adjusted for delivery of all available power. So the questions are: "Why are the changes in output impedance indirectly proportional to the load resistance?" and "Why does the measured output resistance equal the load resistance when the change in load resistance is small?" The answer is in the resulting change in plate current with change in load, which directly affects the slope of the load line and network output resistance. As we will see, when the change in load is sufficiently small, the change in plate current is also so small that the resulting change in amplifier operation is insignificant in relation to causing error in the measurement of the network output resistance. So let's examine the pertinent changes.

With the matched load (53.1 Ω) the plate current was 260 mA; 280 mA with the 27.7-Ω minimum load resistance, (network output impedance Z = 76.7 Ω) and 210 mA with the 85.7-Ω maximum load resistance (network output imped-

Table 4—π-network Output Impedance Magnitude versus Load Resistance Kenwood TS-830S at Approximately 100 W Output, 4.0 MHz

R_{LOAD} (Ω)	Load Mismatch	E_{LOAD} (V)	I_{LOAD} (A)	Z_{OUT} (Ω)	Sqr Root $R_{LOAD} \times Z_0$	Power Delivered (W)
53.1		71.0	1.34	76.7		94.9
27.7	1.92:1	46.0	1.66		46.1	76.4
53.1		73.0	1.37	67.6		100.4
37.3	1.42	59.0	1.58		50.2	93.3
53.1		72.0	1.36	56.1		97.6
43.9	1.21	65.0	1.48		49.6	96.2
53.1		75.0	1.41	52.7		105.9
49.0	1.08	72.0	1.47		50.8	105.8
53.1		74.0	1.39	41.35		103.1
61.1	1.15	78.5	1.28		50.25	100.9
53 1		73.0	1.37	39.0		100.4
63.0	1.19	78.2	1.24		49.57	97.1
53.1		74.0	1.39	31.95		103.1
66.6	1.25	80.1	1.20		46.13	96.3
53.1		74.0	1.39	29.1		103.1
73.3	1.38	82.0	1.12		46.13	91.7
53.1		74.5	1.40	22.5		104.5
85.7	1.61	84.0	0.980		43.95	82.3
				Average 48.08		

ance Z = 22.5 Ω). However, when the load was changed from 53.1 to 49.0 Ω to obtain Z_{OUT} = 52.7 Ω (see Table 4), any change in the 260 mA of plate current resulting from this small change in load was too small to be discernible on the 0 – 300 mA meter.

Observing from Fig 2, it is interesting to note that in the range from the load resistance of 37.3 Ω (mismatch 1.42) to 63 Ω (mismatch 1.19), the square root of the product of the network output impedance and load resistance remains close to the value of the load resistance that allows delivery of all the available power (≈50 Ω), but begins to decrease with increased load mismatch on either side of the matched condition. It will be shown later that the reactance introduced by the nonsymmetrical network while transforming the nonreactive load resistances to the network input are responsible for the decrease as the mismatch increases in either direction away from the resonant condition.

Let's turn now to Table 5, where my measured data is shown to agree with the statement that impedance transformation through a nonsymmetrical network is not the same as that through a perfect transformer having a linear ratio of transformation. Although the statement on this point is shown true, we will see that the reasoning that some claim renders the IEEE procedure invalid is not correct. Notice that except for the 50-Ω load resistance that matches the network output, all other purely resistive load resistances transform through the network to obtain reactive impedances at the input. When plotted on a Smith chart normalized to 1350 Ω, the loci of these input impedances form a straight line at an angle of 64° clockwise from the resistive axis, intersecting the axis at the chart center where the impedance is 1350 + j0. Still in Table 5, notice that when the load mismatch is approximately 2:1 on either side of the matched point, the angles of the impedances are numerically equal with respect to that at resonance: 32° with the high load resistance and –32° with the low load resistance. However, also notice that with the high and low load resistances, the magnitudes of the input impedances are 1060 and 1780 Ω, respectively. The differences between the matched input impedance (nonreactive) of 1350 Ω are thus 290 and 430 Ω, respectively. Let's now examine the significance of these *measured* input impedances.

First, the 1780-Ω input impedance obtained with the low-resistance load is 140 Ω further off the resonant value than the 1060 Ω with the high-resistance load, resulting in the higher off-resonance plate current with the low-resistance load than for the high-resistance load with same degree of mismatch (1.92:1 and 2.0:1, respectively).

Second, the measured input impedances resulting from two load resistances of ±5 Ω relative to 52 Ω were interpolated on the straight-line locus of all the impedances in Table 5. At the resulting two equal input mismatches of 1.12:1, the resulting input impedances normalized to 1350 Ω are 0.95 + j0.10 and 1.10 –j0.10, respectively, for real values of 1282.5 + j135 and 1485 –j135 Ω, respectively. With these mismatches of only 1.12:1 on either side of resonance, the angle of the mismatched impedances is only ±6° and the change from normal amplifier-system performance of 1.0 only to 0.9968, (99.68%) amounts to a change of only 0.014 dB.

Consequently, the small change in amplifier performance under these conditions is insignificant with respect to contributing to error while using the standard IEEE load-variation method in the measurement of network output impedance. Ergo, criticism of the IEEE method of measuring the output impedance of π-networks in RF power amplifier operation is unfounded and the method is proved valid.

In conclusion, my measurements and discussion prove that the output of a properly designed RF tank circuit

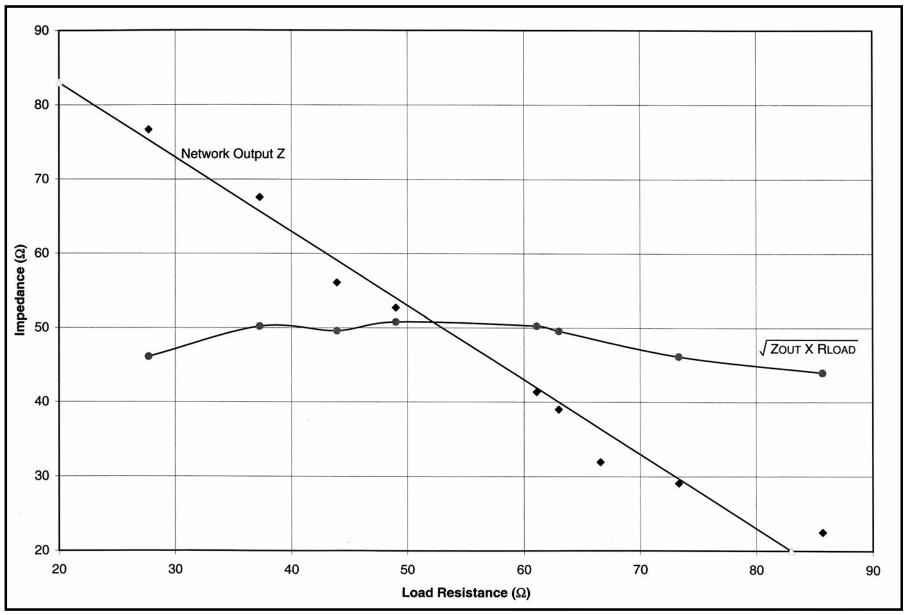

Fig 2—Kenwood TS-830S RF amplifier π-network output impedance, *Z*, versus load resistance. All available power of approximately 100 W is delivered initially to 53.1-Ω load. All adjustments left undisturbed during measurement of other load values (see Table 4).

Table 5—π-Network Input Impedance versus Nonreactive Load Resistance

HW-100 Transceiver Initially Resonated with 52.0-Ω Load at 100 W Output

Network Load (Ω)	*Mismatch Re 52.0 (Ω)*	*Network Input Z Polar*	*Network Input Z Rectangular*
240.0	4.6:1	950@58°	503.4 + j805.6
160.0	3.17	980@48°	655.7 + j728.3
100.0	1.92	1060@32°	898.9 + j561.7
83.0	1.6	1150@20°	1080 + j393.3
52.0	1.0	1350@0°	1350 + j0
41.0	1.22	1580@–14°	1533 –j382.2
34.2	1.52	1630@–18°	1550 –j503.7
26.0	2.0	1780@–32°	1509 –j943.3
20.6	2.52	1900@–41°	1433 –j1245
17.5	2.97	2000@–48°	1338 –j1486

performs with a linear relationship between output voltage and current. Consequently, the theorems associated with linear circuits, such as the Conjugate-Matching Theorem and the Maximum-Power Transfer Theorem, are valid and applicable in both the analysis and practical operation of RF power amplifiers performing in both CW and SSB modes.

Several professional RF engineers and university EE professors have reviewed the material in this article for accuracy, all having pronounced the material to be an accurate and true presentation of the subject. They are Warren Amfahr, W0WL; Jan Hornbeck, N0CS; and others. In addition are John C. Fakan, PhD (KB8MU), private consultant, formerly a consultant to NASA; John S. Belrose, PhD in Radio Science, Cantab (Cambridge, UK) (VE2CV), Senior Radio Scientist, Communications Research Center, Canada; Forrest Gehrke, BSEE (K2BT), microwave engineer with the former RCA; Al Helfrick, PhD in Applied Sciences, (K2BLA) Professor of Avionics at Embry-Riddle Aeronautical University; Robert P. Haviland, EE, (W4MB) retired General Electric RF engineer.

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Notes

[1]The use of "linear" and "nonlinear" relates to the voltage/current relationship at the output terminals of the amplifier tank circuit or at the terminals of a network. This usage does not relate to nonlinearity between the input and output of an amplifier that results in generation of distortion products in the output signal.

[2]In addition to the data in Terman's example, I have made measurements that determine the output impedance of RF power amplifiers, which prove the existence of a conjugate match. The data show that when the amplifier is loaded to deliver all of its available power, the output impedance of the amplifier equals the load impedance, thus signifying a linear voltage/current relationship at the output of the tank circuit and thus a conjugate match. The description of my measurement procedure and the resulting data showing the proofs are included here in following sections.

[3]The material discussed in this article pertains only to RF amplifiers used in the Amateur Service with vacuum tubes and π-networks in the output circuit. The material does not necessarily pertain to amplifiers used in various commercial services, or any amplifier using solid-state components.

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Reducing IMD in High-Level Mixers

If the question is "How can I get the greatest dynamic range out of diode mixers?" a high-level mixer is often the answer. Come learn how to get great performance out of these useful devices.

By John B. Stephensen, KD6OZH

About four months before this writing, I started designing a new HF transceiver. I live in Los Angeles, where there are many other active hams and several within a one-mile radius. I wanted a good strong receiver that would be immune to interference from strong local signals on the same band where I was listening for weak signals. After looking at several *QST* and *QEX* articles, I decided that a passive high-level mixer was the best solution.

All went well until I finished construction and tested the receiver. The measured intermodulation distortion (IMD) was not anywhere near the performance expected. The actual third-order intercept (IP3) was 15 dB lower than could be achieved according to the manufacturer's published data.

My design followed the guidelines published in Amateur Radio publications, but there must have been something that I didn't know. Thus, I started performing a series of tests with different terminations on the mixer's LO, RF and IF ports to see what would solve the problem. What I found can help you use those types of mixers more effectively.

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The Mixer

The Mini-Circuits ZFY-1 doubly balanced mixer (DBM)[1] used in the transceiver has impressive specifications and is usable from the 160-meter to 70-centimeter amateur bands (see Table 1). Other models are available for applications through 3 GHz.

Mini-Circuits application notes indicate that the third-order intercept of a DBM is approximately 15 dB higher than the 1-dB compression point. This mixer should be capable of a third-order intercept of +35 dBm (3 W!) if used properly.

Manufacturers test DBMs with broadband 50-Ω terminations at all ports so these numbers do not show exactly what will happen when the mixer is embedded in a transceiver. However, they are reproducible and do give a good indication of relative performance. With the proper design, performance close to these numbers should be possible.

Testing the Conventional Wisdom

I initially assumed that termination

[1]Notes appear on page 50.

Table 1—ZFY-1 Specifications

LO/RF	100 kHz - 500 MHz
IF	10 kHz - 500 MHz
LO Power	23 dBm
RF 1-dB Compression Point	20 dBm
Conversion Loss, 0.2-250 MHz	4.9 dB typ, 6.0 dB max
Conversion Loss, 0.1-500 MHz	7.5 dB max
LO-RF Isolation, 0.1-1 MHz	20 dB min, 40 dB typ
LO-RF Isolation, 1-250 MHz	35 dB min, 46 dB typ
LO-RF Isolation, 250-500 MHz	30 dB min, 40 dB typ
LO-IF Isolation, 0.1-1 MHz	23 dB min, 37 dB typ
LO-IF Isolation, 1-250 MHz	35 dB min, 46 dB typ
LO-IF Isolation, 250-500 MHz	30 dB min, 40 dB typ

of the mixer's IF port was the only important requirement for using a DBM. Competent advice told me that the IF port must be terminated in a 50-Ω broadband load, but other ports were not very sensitive to their terminations. The following passage from *Solid State Design for the Radio Amateur*[2] is similar to those in many other articles in amateur literature:

"After initial measurements were performed with broadband terminations at all ports, tuned circuits were inserted in various lines to the mixer. These were single-tuned LC circuits. The results were profound! When a single-tuned circuit was put in the IF port it had the effect of still presenting a 50-Ω termination at the desired IF of 9 MHz. (The RF energy was at 14 MHz and the LO was at 23 MHz.) However, at frequencies other than the 9-MHz IF, the impedance was highly reactive. This has the effect of decreasing the output intercept from +15 dBm to +5 dBm in several of the mixers studied. The conversion loss did not change significantly."

"When a narrow-band termination was used at the RF and LO ports of the mixer, a degradation in output intercept was also observed. However, it was not nearly as severe as that seen at the IF port."

This testing was done on a low-level mixer with a +10-dBm LO. I was using a high-level mixer with a +23-dBm LO. The LO and IF frequencies were identical. Perhaps there was a difference in behavior of high-level mixers.

The circuit used to test this assumption is shown in Fig 1. A crystal-controlled oscillator and an HP-8640 signal generator were used to generate RF signals at 14.318 MHz and 14.218 MHz, respectively, which were combined in a hybrid combiner and sent to the mixer under test. The output level was 4 dBm per tone. This is a typical setup for two-tone IMD testing. For more information on IMD, see *The ARRL Handbook*.[3] The LO frequency range used was 23 to 23.4 MHz and the IF was 9 MHz. The LO was tuned to place each distortion product into the IF filter passband for measurement.

The two tones pass through a three-pole RF band-pass filter connected to the RF port of the mixer. The band-pass filter has a loss of 3 dB. The mixer's LO port was driven directly by an amplifier with an untuned output. The IF port was terminated with an attenuator that fed directly into a KVG XF-9B10 2.4-kHz bandwidth crystal filter, matched to 50 Ω by an L network. This was followed by an IF strip and the rest of the receiver. The AGC system is this receiver has a digital readout accurate to ±1 dB over the range used. The filter exhibits a very low return loss (high SWR) outside its passband. Consequently, return loss outside the passband was determined by the loss in the attenuator as shown in Table 2.

The attenuator was adjusted to determine the mixer sensitivity to IF-port termination impedance and the results are shown in Table 3. This was hardly the improvement expected from the conventional wisdom on DBMs. Only 4 dB of improvement was seen versus the 10 dB expected.

