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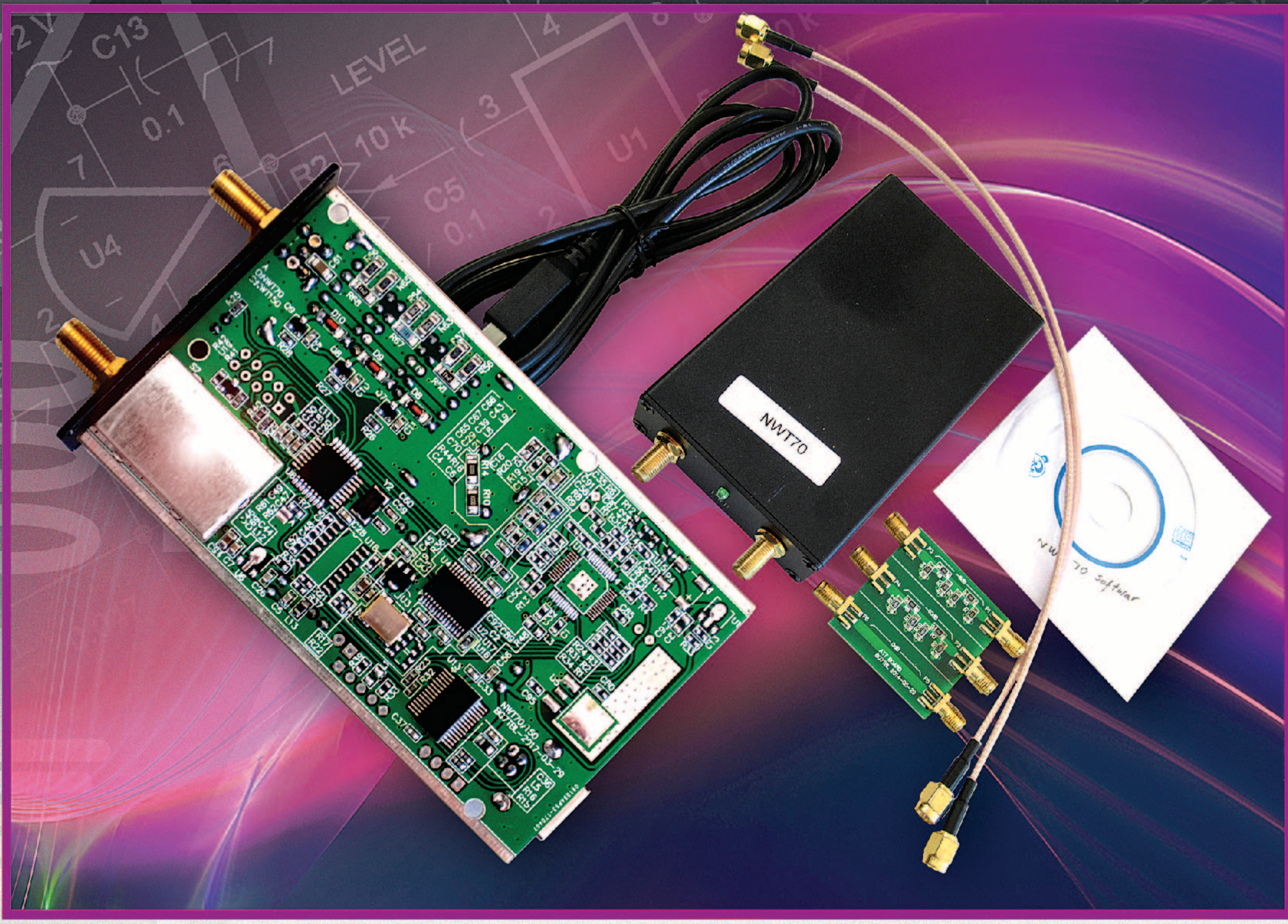

WB9LVI evaluates a very economical Scalar Network Analyzer.