When attenuators were inserted at the RF port in place of the band-pass filter, considerable performance improvement was measured, as shown in Table 4. These results showed that, at least in high-level mixers, the IF port is not the only port sensitive to its termination. An 11-dB improvement could be obtained by paying attention to the RF-port termination.

A test with a 3-dB attenuator on the LO port showed no discernable difference in IMD; however, LO waveform symmetry does affect IMD performance, and the LO voltage at the mixer was asymmetrical. I had not seen any quantitative measurements published so I added a filter to the LO port and made the measurements shown in Table 5. The filter was a π network tuned to 23 MHz with a Q of 3. With this filter in place, waveform symmetry was 50/50 and IMD dropped by 6 dB.

These measurements show that both the IF and RF DBM ports must be well matched and the LO port must be supplied with a clean signal to achieve the best performance. The next problem was how to translate these requirements into a real circuit inside the transceiver.

Achieving Proper Termination at the RF Port

The RF and LO frequencies are the most important and must be terminated properly. When the IF frequency is 1.5 times the highest R frequency, or higher, proper termination can be provided by parallel low- and high-pass filters with the high-pass or "idler" filter terminated in 50 Ω as shown in Fig 2. Doug Smith, KF6DX, referred to this in a recent *QEX* article[4] describing a transceiver with a level-17 mixer and a 75-MHz IF. In most cases, a set of band-pass

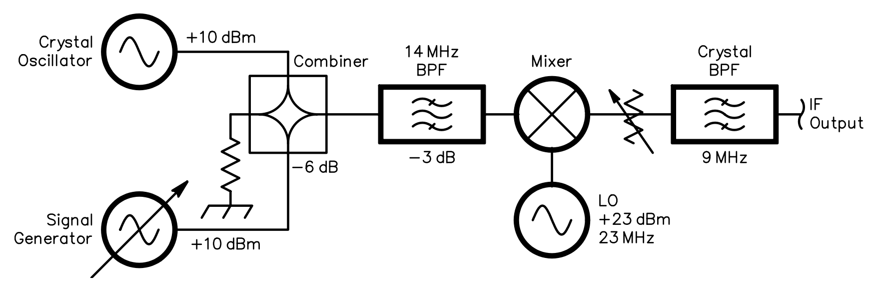

Fig 1—IMD Test Setup

Table 2 —Attenuation, Return Loss and SWR

Attention	*Return Loss*	*SWR*
3 dB	6 dB	3:1
6 dB	12 dB	1.67:1
9 dB	18 dB	1.29:1

Table 3—Receiver IMD versus IF-Port Match

IF Port Attenuation	*IMD*	*IP3*
0 dB	–33 dBc	+20.5 dBm
–3 dB	–37 dBc	+22.5 dBm
–6 dB	–40 dBc	+24.0 dBm
–9 dB	–41 dBc	+24.5 dBm

Table 4—Receiver IMD versus RF-Port Match

RF Port Attenuation	*IMD*	*IP3*
(BPF)	–37 dBc	+22.5 dBm
–3 dB	–52 dBc	+30.0 dBm
–6 dB	–65 dBc	+33.5 dBm

Table 5 —Receiver IMD versus LO-Waveform Symmetry

LO Filter	*IMD*	*IP3*
no	–37 dBc	+22.5 dBm
yes	–49 dBc	+28.5 dBm

filters that are one-half to one-octave wide precede the low-pass filter to prevent severe second-order intermodulation products. Since LO harmonics fall within the high-pass filter's passband, they are terminated properly. Note that any IMD products from signals inside the band-pass filter's passband that fall outside both the band-pass filter's and the high-pass filter's passbands are not terminated.[5] This effect is probably insignificant compared to other sources of IMD.

When the IF is lower than the RF (as with VHF/UHF transceivers or HF transceivers having a first IF in the 3- to 11-MHz range), the mixer RF port requires a band-pass filter to reject the image frequency. This filter is also desirable to reject strong signals that are present from international broadcast stations or nearby amateur transmitters on adjacent bands.

RF band-pass filters can have either very high impedance or very low impedance outside the passband depending on the type of filter. Filters with series capacitors or inductors at their inputs for matching to the first resonator will have high impedances outside the passband, as shown in Fig 3. Filters that have tapped inductors or shunt capacitors on their inputs will have low impedances outside the passband, as shown in Fig 4. Either situation can cause excess IMD in the mixer. Unfortunately, most amateur-built equipment ignores this problem.

To provide a proper termination, a diplexer circuit can be used between the filter and mixer, as is commonly done at the IF port. This achieves a broadband 50-Ω match at one or both ports. However, achieving such a broadband match is only practical for simple filters with a Q of 10 or less. When a multipole filter is needed to achieve steep skirts and reject the RF image frequency, a different approach must be taken. The RF and LO frequencies are the most important and the mixer termination need not be perfect far away from these frequencies. This fact can be used to design a simpler and lower-loss solution for the RF port of the mixer.

If the filter presents low impedance outside the passband, place a 50-Ω resistor in series with the filter input. Alternatively, a 50-Ω resistor may be placed in parallel with the filter if it presents a high impedance outside the passband. This gives a good match far from the passband. The resistor can then be bypassed by a series-resonant

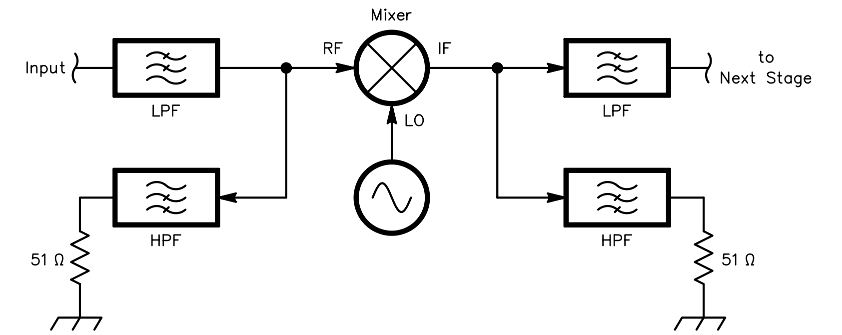

Fig 2—Parallel high- and low-pass filters for LO termination at RF port.

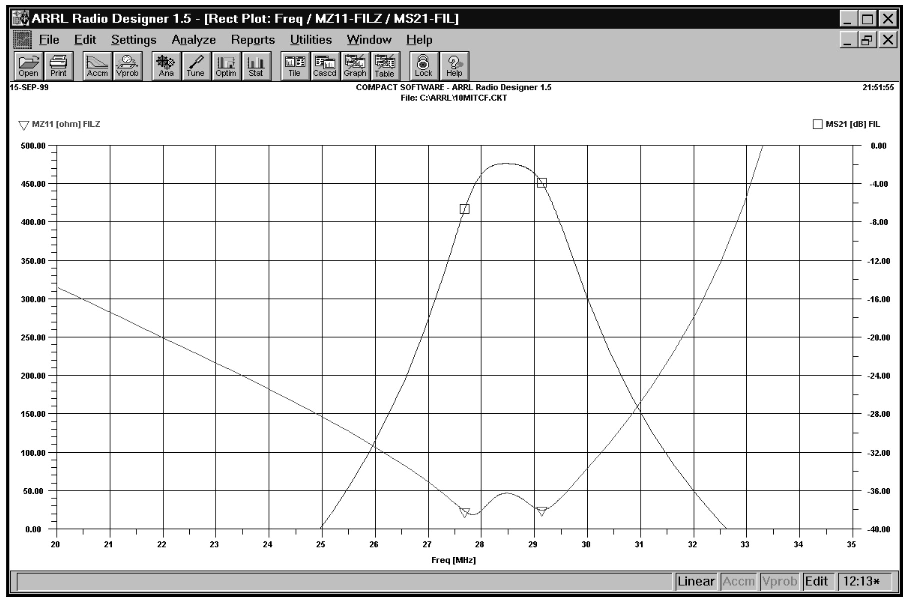

Fig 3—Impedance (MZ11) and attenuation (MS21) plots of a three-pole 10-meter band-pass filter with series-capacitor matching at the input port.

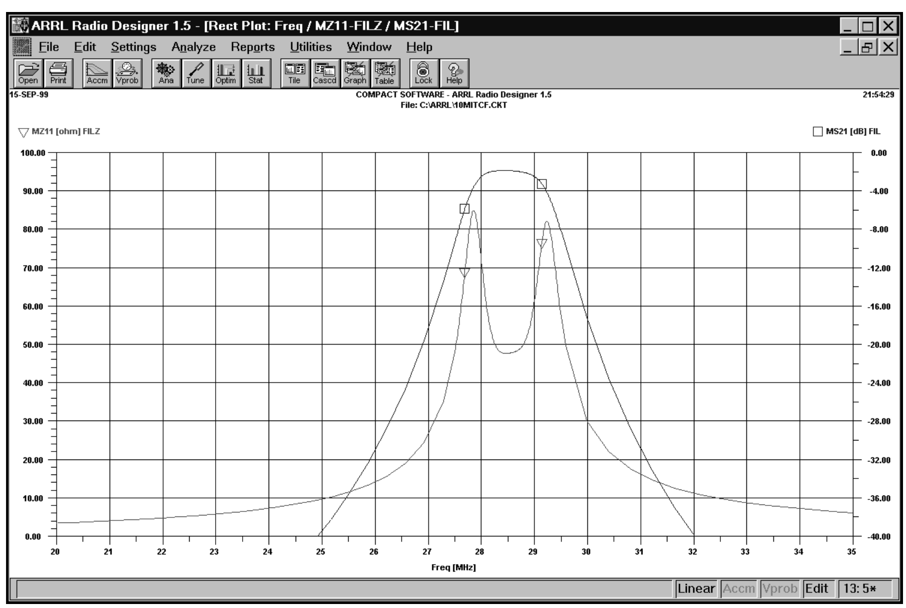

Fig 4—Impedance (MZ11) and attenuation (MS21) plots of a three-pole 10-meter band-pass filter with tapped-inductor matching at the input port.

trap or disconnected by a parallel-resonant trap for signals inside the passband, as shown in Fig 5. The trap need not have high Q, as it must only have a bandwidth smaller than the difference between the RF and LO frequencies. It will affect the adjacent resonator slightly, but this may be retuned and no significant distortion of the passband will result.

A band-pass filter, shown in Fig 6, was designed for the 20-meter amateur band and tested. Figs 7 and 8 show the measured attenuation and return loss. A Chebyshev filter with 0.1-dB ripple and shunt capacitive coupling between resonators was chosen because of its low-pass characteristic. This provides maximum attenuation at the image frequency with a high-side LO. The resonator impedance was chosen to maximize the unloaded Q of the inductor for lowest filter loss.

The filter attenuation is 90 dB at the LO frequency (23 MHz) and over 100 dB at the image frequency (32 MHz). This was achieved with the help of shielding between resonators. The series-resonant trap is constructed with a fixed-value inductor and capacitor. The inductor is a molded RF choke with a Q of no more than 50. However, the return loss is 23 dB or more at both the LO and RF and insertion loss at 14.2 MHz is only 2.8 dB. The return loss is optimized above the filter passband. This is an advantage when the LO is above the RF. A transformer-coupled design showed better return loss below the filter passband.

Achieving Proper Termination of the IF Port

With low-level mixers, a post-mixer IF amplifier is usually included to provide a broadband termination. When using a high-level mixer in a receiver, the output intercept of the mixer is often equal to or higher than the input intercept of any low-noise amplifier that can be designed to follow it. Any amplifier that is used will also promote IMD in the crystal filter that follows. For example, the ZFY-1 mixer can produce 25 mW of output, which exceeds the 10-mW maximum input rating of most crystal filters without any amplification. It is best to avoid any IF amplifier prior to the first crystal filter. However, crystal filters present a very reactive load outside their very narrow (250 Hz to 15 kHz) passbands.

The IF port may be terminated using the same strategy as the RF port. A low-Q diplexer will terminate any LO energy at the IF port and eliminate much of the IMD problem. Fig 9 shows such a circuit designed for 9 MHz and Fig 10 shows its characteristics. The low Q (1.6) allows the use of fixed capacitors and inductors and ensures a low insertion loss (0.3 dB) at the IF, while providing a high return loss at the 23-MHz LO frequency. The main crystal filter or a roofing filter can then be connected to the diplexer output. There will be some IMD generated by strong signals inside the diplexer passband and outside the

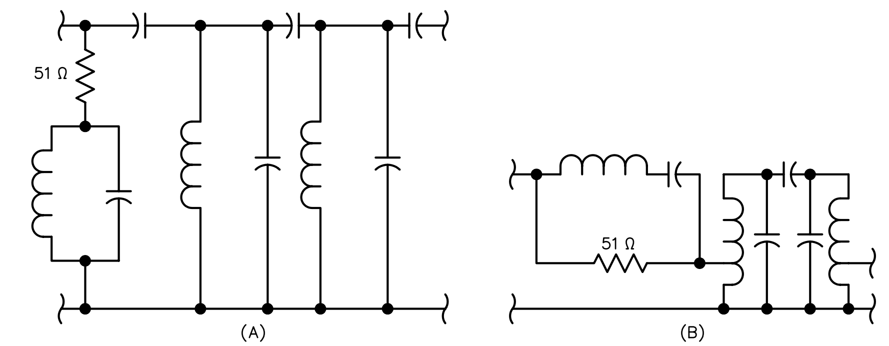

Fig 5—Multipole band-pass filters with out-of-band impedance matching on one port. The filter at (A) has a high impedance outside the passband and the filter at (B) has a low impedance outside the passband.

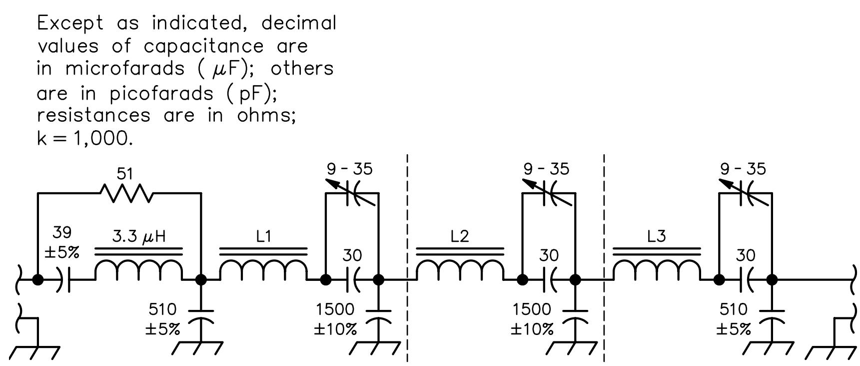

Fig 6—20-meter band-pass filter schematic. L1, L2 and L3 are each 24 turns of #22 AWG enameled wire on a T50-6 powdered-iron toroid core (2.5 µH).