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QEX

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Kazimierz "Kai" Siwiak, KE4PT
Editor

Lori Weinberg, KB1EIB
Assistant Editor

Zack Lau, W1VT
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Contributing Editors

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Brian Washing
Technical Illustrator

Advertising Information Contact:

Janet L. Rocco, W1JLR
Business Services
860-594-0203 – Direct
800-243-7768 – ARRL
860-594-4285 – Fax

Circulation Department

Cathy Stepina, *QEX Circulation*

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Fax: 860-594-0259 (24 hour direct line)
e-mail: **qex@arrl.org**

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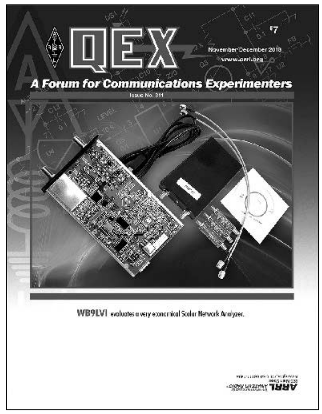
About the Cover

Dr. George R. Steber, WB9LVI, puts a very economical Scalar Network Analyzer (SNA) through its paces. A low cost SNA is studied to see how it compares with a classic SNA and how the SNA differs from a Vector Network Analyzer. Dr. George begins by looking at the components that make up an SNA and develops a generic block diagram. Next, a popular low cost commercial unit used for this study is reviewed. That unit covers the frequency range 0.05 MHz to 85 MHz. Finally, examples of measuring filters, crystals and standing wave ratio are presented.



In This Issue

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The American Radio Relay League

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The purpose of *QEX* is to:

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

2) document advanced technical work in the Amateur Radio field, and

3) support efforts to advance the state of the Amateur Radio art.

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and letters for publication in *QEX* should be marked Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted in word-processor format, if possible. We can redraw any figures as long as their content is clear.
Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX* or high-resolution digital images (300 dots per inch or higher at the printed size). Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Kazimierz "Kai" Siwiak, KE4PT

Perspectives

It's a Digital World: Continued

We quote and paraphrase the comments by Joe Kornowski, KB6IGK, Editor of the *AMSAT Journal*, from his July / August 2018 *AMSAT Journal* editorial.

"The use of digital modes in Amateur Radio has grown dramatically in recent years. One of the newest modes, FT8, developed by Joe Taylor, KIJT, and Steven Franke, K9AN, and incorporated in WSJT-X, has gained phenomenally rapid adoption within the Amateur Radio community. As of March 2018, Taylor reported that FT8 usage globally hovered around 15,000 users per week. Club Log reported that, by the end of 2017, FT8 represented more than half of the QSOs uploaded to Club Log. For the full year, of the 32 million QSOs uploaded, 5 million were FT8.

According to the ARRL, "For newcomers, data emissions are far more popular than telegraphy" (Petition to FCC for Rule Making, February 2018). Inevitably, computer generated data 'bits' have now overtaken 'dahs' and 'dits'.

This trend may be reflected in anecdotal data from the 2017 AMSAT journal readers' survey in which a couple of young aerospace engineers characterized the notion of real-time, two-way voice communication as passé, explaining that young engineers expect current technology to be digital and delay-tolerant. For them, working satellites in real-time simply requires too much planning and time commitment for their busy schedules. Having a digital station setup to remotely and automatically exchange data via satellites makes more sense to them."

We agree with Kornowski's editorial. Creating innovative ways to develop and operate digital communications might help inspire the next generation of Amateur Radio enthusiasts. We'd like to hear from you (**qex@arrl.org**), and we solicit your articles on Amateur Radio digital communication.

In This Issue

We feature a range of topics in this issue of *QEX*.

David Birnbaum, K2LYV, calculates the losses in reversible L network antenna tuners.

Dave Leeson, W6NL, shows how a dipole can be broadbanded by a number of techniques including by matching with resonant sections of transmission feed lines.

Jim Satterwhite, K4HJU, designs a test set to measure the competency of ground installations.

Dr. George R. Steber, WB9LVI, explores the capabilities of low-cost scalar network analyzers.

Andrzej (Andy) Przedpelski, KØABP, comments on high stability crystal oscillators.

Scott Roleson, KC7CJ, comments on handheld transceiver earpiece accessories.

Keep the full-length *QEX* articles flowing in, or share a **Technical Note** of several hundred words in length plus a figure or two. Let us know that your submission is intended as a **Note**. *QEX* is edited by Kazimierz "Kai" Siwiak, KE4PT, (**ksiwiak@arrl.org**) and is published bimonthly. *QEX* is a forum for the free exchange of ideas among communications experimenters. The content is driven by you, the reader and prospective author. The subscription rate (6 issues per year) in the United States is $29. First Class delivery in the US is available at an annual rate of $40. For international subscribers, including those in Canada and Mexico, *QEX* can be delivered by airmail for $35 annually. Subscribe today at **www.arrl.org/qex**.

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Very best regards,

Kazimierz "Kai" Siwiak, KE4PT

Dr. George R. Steber, WB9LVI
11105 N. Riverland Ct., Mequon, WI 53092; steber@execpc.com

Low Cost RF Scalar Network Analyzers

A very economical SNA unit covering 50 kHz to 85 MHz is put through its paces.

An RF network analyzer is an instrument for measuring electrical parameters of antennas, components, filters and more. Professional, full featured analyzers are quite expensive, precluding them from use by many experimenters. Today, simple low cost RF scalar analyzers, that provide many of the same functions, are becoming widely available. Kits and wired units provided through non-traditional channels, web stores or auction sites compete to make the prices amazingly low. So how do these analyzers stack up for Amateur Radio use? In this study a very economical unit covering the frequency range 0.05 MHz to 85 MHz is put through its paces. The results may surprise you.

Radio frequency designers consider the network analyzer to be one of the most valuable tools at their disposal. It is used to characterize or measure the response of devices at RF or even at microwave frequencies. Obtaining the response of the device or network makes it possible to understand how it will work within the circuit for which it is intended. Typically RF analyzers are used to measure a variety of devices ranging from antennas, filters, frequency sensitive components such as crystals, to devices such as transistors, amplifiers and mixers.

Two major instruments in the RF network analyzer family are the scalar network analyzer (SNA) and the vector network analyzer (VNA). The SNA is the simpler of the two measuring only the amplitude (scalar) properties of the components and devices. A VNA measures both amplitude and phase and generally can measure more parameters than a SNA. The capabilities and construction of VNAs are discussed in more detail in the Notes[1, 2]. The SNA, in spite of its limitation, can be very useful in many situations, and can be easier to setup and use.

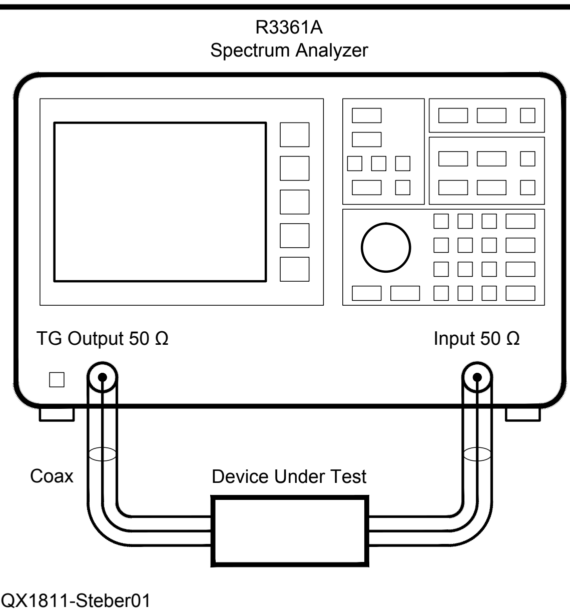

Figure 1 — Spectrum analyzer and tracking generator used as a classic scalar network analyzer.

The Scalar Network Analyzer

A spectrum analyzer in combination with a tracking generator functions as an SNA. The author's setup, see Figure 1, is typical and consists of a 2.6 GHz Advantest R3361C spectrum analyzer equipped with a tracking generator. When a tracking generator (TG) and spectrum analyzer (SA) are used together, their operation is closely linked.

The tracking generator produces a constant amplitude swept signal on exactly the same frequency that the spectrum analyzer is receiving within its resolution band-width filter (RBW). Thus, if the output from the TG is connected directly to the input of the spectrum analyzer, a line would be plotted across the screen of the analyzer indicating the amplitude of the tracking generator output versus frequency.

If a device under test (DUT) is placed between the TG and SA as shown in Figure 1, the spectrum analyzer will plot any amplitude variations caused by the device. Assume that a filter is the DUT. For this case the output signal of the tracking generator will pass into the filter, where its response will change according to the reaction of the filter at that frequency. In this way the spectrum analyzer is able to plot the response of the filter versus frequency.

Simple low-cost scalar network analyzers operate in a similar way but there are some important differences. The main disparity is that the frequency selective spectrum analyzer is usually replaced by a wide band multistage logarithmic amplifier like the AD8307. Hence there is no filtering of the input signal as in a spectrum analyzer. This provides simplicity at the expense of selectivity.

In this article a low cost SNA will be studied to see how it compares with the classic SNA described above. We'll begin by looking at its components and develop a generic block diagram. Next, a popular low cost commercial unit used for this study will be reviewed. Finally, examples of measuring filters and other components will be presented. To begin, let's take a closer look at typical low cost SNAs.