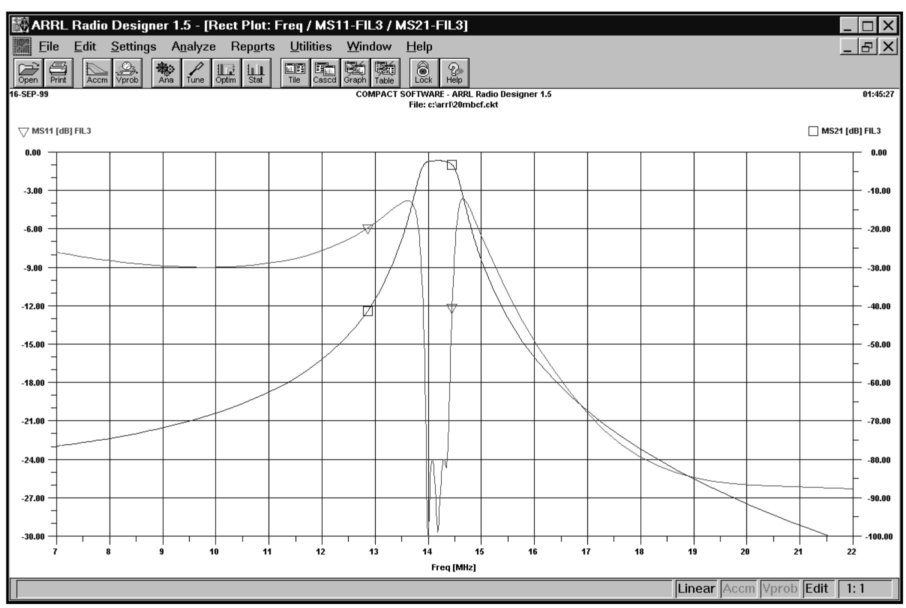

Fig 7—Predicted attenuation and return loss for the 20-meter band-pass filter.

crystal filter passband, but this effect turns out to be minimal compared to other sources of IMD.

Driving the LO Port

It has long been known that it is very important for the LO waveform to be as symmetrical as possible to minimize IMD. In this case, the original LO signal was asymmetrical with a 47/53% duty cycle with respect to zero crossings. Since the LO tunes only a relatively narrow range (the 20-meter amateur band), a π network is suitable to filter the input to the LO port. The network shown in Fig 11 has a Q of 2.1 and attenuates the second harmonic of the LO by more than 20 dB. Fixed-value capacitors and inductors were used for construction; loss is minimal.

Results

Tests were made using a ZFY-1 mixer and two crystal-controlled RF sources with a 25-kHz separation. Crystal-controlled sources were used to ensure that phase noise did not affect measurements at such a close spacing. The two-tone RF signal generator circuit is very similar to one previously published in *QEX*,[6] but uses the oscillator and filter circuit described in my ATR-2000 article.[7] The signal levels at the RF filter were +1 and +2 dBm at approximately 14.300 MHz and 14.325 MHz. The test setup is shown in Fig 12.

The LO filter, terminated RF filter, IF diplexer and IF attenuator described above were all constructed using connectors so that they could be removed or reversed to measure their effects. Various circuit configurations were tried and the results are shown in Table 6.

The results are impressive with all modifications in place (test 1). The third-order intercept point (IP3) is within 1 dB of that achievable according to the manufacturer's application notes. It is very interesting considering that the return loss at the IF port for the down-converted test signals is 6 dB (3:1 SWR). In fact, removing the 3-dB attenuator at the IF port only degraded the IP3 by 3.5 dB. Going beyond 3-6 dB of attenuation had no measurable effect. Clearly, termination of the signals near the IF has a smaller effect on IP3 than any other factor.

When tests were repeated with a second mixer (tests 9 and 10), insertion of the 3-dB attenuator at the IF port made only a 0.5-dB difference. The IP3 with the 3-dB IF attenuator was 1.7 dB lower and, without the attenuator, was 1.3 dB higher than the first mixer. This is important on the higher HF bands because the IF filter can be connected directly to the IF diplexer to lower the receiver noise figure.[8] An RF amplifier would increase complexity and cause

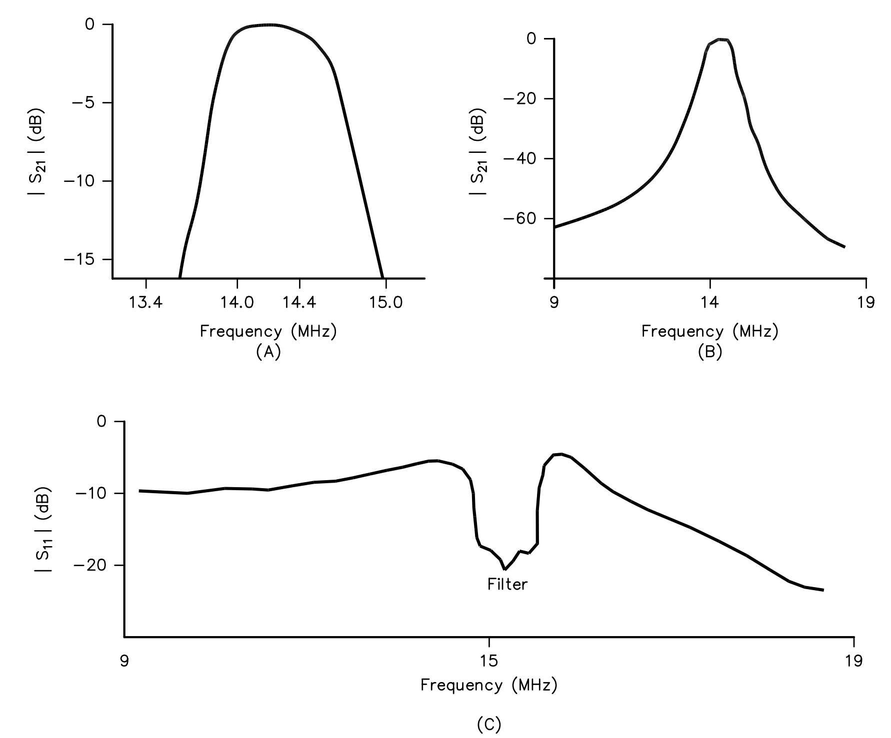

Fig 8—20-meter band-pass filter showing measured attenuation (at A and B, S21) and return loss (at C, S11).

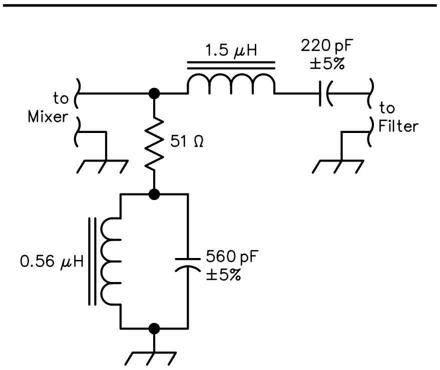

Fig 9—IF diplexer circuit.

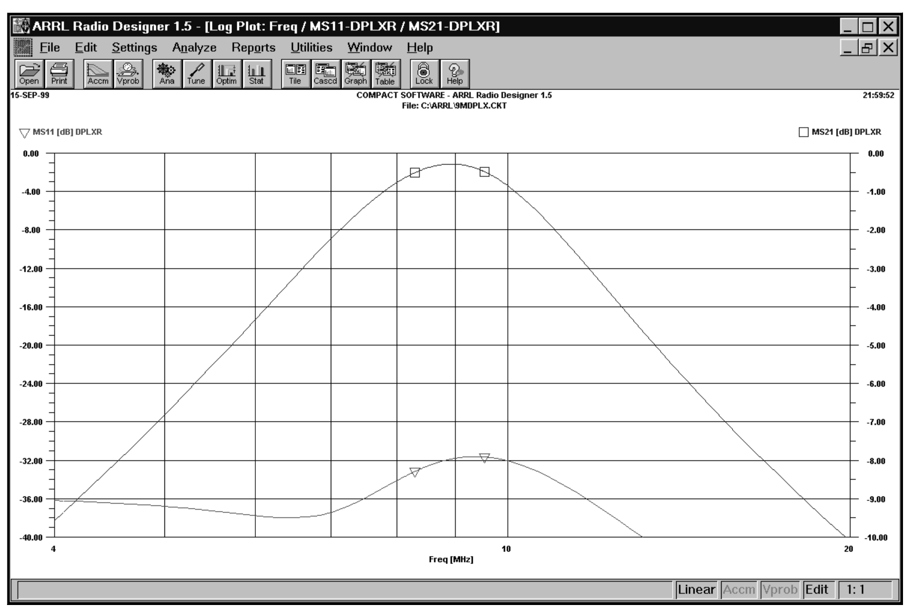

Fig 10—IF diplexer attenuation and return loss.

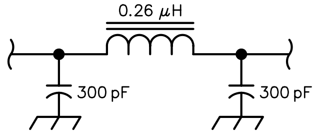

Fig 11—LO-port filter.

greater degradation in IP3 than was shown with either mixer.

The largest single improvement in IP3 is caused by terminating the LO and image frequencies at the mixer's RF port (test 7). Termination of the LO frequency at the IF port and ensuring symmetry of the LO waveform are also large contributors. Interestingly, the IF diplexer has little effect if the RF port is not terminated properly at the LO frequency (test 8). Proper termination of all ports is required for the best performance.

Table 6—IMD Test Results with Various Configurations

Test #	*LO Filter*	*RF Filter Terminated*	*IF Diplexer*	*IF Attn*	*IP3*	*Δ IP3*
1	Yes	Yes	Yes	3 dB	37.0 dBm	0
2	Yes	Yes	Yes	0 dB	33.5 dBm	–3.5 dB
3	Yes	Yes	No	3 dB	30.0 dBm	–7.0 dB
4	Yes	Yes	No	0 dB	24.5 dBm	–11.5 dB
5	No	Yes	Yes	3 dB	30.0 dBm	–7.0 dB
6	No	Yes	No	3 dB	23.5 dBm	–13.5 dB
7	Yes	No	Yes	3 dB	28.0 dBm	–9.0 dB
8	Yes	No	No	3 dB	27.5 dBm	–9.5 dB
9*	Yes	Yes	Yes	3 dB	35.3 dBm	0
10*	Yes	Yes	Yes	0 dB	34.8 dBm	–0.5 dB

*Test with ZFY-1 option B mixer date code 9448 02—others with date code 9602 03

Notes

[1]The Mini-Circuits ZFY-1 is the "connectorized" version of the SAY-1 mixer and both use 200 mW of LO injection to achieve high performance levels. The ZFY-1 is $75 from Mini-Circuits distributors but I have also seen it for $30 on the surplus market. A lower-cost mixer with specifications almost as good is the TUF-1H, which uses a 50-mW LO and is available for $12 new. Surplus Sales of Nebraska, RF Parts and Down East Microwave are good sources of these or similar high-level mixers at low cost.

[2]W. Hayward, W7ZOI, and D. DeMaw, W1FB, *Solid State Design for the Radio Amateur* (Newington: ARRL, 1977), p 119.

[3]R. Straw, N6BV, Editor, *The ARRL Handbook for Radio Amateurs* (Newington: ARRL, 1999), Chapter 15.

[4]D. Smith, KF6DX, "Signals, Samples, and Stuff: A DSP Tutorial (Part 2)," *QEX*, May/June 1998, pp 22-37.

[5]The original signals inside the band-pass filter passband can produce odd-order distortion products outside the band-pass filter passband. For example, if the band-pass filter covers 5 to 7 MHz and the original signals are 5.5 and 6.5 MHz, the mixer will produce third-order distortion products at 4.5 and 7.5 MHz, fifth-order distortion products at 3.5 and 8.5 MHz and so on. The mixer RF port is not terminated in 50 Ω at these frequencies.

[6]S. Rumley, KI6QP, "A Precision Two-Tone RF Generator for IMD Measurements," *QEX*, April 1995, pp 6-12.

[7]J. Stephensen, KD6OZH, "The ATR-2000: A High Performance Homemade Transceiver, Pt 1" *QEX*, Mar/Apr 2000, pp 3-15. The crystals shown in Appendix A were pulled up or down in frequency with series connected 12-pF or 10-µH components to achieve a 25.8-kHz spacing between oscillator frequencies. (Part 2 of the ATR series appeared in May/June 2000 and Part 3 in Mar/Apr 2001.)

[8]Atmospheric noise below 30 MHz is nearly always more than 18 dB above the thermal noise, so a receiver noise figure of 13 dB is adequate for all HF bands.

John Stephensen, KD6OZH, has been interested in radio communications since building a crystal radio kit at age 11. He went on to study Electronic Engineering at the University of California and has worked in the computer industry for 26 years. He was a cofounder of Polymorphic Systems, a PC manufacturer, in 1975 and a cofounder of Retix, a communications-software and hardware manufacturer, in 1986. Most recently, he was Vice President of Technology at ISOCOR, which develops messaging and directory software for commercial users and ISPs. John received his Amateur Radio license in 1993 and has been active on the amateur bands from 28 MHz through 24 GHz. His interests include designing and building Amateur Radio gear, digital and analog amateur satellites, VHF and microwave contesting and 10-meter DX. His home station is almost entirely home-built and supports operation on SSB, PSK31, RTTY and analog and digital satellites in the 28, 50, 144, 222, 420, 1240, 2300, 5650 and 10,000 MHz bands from Grid Square DM04 in Los Angeles. The mobile station includes 10-meter SSB, 144/440-MHz FM and 24-GHz SSB.

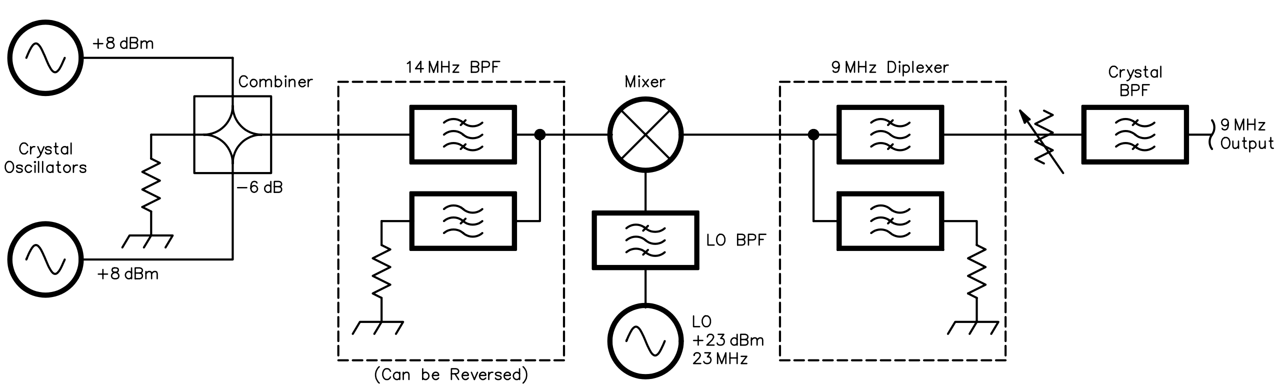

Fig 12—IMD test setup for data of Table 6.

A Perl /Tk Package for NEC-Based Antenna Design

Here is some antenna-modeling software for Linux/Unix users. You no longer need Windows to see your antenna pattern.

By Bill Walker, W5GFE

There are many useful programs available for modeling antenna systems. Among the most popular are those developed by Roy Lewallen, W7EL.[1] Roy's programs are intended for use on machines running Microsoft operating systems.

Other good sources of information include "A Beginner's Guide to Using Computer Antenna Modeling Programs,"[2] from L. B. Cebik, W4RNL, and "Modeling HF Antennas with MINI*NEC*—Guidelines and Tips from a Code User's Notebook,"[3] by John Belrose, VE2CV.

[1]Notes appear on page 54.

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Enter *NEC*

If, like me, you're a user of Unix-based operating systems, you may find the program *NEC* more to your interest.[4] There's very useful downloadable documentation for *NEC* on the Internet.[5]

The *NEC* program is very versatile. It consists of about 9000 lines or so of *FORTRAN* (see Note 4 for a *C* version). To compile the data, you'll need a good *FORTRAN* compiler. If you don't have one, check out the GNU project on the Internet.[6]

NEC is a complex program. It produces output that consists of pages and pages of numbers arranged in seemingly never-ending columns. This may have accounted for its lukewarm acceptance in the amateur community over the years.

In an article published on the Internet,[7] I described a system of *Tcl/Tk* programs that accepts the voluminous output of *NEC* and produces a series of graphs which are, to me at least, more useful than the numeric output of *NEC*. I find these graphs especially helpful when I'm just fiddling with an antenna.

Although that software proved both useful and popular, it suffered from several problems—including the fact that the *Tcl/Tk* language needs the BLT extension to handle the programs. Also, I must confess that *Tcl/Tk* itself isn't the most comfortable language for me, and I was never satisfied with the "straight line" code produced when writing in *Tcl/Tk*. The current effort, written in *Perl/Tk*, seems more robust and satisfactory.

Perl/Tk is an extension of the enormously popular *Perl* language (see Notes 8 and 9 for authoritative work). Both *Perl* and the *Tk* extension are available for most operating systems in either binary or source form from the Internet (see CPAN at Note 10). I'm using *Perl/Tk* on SCO *Unix* 5.0.4, but many of my students are using it under either *Linux* or *Windows NT*. One intrepid soul is even running it under *Windows 98*!

Modeling antenna patterns with NEC

To model antenna patterns with *NEC*, you must first acquire (or compile) a binary version of the program. Try the Web page in Note 4 for either source code or binary files. While you're at it, you might also try to print the helpful documentation from the source in Note 5.

NEC consists of a single binary file, so installation is easy. Just locate it in a convenient place in your search path. *NEC* does produce some intermediate files, so you need to run *NEC* in a directory where you have write permission. The input to *NEC* consists of a series of lines that describe the geometry of an antenna and the applied excitation.

Informally, each element of an antenna consists of a straight line. Each element is "tagged" with an integer. The two endpoints of each element represented as points with *X*, *Y* and *Z* coordinates. An element is further divided into segments for calculating currents in each segment. It's also possible to specify the radius of wire in the element. The data lines that describe elements are called "GW" lines.

There's usually a GW line for each element of the antenna; however, an antenna such as a discone can be described by rotating several elements if need be, so that isn't always true. Because many who model antennas may wish to use English units and *NEC* requires metric units, it's possible to include a "GS" data line that applies a scaling factor to all elements of an antenna. This allows the designer to work with any convenient unit of measure.

The end of the physical description of the antenna is marked with a "GE 1" data line. Other data lines ("GN" data lines) are used to describe the ground under an antenna, and the frequency ("FR" data lines) and excitation ("EX" data lines) applied to an antenna. Other lines ("LD") can be used to indicate termination impedances.

If you provide a "RP" data line, the program calculates radiation patterns. Different parameters of the RP line ask for E and H fields to be calculated at various angles of azimuth or elevation. The W5GFE antenna package expects you to provide one of two different RP cards to provide correct input to the plotting programs. The "EN" data line indicates the end of the input data. For a complete discussion of all the possible input lines and parameters, see the *NEC* documentation.

Modeling a Rhombic with NEC

I live on about 40 acres of land in southeastern Oklahoma. There's a single diamond-shaped meadow on the property that could accommodate a rhombic antenna that is 120 feet by 240 feet. If I use telephone poles that can place the wires 50 feet in the air, is it worth building that rhombic for 20 meters? Table 1 is a *NEC* data file that describes the situation. The four GW lines describe the rhombic antenna. The rhombic has four elements. The line

GW 1 10 –120.0 0.0 50.0 0.0 60.0 50.0 0.01

describes the first element, which is divided into 10 segments. The coordinates of the endpoints are (–120.0, 0.0, 50.0), as shown in Fig 1. (The 0.01 represents the radius of the wire in the element.)

The figure makes it easy to determine the other three GW lines. Because the antenna is measured in feet, we use a GS line to scale the entire antenna to metric units. The GE line indicates an end to the antenna geometry, while the GN line describes the ground.

Other lines include the FR line (frequency = 14.0 MHz), two EX lines (the excitation applied to element 1 and element 2) and two LD lines (describe the 600-Ω termination resistors on elements 3 and 4).

The RP Data Line

For purposes of the W5GFE antenna package, it's important to use a special RP data line. The RP line included in Table 1 asks for an elevation profile (angles 0° to 180°) to be computed at a distance of 10 kilometers from the origin. This will instruct *NEC* to provide output the program *Elevation* expects.

When computing azimuth fields, a different RP line is used. If you wish to compute azimuth patterns, use this RP line instead:

RP 0 1 361 1001 75.0 0.0 0.0 1.0 10000.0 0.00E+00

The RP line isn't used if you simply want to view the antenna. Look at it using either *Wires* or *NECview*.

The Programs

The programs in the W5GFE antenna package include: *Azimuth*, *Elevation*, *Wires* and *NECview:*

Azimuth

The *Azimuth* program expects to use output from *NEC* where the RP data line is of the form:
RP 0 1 361 1001 75.0 0.0 1.0 10000.0 0.00E+00

Table 1—A *NEC* data file.

```
CM NEC Input File for W5GFE Rhombic
CE
GW 1 10 –120.0 0.0 50.0 0.0 60.0 50.0 0.01
GW 2 10 –120.0 0.0 50.0 0.0 –60.0 50.0 0.01
GW 3 10 0.0 60.0 50.0 120.0 0.0 50.0 0.01
GW 4 10 0.0 –60.0 50.0 120.0 0.0 50.0 0.01
GS 0 0 0.3048
GE 1
GN 1 0 0 0 0.0 0.0 0.0 0.0 0.0 0.0
FR 0 1 0 0 14.0 0.0 0.0 0.0 0.0 0.0
EX 0 1 1 0 1.0 0.0 0.0 0.0 0.0 0.0
EX 0 2 1 0 –1.0 0.0 0.0 0.0 0.0 0.0
LD 0 3 10 10 600.0 0.0 0.0 0.0 0.0 0.0
LD 0 4 10 10 600.0 0.0 0.0 0.0 0.0 0.0
RP 0 180 1 1001 –90.0 0.0 1.0 0.0 10000.0 0.0
EN
```

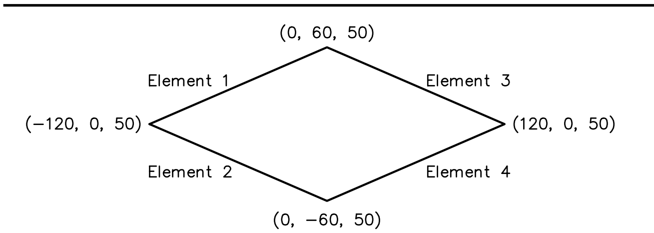

Fig 1—The rhombic antenna.

If the output file from *NEC* were named rhombic.out, the command **Azimuth rhombic.out** would produce four windows on the screen. Two of the windows display polar plots of the azimuth pattern, while a third displays a Cartesian plot of the same information. The fourth window provides a "fact box" with useful information about the antenna. An example of the *Azimuth* screen appears in Fig 2.

Elevation

The *Elevation* program expects to use output from *NEC* where the RP data line is of the form:

RP 0 180 1 1001 -90.0 0.0 1.0 0.0 10000.0 0.00E+00

The command **Elevation rhombic .out** produces four windows similar to those from *Azimuth*, but that contain information about elevation patterns. An example of the *Elevation* screen in shown in Fig 3.

Wires

Unlike *Azimuth* or *Elevation*, *Wires* uses the *NEC* input file, not the file produced as output by *NEC*. *Wires* ignores all data lines except the GW lines. Therefore, *Wires* isn't useful for antennas created using rotations (such as "GH" or "GR" data lines).

The *Wires* program provides three different views (XY plane, YZ plane and XZ plane) of the same antenna. The corresponding elements of the antenna are colored the same in each view. *Wires* has a file-selection window that lets you pick the antenna description you wish to view.

Wires provides a "static" view of the antenna. Since the arrival of the "active" program, *NECview*, interest in *Wires* seems to have waned considerably.

NECview

*NEC*view, like *Wires*, provides a picture of an antenna by reading *NEC* input files. Unlike *Wires*, *NECview* provides only a single window. This window has "sliders" that allow the antenna to be rotated continuously so you can view it from different angles. *NECview* is visually stimulating and seems to catch the imagination of users. It's based quite heavily on a program called *xNECview* by P. T. deBoer.[11]

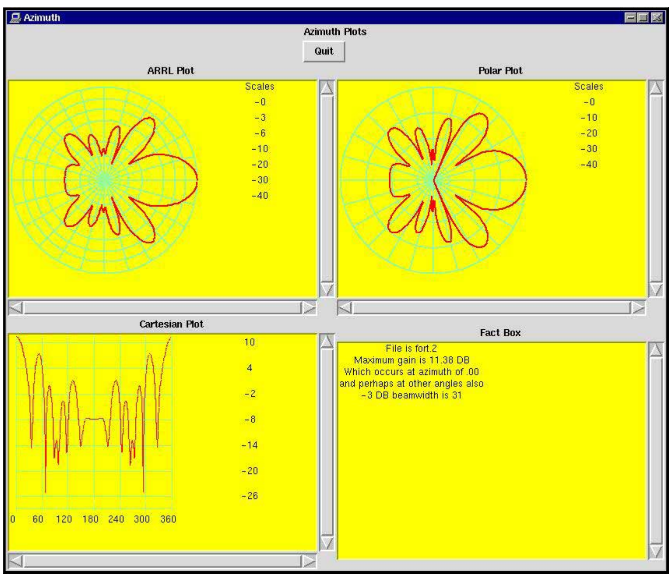

Fig 2—An example of the *Azimuth* screen.

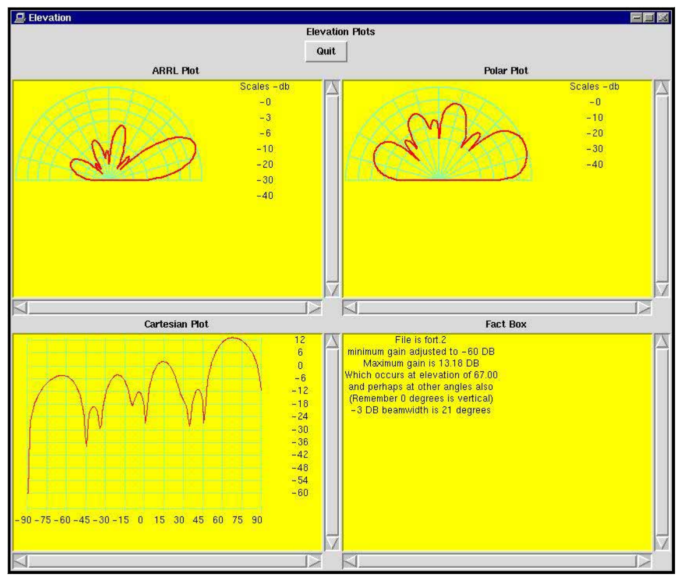

Fig 3—An example of the *Elevation* screen.

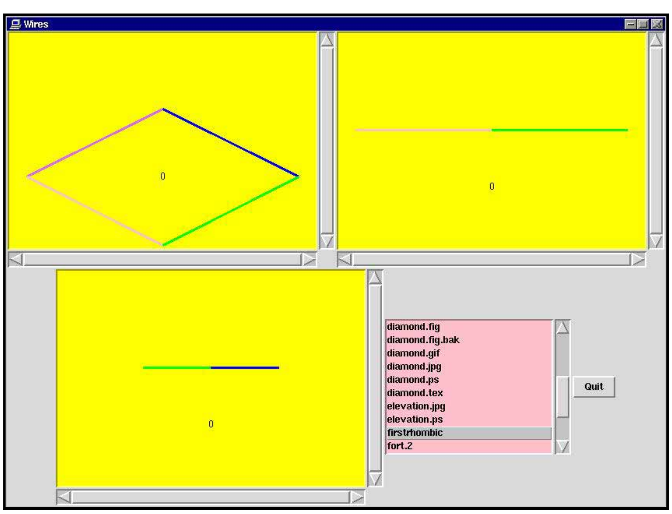

Fig 4—The *Wires* screen

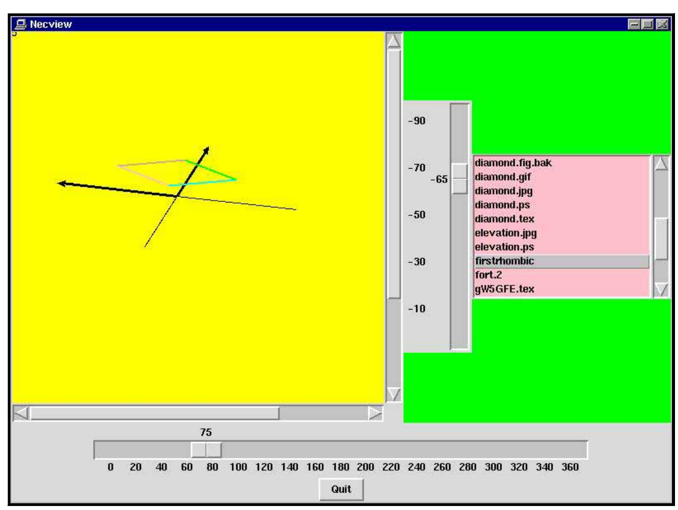

Fig 5—*NECview* in action with the rhombic input file.

Currently, *NECview* only supports the GW and the GR data lines. Fig 4 is a screen shot of *NECview* in action with the rhombic as an input file.

Availability

The entire W5GFE antenna package is available via anonymous FTP.[12]

Conclusion

In an article long ago,[13] I explored some propagation effects at VHF. How I wish I'd had modeling software such as *NEC* then!

If you like to fiddle with antennas, the *NEC* modeling software combined with the W5GFE antenna tool kit will provide many enjoyable hours at the computer and inspire more than a few hours on a ladder. It is no longer *NEC*essary to "fly by the seat of your pants" when designing antennas. Enjoy!

Notes

[1]R. Lewallen, W7EL, W7EL Software, PO Box 6658, Beaverton, OR 97007.

[2]L. B. Cebik, W4RNL, "A Beginner's Guide to Using Computer Antenna Modeling Programs," *The ARRL Antenna Compendium*, Vol 3.

[3]J. Belrose, VE2CV, "Modeling HF Antennas with MINI*NEC*—Guidelines and Tips from a Code User's Notebook," *The ARRL Antenna Compendium*, Vol 3.

[4]A really useful and rich reference for *NEC* documentation, code and utilities is at **www.qsl.net/wb6tpu/**.

[5]*NEC* documentation is downloadable from **members.home.net/nec2/**.

[6]The EGCS group merged with the GNU *C* project located at **gcc.gnu.org/**. Look for their *C* and *Fortran* compilers at **ftp://ftp.freesoftware.com/pub/sourceware/gcc/releases/index.html**.