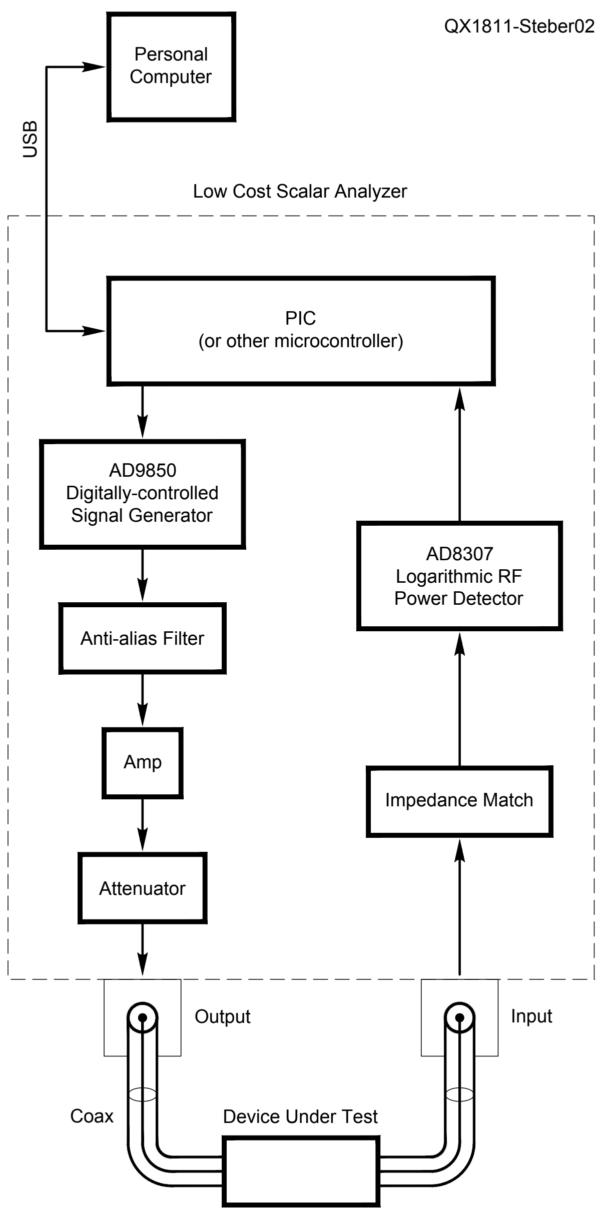

Figure 2 — Generic low cost scalar network analyzer block diagram.

Low Cost Scalar Network Analyzers

There are a number of low cost SNA designs and products on the internet. Some designs have been group efforts[3] where kits of parts are available and some are one-of-kind efforts[4] by individuals. Finished products can be found on auction sites or web shops. These units seem to have several things in common including heavy reliance on a personal computer (PC) for calculations and plotting. The main board of the unit typically has a microprocessor like a PIC which communicates with the PC. Also included is a programmable signal generator like the AD9850 and a sensitive RF detector, customarily the AD8307.

A block diagram of a simple SNA design is shown in Figure 2. Here the PC communicates with the SNA via the PIC and a USB port. The PIC is responsible for controlling the signal generator and collecting data from the RF power detector. The signal generator usually has an anti-alias filter to clean up the signal, followed by an amplifier and attenuator. The attenuator provides a reasonably stable 50 Ω termination for the amplifier and load. On the receive side, the input of the AD8307 has an impedance matching circuit to provide a 50 Ω input impedance. It is generally placed in a shielded box to prevent extraneous signals from reaching it. This is very important since the sensitive AD8307 will process any and all signals received and this will raise the noise floor.

A family of NWT scalar analyzers similar to the generic version just described can be found on the internet. To my knowledge they are based on the plans of our German radio colleague Bernd Kernbaum, DK3WX. The first design was published in the magazine *Funkamateur* around 2000. Bernd offered kits with the project and it was apparently successful. Kits were ended around 2004, but it continues to be supported today by other individuals with improvements and software. Incidentally, the name NWT stands for NetWerkTester. According to my German translation, Bernd originally described the NWT as a measuring device for passive and active filters, antennas, and finding the values of inductors and capacitors. He also describes its use as a precision oscillator, a local oscillator for a receiver, and with an external mixer as a spectrum analyzer.