[7]Bill Walker, "Jiffy" Visualization Software for NEC-based Antenna Design can be downloaded as W5GFE.tar.gz from **www.qsl.net/wb6tpu/swindex.html**.

[8]L. Wall, T. Christiansen and R. Schwartz, *Programming in Perl,* Second Edition (Cambridge, Massachusetts: O'Reilly & Associates, Inc; **www.oreilly.com**; 1996).

[9]N. Walsh, *Learning Perl/Tk*, (O'Reilly & Associates, 1999).

[10]The Comprehensive *Perl* Archive Network (CPAN) home page is at **www.cpan.org/**.

[11]P. T. deBoer, **pa3fwm@amsat.org**, **ptdeboer@cs.utwente.nl**.

[12]The entire software package is available as W5GFE2 .tar.gz at **www.qsl.net/wb6tpu/swindex.html**.

[13]Bill Walker, "Predicting Radio Horizons at VHF," *QST*, June 1978, pp 28-29.

Bill Walker holds an Extra-class license. He received his "ticket" in 1961 at the age of 14 and has held the same call throughout his ham life. He also holds BS, MS and PhD degrees in mathematics and is currently Professor and Chairman of Computer Science at East Central University in Ada, Oklahoma, where he lives with his wife Anita and their son Dalton. Anita also holds the PhD degree in mathematics and she is Professor of Mathematics at the same institution. Professor Walker has authored three textbooks on computing and writes a regular column ("Pow-wow Circle") for Whispering Wind magazine, a publication which is devoted to Native American crafts and culture. Dr. Walker and his family are active participants at powwows throughout Oklahoma. He has written articles that have appeared in QST and 73.

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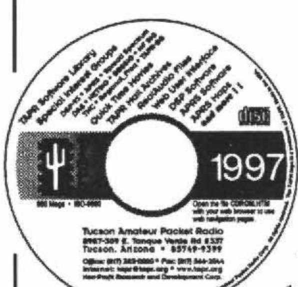

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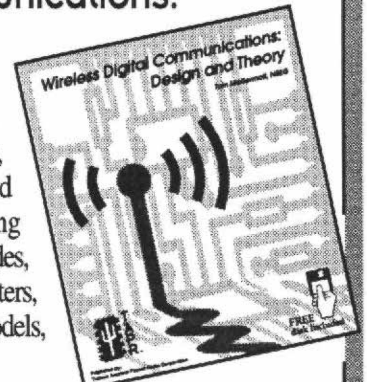

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Tunable Toroids

[While Varicap diodes are the component of choice in VCO designs, at one time at least one company, Vari-L, marketed a line of small magnetically tunable inductors. By varying the current through an internal electromagnet, the inductance could be adjusted over a large range. In RadCom's *December 2000 and April 2001 Technical Topics columns, Tech Topics Editor Pat Hawker, G3VA revisited the concept of using a variable magnetic field to achieve permeability tuning—along with some possible advantages offered by these implementations. This column has been adapted from those columns.—Peter Bertini, K1ZJH,* QEX *Contributing Editor,* **k1zjh@arrl.org**]

There are several advantages given by permeability tuning, not least the virtually consistent Q throughout the frequency sweep. In the Amateur Radio field, permeability tuning was exploited by Collins Radio in many of their excellent post-war receivers such as the 75A series in the 1950s and their later S-line transceivers. Permeability tuning was also used in several car-radio broadcast receivers made by such firms as Radiomobile in the 1950s, and it was commonly used in television receivers.

The usual technique was mechanical, moving ferrite or powered-iron cores into and out of fixed solenoid inductors, but even then, a few designers took advantage of electrical techniques to change the permeability by magnetic means. Jack Hardcastle suggests that is time to look again at these techniques, both at LF (136 kHz) and HF. He also shows how they can be applied to toroid cores.

from Jack Hardcastle:

The renewed interest in the LF and VLF bands has so far passed me by. Not having acres in which to erect antennas has demotivated me for building equipment for either 73 or 136 kHz. However, I have been intrigued by the particular problems of making tunable circuits at these unfamiliar frequencies. Space considerations rule out the large diameter, multi-tapped inductors and variometers so beloved by our predecessors. Lack of availability rules out very large capacitors. So, what are the alternatives?

It seems to me that the most fruitful avenue to explore is to use relatively low-impedance circuits and to devise some means of varying the inductance. The obvious way to do this is to use a cylindrical coil of wire and to insert a rod of powered-iron or ferrite—(as in a ferrite-rod antenna). This has indeed been done in the past, notably in the RAF's T1154 transmitter and in the Collins S-line equipment. Both of these use a linear motion to control the tuning cores, a technique that is difficult to emulate without considerable mechanical-engineering resources. I felt, however, that a rotary motion could be more readily implemented, so I made tests using a pair of ferrite E-cores (see Fig 1).

So that the variable inductor could be readily tested, I built it into a Hartley test oscillator (see Fig 2). I found that rotating one E-core through 90° resulted in a frequency change from 157 kHz to 232 kHz. I must confess that I never actually made a mechanical drive to perform this rotation. At its simplest, I can visualize it being comprised of an epicycle slow-motion drive with one of the cores cemented to a rod attached to the drive.

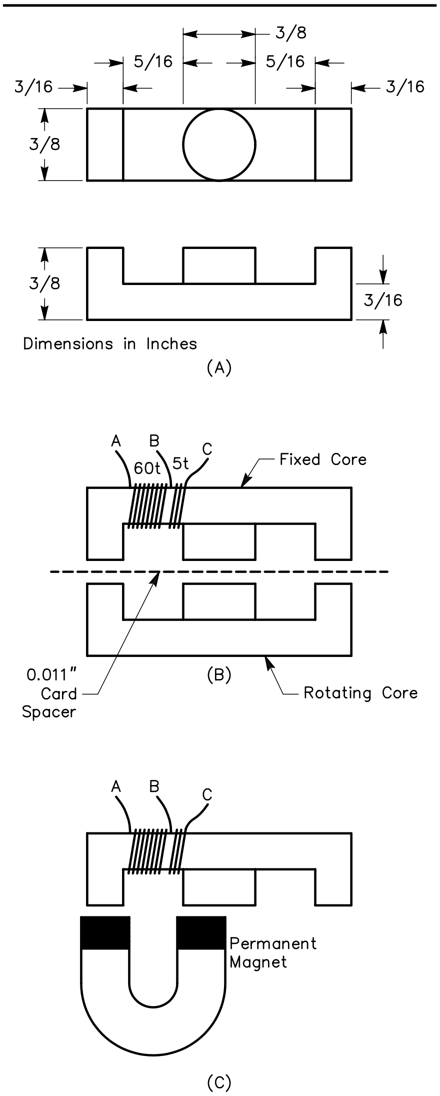

Fig 1—Experiments with a coil wound on an E core. (A) shows front and top views of a bare core. At B, a coil is wound on one half of the core and the second half is rotated (on the axis of the E's center leg). At C, one half of the core is removed and a permanent magnet is brought close to magnetically bias the core.

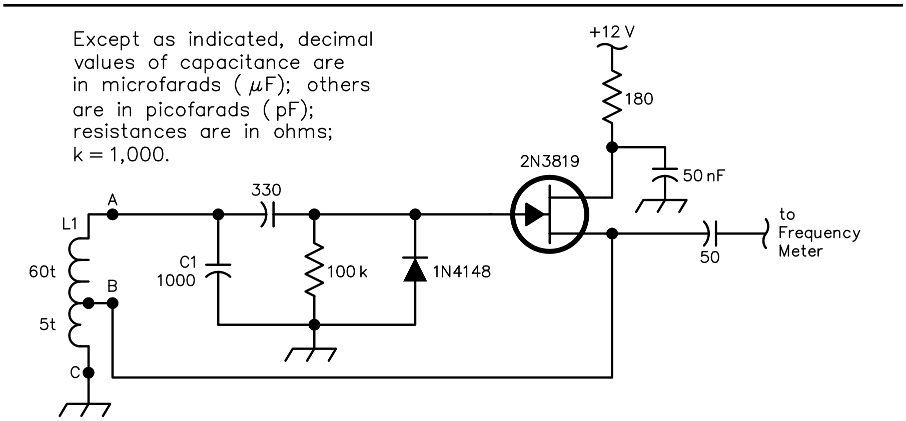

Fig 2—A simple Hartley oscillator used to test inductance variation.

Another, fairly obvious, way of changing to effective permeability of the ferrite cores is to increase the air-gap between them, but there is also a more subtle approach. If an external magnetic field is applied to a ferrite core, it drives the material toward saturation, lowering the permeability as it does so, thereby raising the oscillator frequency.

The use of an external magnetic field to lower the permeability of ferrite cores is a well-documented technique.[1,2,3] The technique was used before semiconductor tuning diodes were available to make 'wobbulators' (or gyrators) and to apply automatic frequency control.

In all these cases, the magnetic field was produced by an electromagnet so that a frequency sweep could be produced. However, for manually tuning a resonant circuit, the same result can be produced using a permanent magnet as shown in Fig 1C. For instance, the above test circuit showed that by bringing a strong horseshoe magnet near one of the E-cores (with one core removed) the frequency could be increased from 232 kHz to 421 kHz.

It is also possible, as noted by M. G. Scroggie, to tune inductors wound on toroids. By applying the magnetic field generated by a relay coil to a coil wound on an FT50-30 ferrite toroid, the oscillator frequency can be swept from 421 kHz to 1 MHz when the relay current was increased from 0 to 117 mA (see Fig 3). In practice, it is not normally required to tune over a very wide range, so the system is not as power-hungry as it appears at first glance.

So, what is the significance of this technique to radio amateurs? It allows resonant circuits to be tuned over a very wide range, particularly at very low frequencies. The bias toward the low-frequency end of the spectrum is a consequence of needing to use materials of lower permeability at higher frequencies. This limits the potential for reducing the effective permeability by whichever means is used.

It provides an alternative to variable capacitance as a means of manual tuning. It provides an al-

[1]Notes appear on page 57.

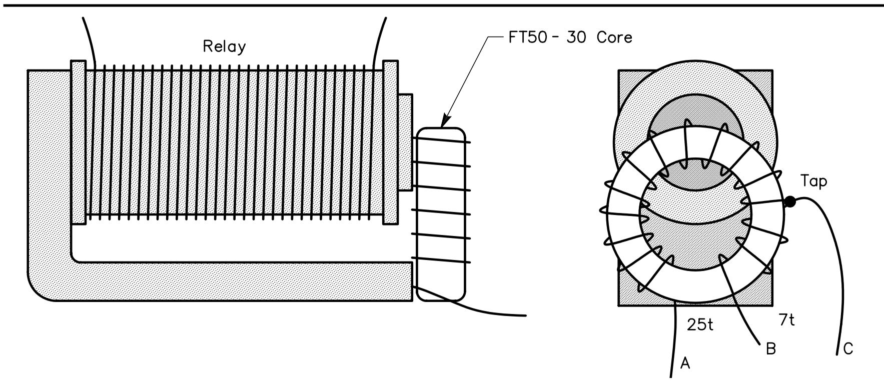

Fig 3—A toroid inductor is positioned near a relay coil so that a dc-generated magnetic field from the relay can affect the permeability of the toroid core.

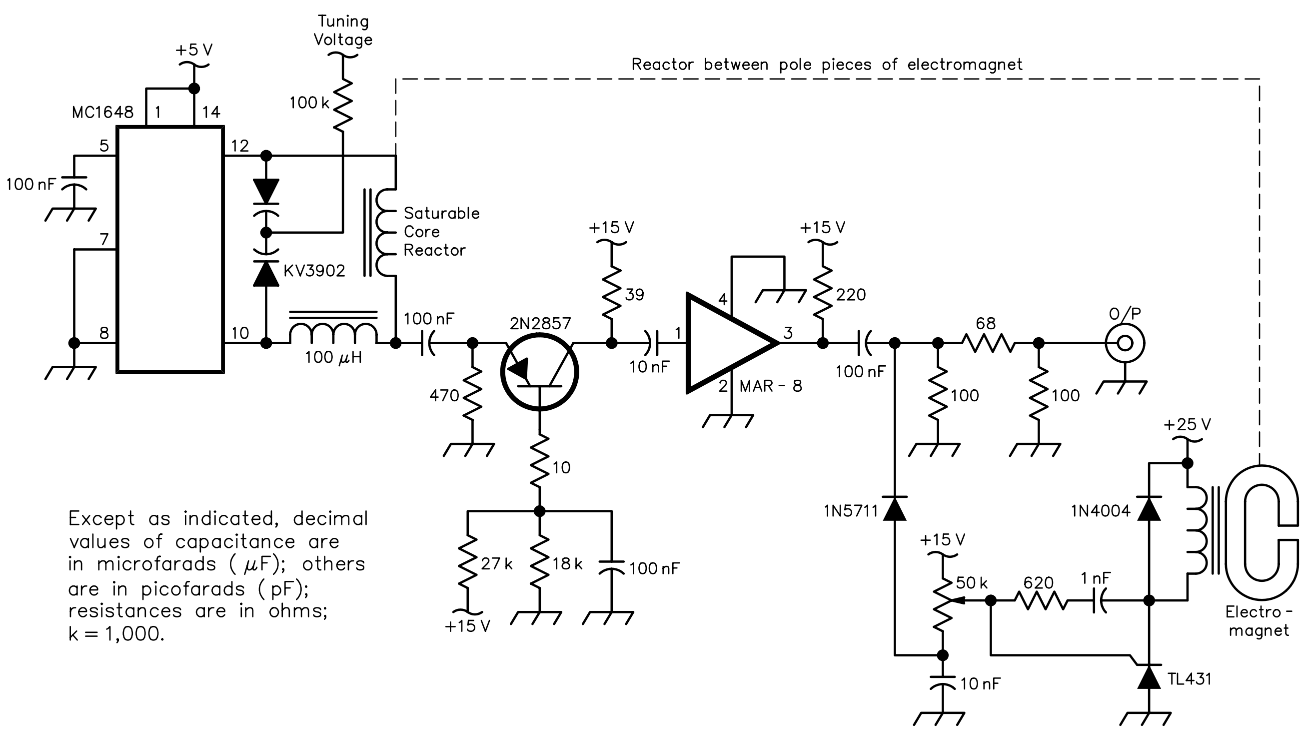

Fig 4—Circuit diagram of the wide-range constant-reactance voltage controlled oscillator (CRVCO) with a permeability-tuned TO-50 toroid-core inductor. Although the *RF Design* article was originally published some years ago, the *2000 Data Digest* books show that versions of all the semiconductor devices were still in production last year. The MC1648D is an 8-pin SO device; the MC1648L and P devices are 14-pin DIP versions.

ternative to varactor diodes as a means of electrical tuning. More speculatively, it may be a means of reducing phase noise in oscillators and synthesizers. Because varactor diodes are also noise sources, any means of eliminating them from a circuit is advantageous to designers. It remains to be seen whether ferrite materials are a significant source of noise, comparable with a varactor.

A WIDE-SPAN TUNED-TOROID VCO

David Mackenzie, GM4HJG, recently stumbled onto an Internet page that provides an interesting example of the practical value of permeability tuning—using an FT50 toroid inductor placed in the field of an electromagnet.

It appears on the Wenzel Web site (**www.wenzel.com/pdffiles/crvco.pdf**) as a reprint of one of the *RF Design* Awards articles as originally published in the magazine, probably in the period from 1985 to 1989. (I searched the contents pages of issues later than 1989 without success).

The article "Constant Reactance Voltage-Controlled Oscillator" (CRVCO) is by Raymond T. Page of Wenzel Associates Inc. It shows how a very-wide-range VCO providing good frequency and amplitude stability over a tuning range of 20 to 150 MHz (7.5:1 span) can be implemented without any complex band switching. These characteristics are much superior to the conventional VCO, which tends to be limited to a frequency span of not more than 3:1 with the loaded Q and resulting circuit stability degraded at the end of the frequency range. By actively holding the inductor's reactance constant in a feedback system that tracks the tuning varactor, high Q and exceptionally constant output power are maintained over the entire frequency span.

It is stressed that this form of CRVCO implementation need not be restricted to high-ratio tuning applications. Designs with smaller-ratio tuning spans can benefit from its inherent stability. The key feature is the use of permeability tuning as suggested by G3JIR, with the toroid mounted between the poles of an electromagnet, in conjunction with an ingenious feedback (AGC) system. The "saturable-core reactor" (tunable toroid) consists of a ferrite FT50 toroid-core inductor sited in the variable magnetic field. As the magnetic flux is increased, the 4C4 core loses permeability without significant changes in Q. A convenient electromagnet is formed from a modified Wabash reed relay (#208-31-1) with a soft iron rod replacing the reed switch and used to direct the saturating magnetic field to the toroid. The coil uses seven turns of #30 AWG wire on a Ferroxcube core (#135TO504C4). This combination requires less than 100 mA to saturate the inductor fully. With a tuning voltage of 1 to 24 V, the CRVCO tunes from 20 to 150 MHz with the output response a barely detectable ±0.04 dBm from end to end.

Jack Hardcastle, G3JIR, comments:

The small size of the photocopied Web diagram (Fig 4) was quite challenging *[I heartily agree!—G3VA]* and for a time I wondered whether the position of the 100-µH inductor was correct, since it appears to be in series with the tuning inductance and the Varicap that form the oscillator tuned circuit. In order to clarify this, I visited the Wenzel Web site and downloaded the circuit for myself. This enabled me to view an enlarged picture so I could see more clearly that my first impression was correct.

On investigating the internal circuit of the MC1648, it became clear that the 100-µH choke was a dc path for biasing the IC. The RF path from the tuning inductor is actually via the 100 nF capacitor, the 2N2857 emitter-base circuit, the 10-Ω resistor and the 100-nF decoupling capacitor. Once I realized this, the rest of the circuit fell into place. It is a most ingenious idea that deserves to be better known. I am not sure whether the MC1648 PLL device is still available *[see caption to the diagram—G3VA]* but the idea could be applied to any discrete component oscillator, and I hope to try this some time in the future. GM4HJQ has certainly turned-up a most innovative application of the magnetically-tuned inductor.

Notice, however, that Raymond Page stressed that the exceptional performance of this CRVCO depends, not only on the tunable toroid, but also on the choice of other components.

From Page's article:

The MC1648 is selected as the VCO because it contains an automatic-gain control that precisely sets the voltages across the tank, allowing the inductor's reactance to be determined by measuring its current. This current is metered by connecting the ground end of the coil to the synthetic ground at the collector of a grounded-base stage. A voltage proportional to the emitter current appears at the collector. This voltage is amplified and detected. The low-impedance collector resistor and MMIC amplifier provide very flat wide-band response.

Once detected, the inductor current results in a dc voltage that is scaled by a 50-kΩ potentiometer before it is applied to the reference pin of a TL431 shunt regulator. In this unique application, the TL431 modulates the current into its cathode in an attempt to keep the reference pin at 2.5 V dc.

As an increasing voltage is applied to the varactors, the VCO frequency begins to rise, which makes the inductor current start to drop. Since this drop in inductor current shows up as a proportionate dip in RF voltage at the detector, the voltage at the reference pin of the TL431 will attempt to increase, causing the cathode to sink more current. This increases the saturation of the toroid and lowers its inductance, bringing the current back to its preset level, thereby satisfying the feedback loop. The compensation network (620-Ω resistor in series with a 1-nF CIF capacitor) assures that the frequency response of the TL431 is slower than the frequency response of the electromagnet for good loop stability.

The combined performance of the grounded-base stage and MMIC stage play a crucial role in just how well the toroid reactance can be regulated. This trans-resistance amplifier is useful to 400-MHz.

G3JIR points out that the Wenzel Web site also contains useful material on such items as "Low-Cost Phase-Noise Measurement" (3 pages), "A Low-Noise Amplifier for Phase-Noise Measurements" (3 pages) and "Phase Noise, Harmonics and Sub-Harmonics" (2 pages).—*G3VA*

Notes

[1]Pressman and Blewett, "300-4000-kHz Electrically Tuned Oscillator," *Proceedings of the IRE*, January 1951, pp 74-77.

[2]M. G. Scroggie, "An unconventional FM receiver," *Wireless World,* October 1957, p 505.

[3]A. E. Ford and J. S. White, "An Insertion Loss Display and Recording Equipment for the Frequency Range 50 kHz to 8 MHz," *Post Office Electrical Engineers Journal,* October 1960, pp 145-50.

RF

By Zack Lau, W1VT

A Low-Cost 222-MHz Helical Band-Pass Filter

When building simple transverters, it is often difficult to meet the FCC spectral-purity requirements. This two-resonator 9-MHz-bandwidth 222-MHz filter is designed to reject common mixing spurs while attenuating the desired signals just 1 dB. This filter rejects 194 and 260 MHz spurs by 35 dB. These are the most noticeable spurs resulting from a transmit conversion from 28 to 222 MHz.[1] The 194-MHz spur is the local oscillator. The 260-MHz spur is a second harmonic of the IF plus the LO—a high-order mixing product. The 194-MHz signal often leaks around filters in designs with insufficient shielding. Techniques for lower-loss filters will also be discussed. As a comparison point, a Toko CBW two-pole filter designed for circuit-board mounting has a measured loss of 1.5 dB and a 3-dB bandwidth of 7 MHz, after careful tuning for minimum loss at 222 MHz. While it occupies just a quarter of the volume, it also costs $18.

The basic design equations for helical resonators can be found in any recent edition of the *ARRL Handbook*. These equations describe the optimum coil dimensions for a given resonator size and frequency. Unfortunately, these a just a start toward a design, as most amateur applications require two or more resonators to achieve a flat passband and sufficient stop-band response. Thus, it is necessary to determine and achieve the optimum coupling between the resonators. Obtaining a repeatable filter design is considerably more difficult at VHF than HF.

At HF, small coupling capacitors often have adequate tolerance for building band-pass filters. At VHF, although sub-1-pF capacitors exist, their tolerances are often too loose to be useful, except in custom one-of-a-kind designs built with the assistance of costly test equipment. A more serious problem is coupling via electromagnetic

[1]Notes appear on page 61.

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fields. It is difficult to capacitively couple two resonators without introducing significant electromagnetic coupling. Therefore, most experimenter's skip the idea of capacitive coupling between resonators and use electromagnetic coupling instead.

The most practical method of coupling two resonators is to cut a hole through a common shield, often called aperture coupling. However, I found that the orientation of the coils is critical. Thus, with typical air-core prototypes, the optimum aperture can vary considerably. Slicing off a quarter-turn of a six-turn inductor may result in a considerably different coupling between the resonators.

This may not be a problem with two resonators, if you can sweep the filters. Adjusting the coupling until the passband looks good is an excellent technique, if you have the required equipment. An excellent article that discusses this problem is "The Double-Tuned Circuit: An Experimenter's Tutorial."[2] Merely adjusting the filter for minimum loss at 222 MHz isn't enough—if the resonators are over-coupled the bandwidth might easily be two or three times the desired bandwidth. This may result in much less attenuation of mixing products than expected. This is easily seen by sweeping a filter with no aperture stop—it is possible to have no useful attenuation of one or more of the mixing products 28 MHz away.

Fortunately, there is a simpler way to determine whether the filter is over coupled. Over-coupling of the 222-MHz band-pass filter can be detected with a wideband VHF/UHF SWR analyzer, such as the Autek RF-5. If the filter is terminated in a 50-Ω load, the only frequencies with a low SWR are in the desired filter passband. If you see a low SWR in the stop band, the filter is over-coupled. It is then necessary to reduce the coupling and retune the filter until there is no evidence of the unwanted filter response. It is possible for the peaks to be relatively far away—the Autek RF-5 doesn't have enough range to tune a 2-meter filter. An unwanted response could easily occur in the 75 to 138 MHz frequency gap. The SWR analyzer is also useful for determining whether your VHF load is close to 50 Ω—layout is just as important as the parts used, when working at VHF.

The equations predicted that seven turns of #16 wire wound on an RF transparent 0.60-inch form would be optimum for a 1-inch square enclosure. Experimentally, I found that six turns worked well, allowing the resonators to be tuned down to 222 MHz with tuning screws. I measured an insertion loss of 0.90 dB and a 3-dB bandwidth of 8.6 MHz. A second filter showed an insertion loss of 1.1 dB and a bandwidth of 11.4 MHz.

I also constructed a version without the coil forms in place. The coils were wound tightly on a 0.625-inch form and then removed. Due to the slightly larger diameter, the filter resonated slightly lower in frequency, even with one of the tuning screws removed—the center frequency was 220.6 MHz. The bandwidth was 7.9 MHz and the insertion loss at 222 MHz was just 0.57 dB. Cutting a quarter turn from the coils (to 5.75 turns) allowed the center frequency to reach 222 MHz. The insertion loss was 0.65 dB with a bandwidth of 4.5 MHz. The coupling was significantly altered by this change. The height of the hole was originally 0.55 inches—the new filter had a 0.63-inch height. The hole is bigger and the coupling is less, as indicated by the smaller bandwidth.

This indicates that the CPVC form introduces significant loss. Nevertheless, I think this version is preferred. It can be used to precisely position the spacing and orientation of the coil.

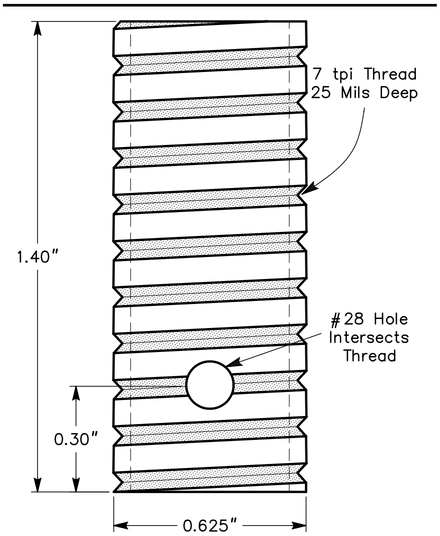

Fig 1—Dimensions for the "1/2" CPVC coil forms. The hole for the coil tap intersects the seven tpi thread.

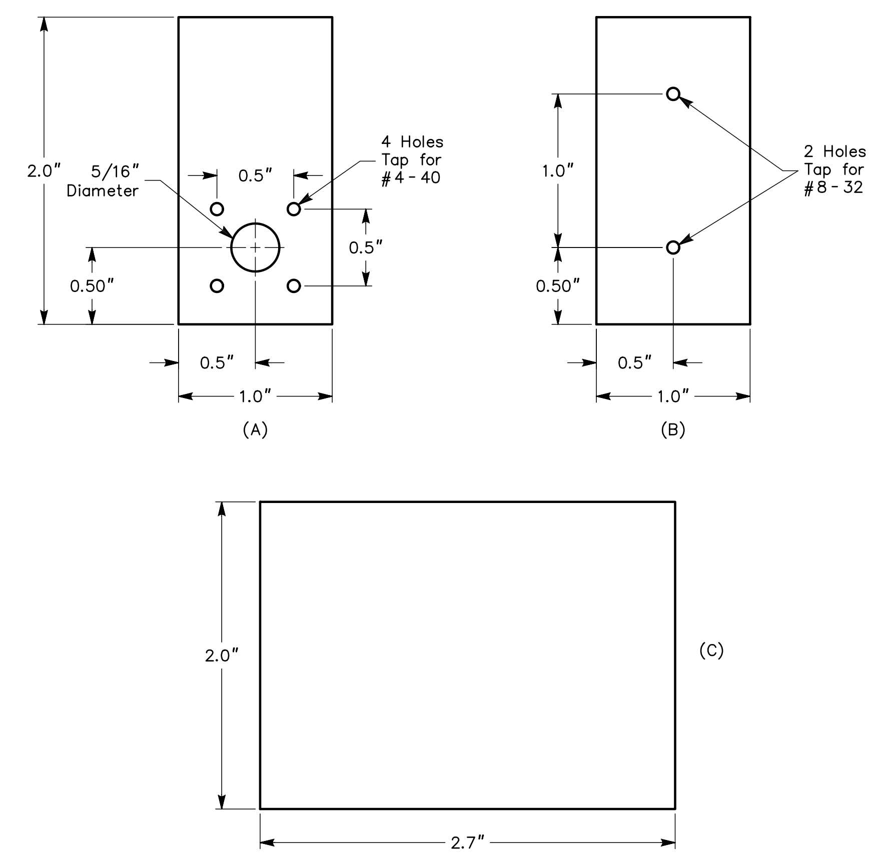

Fig 2—At A, dimensions for the 0.025-inch brass end walls. At B, dimensions for the 0.032-inch brass top plate. At C, dimensions for the copper-circuit-board side walls.

This may allow a design to be more repeatable. The form also holds the coil in place and reduces the effect of vibration on the filter. This may be important for mobile operation. I've seen significantly lower loss using Teflon as the insulator,[3] but Teflon is significantly more expensive and difficult to machine.

In the past, I've experimented with commercially made coil stock, but the wire is too thin for a given pitch to have an optimum *Q*. For most experimenter's, this isn't a viable option, as it has become difficult to find in small quantities. It is possible to make your own—I'd recommend reading the two articles by Robert Johns, W3JIP.[4, 5] A homebrew coil can be easily designed to have a more optimum wire diameter.

Fig 1 shows a drawing of the CPVC form. I used a lathe to cut a seven-turn-per-inch thread—this allows precise turn spacing. The thread is approximately 25 mils deep. When a coil is tightly wound on this form, the inside diameter is approximately 0.60 inches. The length of the form is designed to fit between the center conductor of the BNC connector and the shield. By drilling a hole for the tap point in the form, the height of the coil from the shield can also be defined. Accurately positioning a coil is much more difficult to accomplish without a form or alignment tool.

The coils are each wound with a 13.5-inch length of bare #12 AWG copper wire. I prefer to start with at least 16 inches of wire and bend it to shape. After the copper is formed, I trim the wire to the length. The extra wire makes bending easier. Obtaining the precise length is difficult; I find the actual lengths to be ±0.25 inches. Ordinary two- or three-conductor solid house wire from the hardware store is a cheap source, though it may take some effort to remove the insulation. Three-conductor wire may be preferred—the ground wire is easy to remove. I first wind the cores on a 0.590-inch diameter CPVC form, and then spread the turns apart to approximately the right pitch. The coils are then threaded onto the form. I cut the form down to size with a lathe—this is much easier than winding the coil on a half-inch form and attempting to loosen the turns sufficiently. The final coil should have precisely six turns—the ends of the wire should line up at the same angle with respect to the center of the coil. I aligned the forms to be concentric with the tuning screws.

A formula for the helix wire length is[6]

$$L = \sqrt{a^2 + (n\pi\, d)^2} \qquad \text{(Eq 1)}$$

where

a=coil length
d=coil diameter
n=number of turns

Use #24 AWG bus wire to connect the tap points to the center pins of the BNC connectors. This wire is commonly available from RadioShack.[7] I tested a filter using #20 wire, but there was no significant reduction in loss. The coupling was significantly increased, however, so I do not recommend a substitution unless you want to develop your own design. The thinner wire is easier to work with—thick wire is more easily shorted against adjacent turns. The tap points set the input and output coupling of the filter. For a 9-MHz bandwidth, they should be set one-third of a turn or 120° from the grounded end of the helix. They can be changed if you want to change the bandwidth of the filter. A narrower filter requires the tap points to be set closer to ground, a wider filter requires more coupling. At VHF, the inductance—even of short wires—may be enough to affect the coupling, just as the coupling of HF filters can be changed by inserting inductors in series.

Fig 2 shows drawings for the shield walls. I use 25-mil brass sheet for the end walls with UG-290A/U coax connectors attached. The other walls are made out of single- or double-sided circuit board—single sided might actually be better, as the bare fiberglass can be bolted to aluminum without worry of electrolytic corrosion. However, double-sided unetched circuit board is commonly available as surplus—cheap single-sided board may actually be more difficult to find. The top shield wall is made out of 32-mil-thick brass—it holds an #8-32 thread better than does thinner material. Brass that is 25 mils thick will also work—in this case, you would have three plates that are exactly the same size. This is convenient if you are cutting plates from sheet stock—I've been known to clamp plates together and precisely mill them all to the same size. However, it is more convenient to buy brass strip stock already cut to 1 or 2 inch widths.[8] There is no bottom shield, though you can add one for more complete shielding.

I first solder the top shield wall to one of the copper shield walls, taking time to make sure the spacing from the edge is accurate. The calculation is as follows. The coax connector has an 83-mil-diameter pin with a 23-mil notch cut into it. Thus, the distance is:

$$\begin{aligned} spacing &= (shield\ height) - (coil\ form\ height) - (pin\ spacing) \\ &\quad - (connector\ distance) \\ spacing &= 2.000'' - 1.400'' - 0.019'' - 0.500'' = 0.080'' \end{aligned} \qquad \text{(Eq 2)}$$

If you don't want the form to fit neatly into the notch, but want it to rest against the full diameter of the pin, the spacing is 0.059 inches.

I next solder the brass plates. It's useful to first solder in one of the coils in place—you can get a better idea of how the filter goes together without the last shield wall soldered in place. After the shield wall is soldered in place, the second coil can be soldered to it. The connectors are attached with solder and #4-40 screws. I used 1-inch #8-32 nickel-plated brass screws and hex nuts for tuning. Once the filter is tuned, the hex nuts can be locked down to maintain alignment. Brass screws can also be used, but avoid steel- or zinc-plated screws, as they may adversely affect the insertion loss of the filter.

The challenging part is the alignment of the filter. Ideally, one could look at the filter on a network analyzer while tweaking the filter for optimum insertion and return

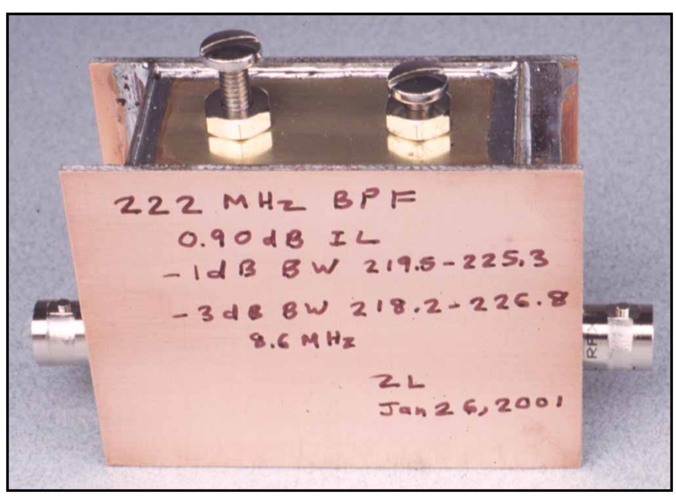

Fig 3—The 222-MHz band-pass filter

losses. However, most of us must improvise with simpler equipment. I've tuned these filters with only 222-MHz signal generator and a spectrum analyzer as a detector, but have often found the filters to be mistuned—overcoupled. I then must reduce coupling and realign the filter until there is no evidence of two separate filter peaks.

The coupling is adjusted with an aperture stop, a 1-inch strip of metal that slides between the two resonators. It should be a tight fit, so that good electrical contact can be maintained without soldering. The initial coupling should be around 0.5 inches—there should be a half-inch gap between the top of the shield and the aperture stop. The tuning screws and aperture stop are then adjusted for minimum loss at 222 MHz. The positioning of the aperture stop is rather critical, 0.1 inch may make a big difference. Once the minimum-loss tuning is achieved, check the coupling by sweeping the filter over a wide frequency range. This can be done with an SWR analyzer—terminate the filter with a good dummy load and look for unwanted filter responses. If you see a good SWR outside the passband, the coupling between resonators needs to be reduced. Making the aperture smaller will reduce the coupling.

It can be difficult to adjust a helical resonator by trimming the loose end—not only is the resonant frequency changed but the coupling may also be changed. Thus, I don't recommend soldering the aperture stop until you are sure of the resonator tuning.

The filter loss can be reduced by using Teflon instead of CPVC for the coil form. (Teflon is much more costly and can be difficult to machine.) It may be worthwhile to experiment with other materials for the coil form. However, a more important consideration is the effect of passband shape and impedance mismatches on insertion loss, particularly if you want to be able to simply add a filter to a system without adjusting it.

While it is certainly possible to build very-low-loss filters, with less than 0.5 dB of loss, it can be difficult to actually implement these filters in practice. Precise coupling is essential—while 1 dB of ripple may be acceptable in a more lossy filter, 1 dB of filter ripple is not acceptable in a filter designed for less than 1 dB of insertion loss. The Toko CBW has a 0.5-dB/3.0-dB ripple/loss specification. Similarly, mismatches resulting from connectors, transmission lines or switches can wreak havoc with attempts to add a filter to an existing system with minimal hassle. With very-low-loss filters, it is often necessary to retune the filters once they have been installed in a system. The retuning can sometimes be accomplished by swapping cables, until one gets a system that performs as expected.

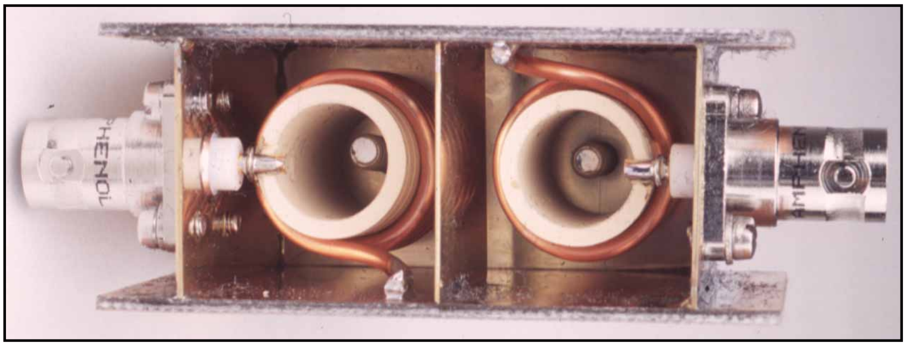

Fig 4—Inside view of the filter.

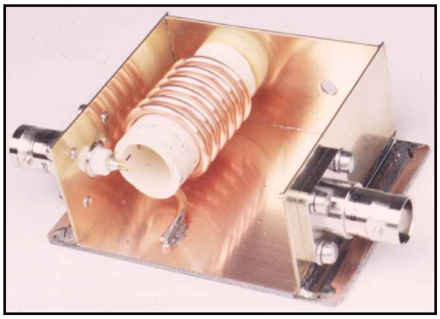

Fig 5—A view of a partially assembled filter.

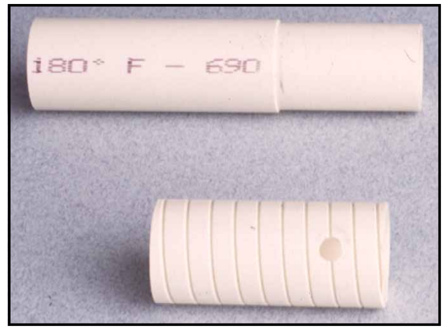

Fig 6—The coil forms.

Notes

[1]Based on measurements of a modified Ten-Tec transverter by Botts. (*QST*, May 2001, pp 28-33).

[2]W. Hayward, W7ZOI, "The Double-Tuned Circuit: An Experimenter's Tutorial," *QST*, Dec 1991, pages 29-34.

[3]A 0.9-MHz-wide 222-MHz filter with 3.5 dB loss, "A Collection of VHF Filters," Lau, Zack, W1VT, *Proceedings of the 20th Eastern VHF/UHF Conference of the Eastern VHF/UHF Society*, pp 109-114.

[4]R. Johns, W3JIP, "Homebrew Your Own Inductors!," *QST*, August 1997, pp 33-35.

[5]R. Johns ,W3JIP, "Rugged Coil Forms and Weatherproof Enclosures," *QST*, Oct 1997 pp 48-50.

[6]L. Campbell, W1CUT now W1HQ,"Finding Wire Length for Helix Antennas," *QST*, Feb 1968, p 56.

[7]RS 278-1341, see **www.radioshack.com**.

[8]Small Parts Inc, **www.smallparts.com**; tel 1-800-220-4242.

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11/98

Letters to the Editor

New Super-Regenerative Circuits for Amateur VHF and UHF Experimentation, (Sept/Oct 2000, pp 18-32)

Hello Doug,

I must take issue with N1TEV's statement on page 23: "The most important variable affecting the selectivity of a super-regen is the shape of the quenching waveform." In his book, J. R. Whitehead shows that the bandwidth of a super-regen detector is related to the rate of change of conductance across the tuned circuit as oscillations commence.[1] This has been confirmed experimentally by R. C. Becker and J. B. Beyer, "Investigation of Electronic Modification of Super-Regenerative Receiver Bandwidth," (University of Wisconsin, 1980).

What N1TEV has so elegantly done by adding a single resistor is to reduce the rate of change of conductance across the tuned circuit, and hence reduce the rate at which oscillations increase after commencement. This can be seen in Figs 7A and B. The shape of the quench waveform is the effect of adding the resistor, not the cause of the improved selectivity.

Whitehead, on pages 105-106, also shows that the bandwidth is dependent upon the signal-to-noise ratio, as N1TEV reports on page 30 [of *QEX*]. This suggests the use of a front-end attenuator to maintain NBFM selectivity.—*John Crabtree, KC0GGH, 5408 Oaklawn Ave, Edina, MN 55424-1609;* **crabtreejr@aol.com**

[1] J. R. Whitehead, *Super-Regenerative Receivers*, (Cambridge, Massachusetts: Cambridge University Press, 1950), pp 41-42.

Hi Doug,

I am very grateful to John Crabtree, KC0GGH, for his comments and his careful study of this issue. My article is based on several years of construction and experimentation rather than on a theoretical approach. It has therefore been difficult for me to determine the *exact* cause of this observed phenomenon; however, I still strongly believe that the right approach is to build the circuit, do some experiments and then try to determine your results.

The introduction of a small resistance in series with the quench capacitor most certainly does slow the build-up of oscillation in the detector. This delay means that the detector spends *more time* operating in the regenerative state, just below oscillation. Logically, this means that *more Q multiplication* occurs and therefore selectivity increases. While the addition of this resistance clearly is the cause of the added Q multiplication, the fact is that the *shape of the quenching modulation* is also changed into something very close to sine-wave modulation. The sine wave does have fewer harmonics than a sawtooth or square wave, and this reduction in harmonics should also improve selectivity by preventing these harmonics from "jamming" the NBFM signal. The fact that reducing the regeneration level (thus decreasing the amplitude of the quenching sidebands) also greatly improves selectivity does lend support to my "jamming" theory.

In any case, I certainly admit that it is *very* difficult to determine exactly which effect is most important, or whether there are other effects happening at the same time. Perhaps other amateurs will help resolve these questions by joining us in studying this truly fascinating subject.—*Charles Kitchin, N1TEV, 26 Crystal St, Billerica, MA 01821;* **charles.kitchin@analog.com**

ATR-2000 (May/Jun 2000, pp 39-51)

Doug,

Readers have caught some errors in Part 2 of the ATR-2000 series. There are two mistakes in Fig 3 on p 42. First, the top end of R2 should connect to pin 2 of U2 rather than pin 6 of U2. Second, the left end of R1 should connect to +5V rather than AGC In.

In the upper right-hand corner of Fig 7 there are two resistors connected to a 1N914 diode. The left end of the 150-Ω resistor should connect to +5 V and the right end of the 1500-Ω resistor should connect to ground. In Fig 3, D5 is backwards. The anode should connect to ground.

These were both errors in my hand-drawn schematic that I found by looking at the actual transceiver to answer some questions about its operation.

Also, it seems that there are some diehards in Europe still using Pascal so I have supplied them with source code for the ATR-2000 control program. Can this also be posted on the ARRL Web site?—*73, John Stephensen, KD6OZH,* **kd6ozh@gte.net**

You bet. The Pascal files have been added to the others in ATR2000.ZIP at **www.arrl.org/qexfiles/**.—*Doug Smith, KF6DX,* **kf6dx@arrl.org**

A Keyed Power Supply for Class-E Amplifiers (Jan/Feb 2001, pp 12-27)

Doug,

I have just looked through *QEX* for Jan/Feb 2001. I find in both *QEX* and *QST* an apparent and growing lack of careful editing. Let me illustrate using [this article] as an example.

On p 22 is the statement, "Switching frequency is determined by R13 and C20." On p 25 is the statement, "R18 and C23 determine the switching frequency." It appears no one read this article for consistency.

Over the past year or so, I have noticed similar discrepancies in both magazines. Errors of this sort certainly detract from the perceived quality of the publication (which used to be quite high). Perhaps you might want to think about these concerns and develop a policy or procedure to minimize them.—*Ernest J. Moore, VE3CZZ, 37 Ashgrove Crescent, Nepean, Ontario, Canada, K2G 0S1;* **ejmoore@trytel.com**

Dear Ernest,

You are right to criticize us for not catching errors. That article had an embarrassing number of them. We would like to make the corrections here and now.

R18 and C23 determine the switching frequency. In the parts list, C22 and C23 should be listed as 10 nF and 5 nF, respectively. Bridge U7 should be rotated one-quarter turn counterclockwise.

I agree that a change in procedure is needed to improve accuracy. We also seek a technical writer to keep us up to date on new product offerings in communications. It would be nice if we could find one person to help us with both. Anyone who is interested or who has a suggestion may contact me directly. The successful candidate will fill an important role in our efforts to continually improve—73, Doug

Doug,

After reading [this article], I became concerned that there might be a safety hazard therein. I am referring to the main transformer T1 and gate transformers T2 and T3. They do not appear to have sufficient primary-secondary insulation. The usual practice in commercial supplies is to wrap three turns of 1-mil Mylar tape between windings. The PVC tape used in T1 would not be considered a substitute. There appears to be no tape in T2 and T3. The insulation of ordinary magnet wire is considered unreliable for off-line supplies, especially when the winding methods are uncontrolled; that is, not smoothly layered.

Otherwise, [it is] a fine article. I am thinking about redesigning T1 for 13.8 V at 35 A.—*Bernard Rate, N7DAL, PO Box 1031, Astoria, OR 97103-1031;* **bernardr@daldesign.com**

Doug,

We would like to thank Bernard for his advice for improving the safety of the power supply.—*Jim Buckwalter, John Davis, Dragan Maric, Kent Potter and Dave Rutledge, KN6EK;* **rutledge@caltech.edu**

Class-E RF Power Amplifiers (Jan/Feb 2001, pp 9-20)

Dear Doug,

At Eq 4 on page 12, in column 2, $f(Q_L)$ means "function of Q_L," not "frequency multiplied by Q_L." That function is given explicitly in the last parenthesized expression of Eq 5.—*Nathan O. Sokal, WA1HQC, Design Automation Inc, 4 Tyler Rd, Lexington, MA 02420-2404;* **NathanSokal@compuserve.com**

PTC . . ., Part 2 (Mar/Apr 2001, pp 9-17)

Hello Doug,

Tell me please: Your "complete 10-stage sub-band decomposition" diagram on page 10 looks to me like it's a discrete wavelet transform. If it isn't, I'd guess it's very close to one. Have you used wavelets at all in your communications DSP work? It strikes me the way wavelets filter down into octave bands must be directly comparable to the way we hear sounds, and that wavelets should be very useful indeed.

Your comment "Most of the texts currently available were written by mathematicians for other mathematicians" rings true. Titles such as "Wavelets For Kids: A Tutorial" must be what passes for humor amongst mathematicians. I've always found articles on wavelet theory somewhat circular, in that there seems to be no obvious starting point: One thing is the transform of the other. I for one, would much like to see an article on wavelets in *QEX,* couched in the more familiar context of radios and electronics, rather than "by mathematicians for mathematicians."—*Chris Cadogan, G3XWB, 8 Horncliffe Close, Rawtenstall Rossendale, Lancs BB4 6EE, Great Britain;* **chris@cadogan.u-net.com**

Hi there Doug,

With reference to your articles on PTC, I have a couple of comments. Wavelet transforms are noted for their data-reduction capability as well as for providing variable time/frequency resolution and *denoising*. In fact, wavelets emerged primarily as a DSP technology to address the shortcomings of FFT and short-time FFT. I suspect they could help with the removal of redundancy in speech.

In the IEEE magazine *Spectrum*, Feb 2001, there is an article describing methods for specifying listening quality and listening effort for speech. Apparently there are two methods developed by the British and the Dutch that are proposed for a new ITU standard called "Perceptual Evaluation of Speech Quality." I particularly liked the idea of including a measure for listening effort, which seems to have a ham context. As amateurs progress towards applications of digital speech, I would expect to hear more about wavelets and measures of speech quality. —*Ron Skelton, W6WO, 4221 Gull Cove Way, Capitola, CA 95010;* **rskelton@surfnetusa.com**

Gentlemen,

You are right that PTC employs something that is very close to a discrete wavelet transform. I am still learning about wavelets and would like to have an article on them that explains things clearly. We, no doubt, will be hearing more about them as work on digital-voice systems progresses. The whole field of digital processing of audio signals is advancing at a remarkable pace. It is already a regular part of Amateur Radio experimentation.

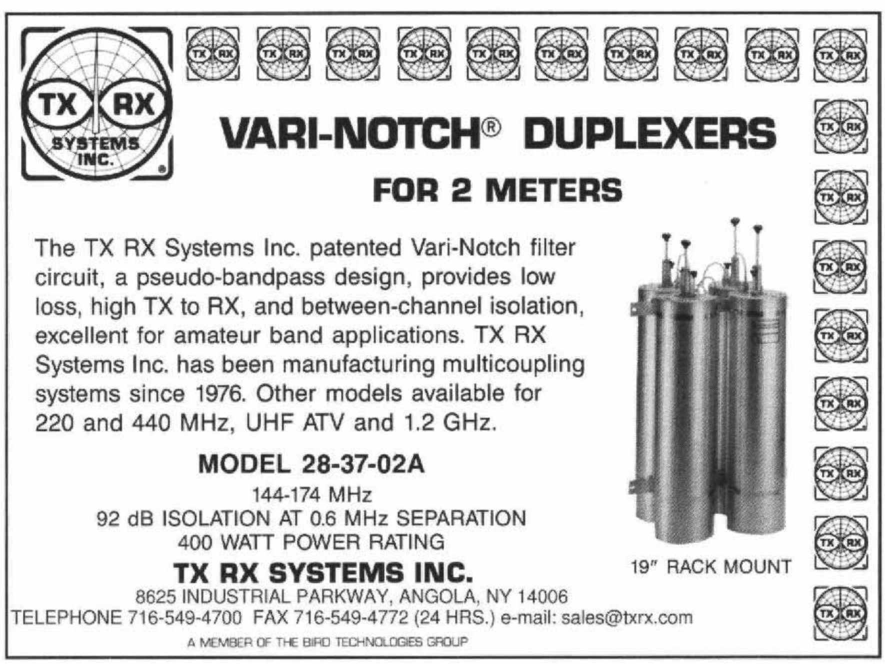
Those interested in learning more about speech-coding systems should visit the ARRL Technical Information Service page ***www.arrl.org/tis/info/digivoice.html*** *and check out some of the links there—73, Doug*

Next Issue in QEX/Communications Quarterly

As a successor to his popular series in *National Contest Journal* about short-boom, log-cell Yagis, antenna wizard L. B. Cebik, W4RNL, brings us his investigations of the long-boom variety. L. B. tackles the notion that long-boom designs tend not to show significant advantages over traditional Yagis of the same length. He sorts through various important parameters and relates each to its effect on performance. The result is a group of preliminary designs; the performance of each candidate is compared with that of a long-boom Yagi.

As with any antenna design, performance means not only gain, but front-to-back ratio, impedance, usable bandwidth and ease of construction, among other things. This article shows that long-boom, log-cell Yagis deliver an interesting and unique balance among those parameters.

Nick Hall-Patch, VE7DXR, contributes the results of his long-term MF propagation experiments. He has what may be the largest body of data recently compiled in his field. It allows him to conduct some fascinating analysis; that analysis reveals new insight into what happens around sunup and points to some possible surprises.



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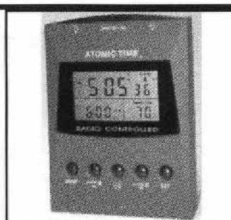

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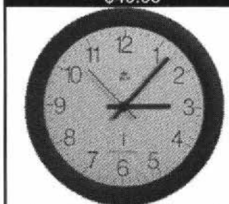

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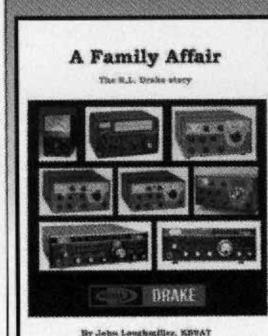

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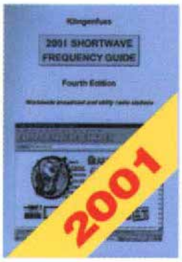

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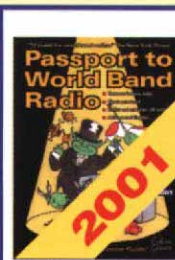

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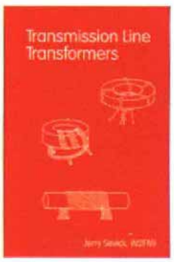

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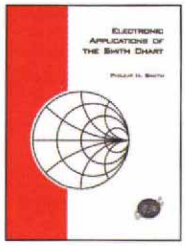

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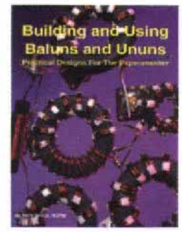

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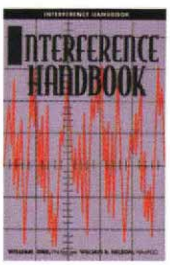

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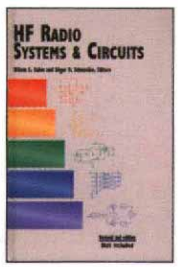

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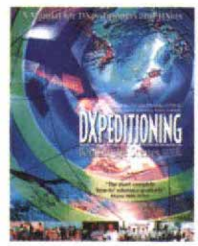

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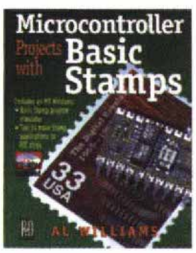

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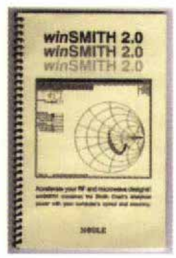

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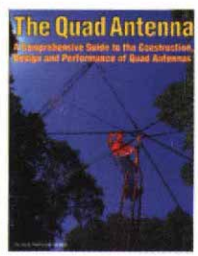

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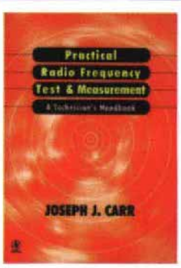

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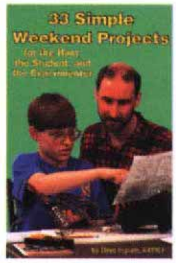

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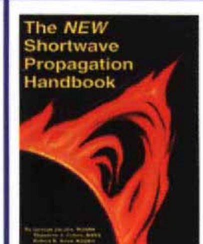

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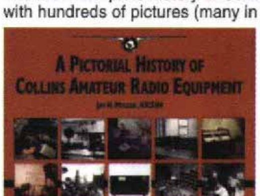

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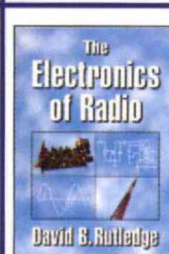

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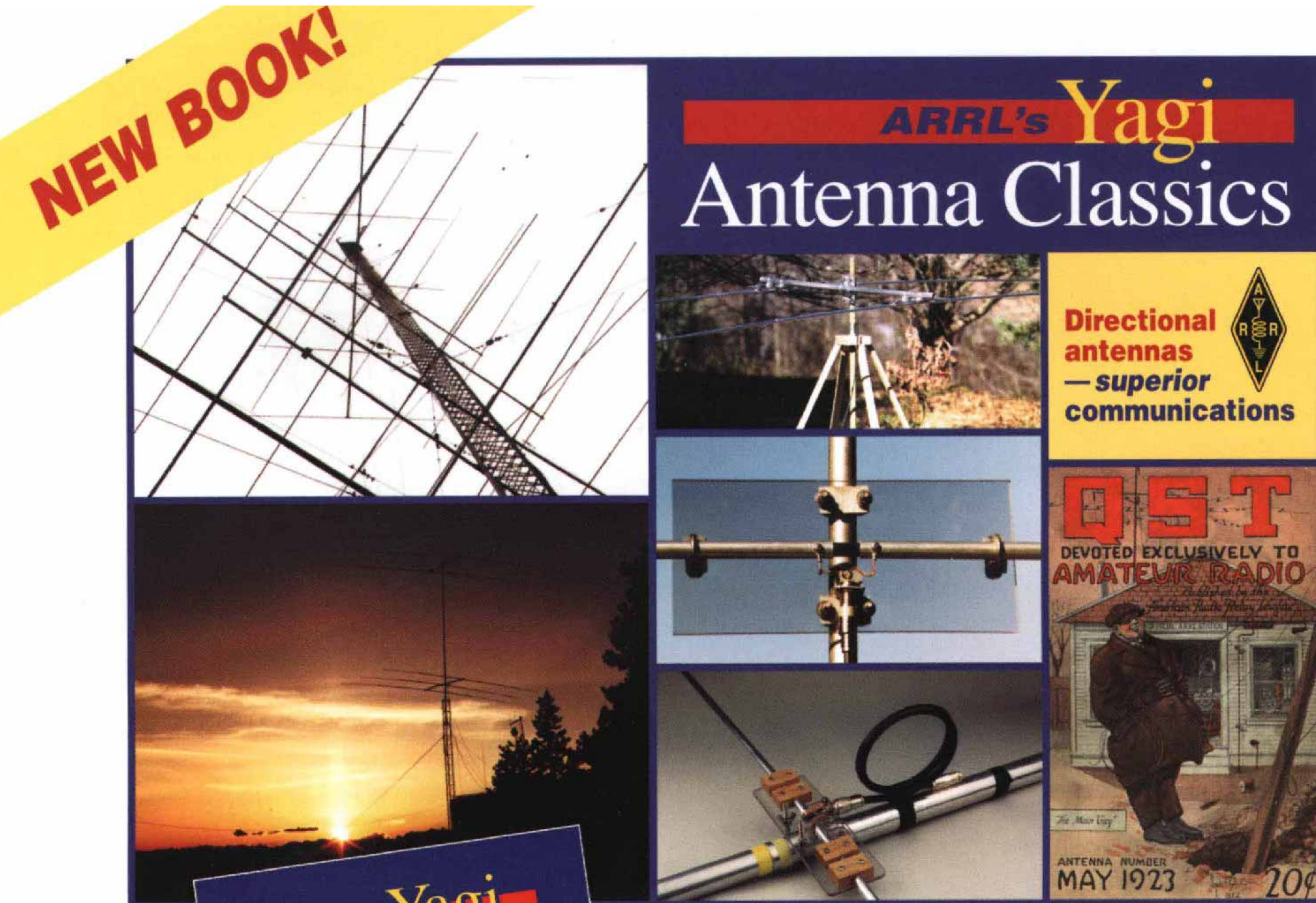

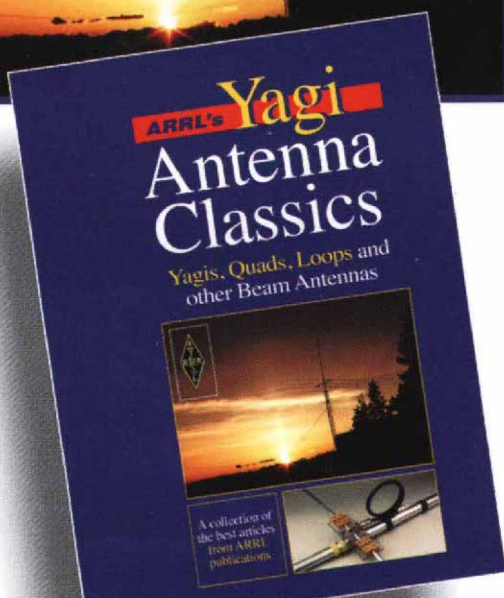

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