One particular device, the NWT70 found on the internet, caught my attention. It covers the frequency range of 0.05 MHz to 85 MHz. The package, see Figure 3, includes the NWT70 unit, two SMA cables, USB cable, three external attenuators (0 dB, 6 dB, 40 dB), and a CD with software, manuals and other things. Because it is USB powered no extra power supply is needed. The price was unbelievably low, only about $68 with free shipping.

While pondering this unit another one came to my attention, the NWT500. Its frequency range is 0.1 MHz to 550 MHz and includes the same items as the NWT70 plus a power adapter, since it is not powered by the USB port. Its cost was $127.77. Yet another popular unit is the NWT4000, which goes to 4.4 GHz, but at a higher cost. Unlike the NWT70 it uses a classical heterodyne concept (spectrum analyzer with tracking generator). More searching found eight different models of NWT hardware with different frequency ranges and options — quite a selection of analyzers.

Since the NWT70 had initially sparked my interest, I ordered it. Hopefully this low cost unit would satisfy my curiosity. A few weeks later it arrived in good shape. In the next section we'll examine the NWT70 in more detail.

The NWT70 Scalar Analyzer

The NWT70 is housed in a compact 4" by 2.5" by 0.875" metal case, and is well fitted. Two SMA connectors and a power-on LED are on the front panel. The rear panel has one slot for the USB connector.

Removing the front screws allows access to the main board. It is nicely laid out using all surface mount parts. Figure 4 shows the top side of the PCB. There is a metal shield over the output circuit to reduce noise, as seen in the lower left corner. Figure 5 shows the bottom of the board. An Atmel MEGA 328P microcontroller, AD9851 signal generator, and an FT232RL USB-to-serial UART were identified using a magnifying glass. There is a metal shield over the input circuit to reduce noise, seen in lower left corner. The Chinese manufacturer of the unit is BG7TBL, as noted in the lower right corner. Overall impression is that the board is well designed and made.

Based on these observations it seems likely that this unit is very much like the generic one shown in Figure 2. A similar block diagram (in Chinese) is in the user manual. There is a nice added feature with this unit: an internal programmable attenuator going from 0 to 50 dB in 10 dB steps. This should be handy when calibrating the unit with the software.

When the NWT70 was plugged into a USB port on a Windows 7 laptop, the FT232RL driver was immediately found and installed. Similarly, when it was plugged into a USB port on a WinXP desktop PC the driver was found. This is probably because

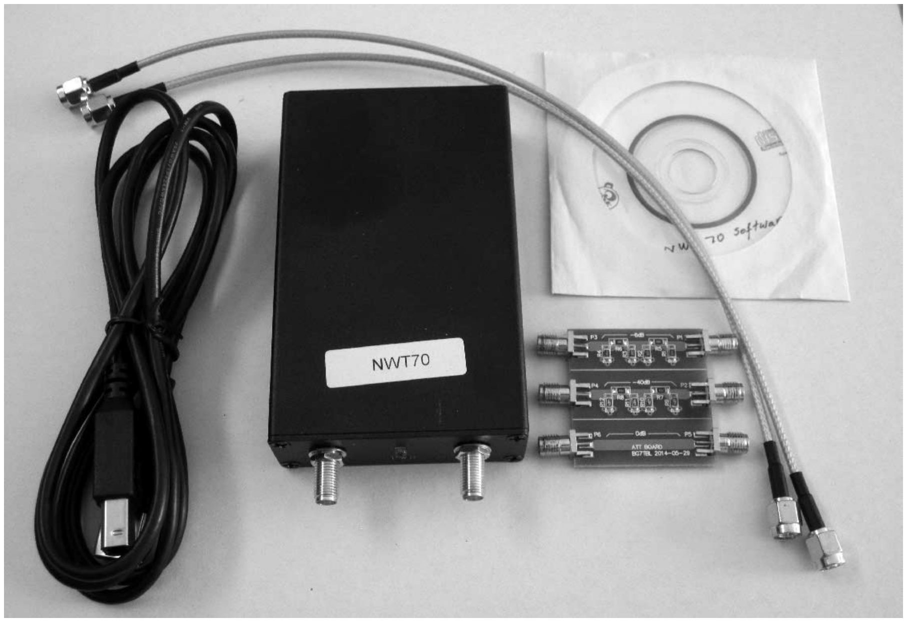

Figure 3—The NWT70 unit and other items received by internet purchase. [George R. Steber, WB9LVI, photo.]

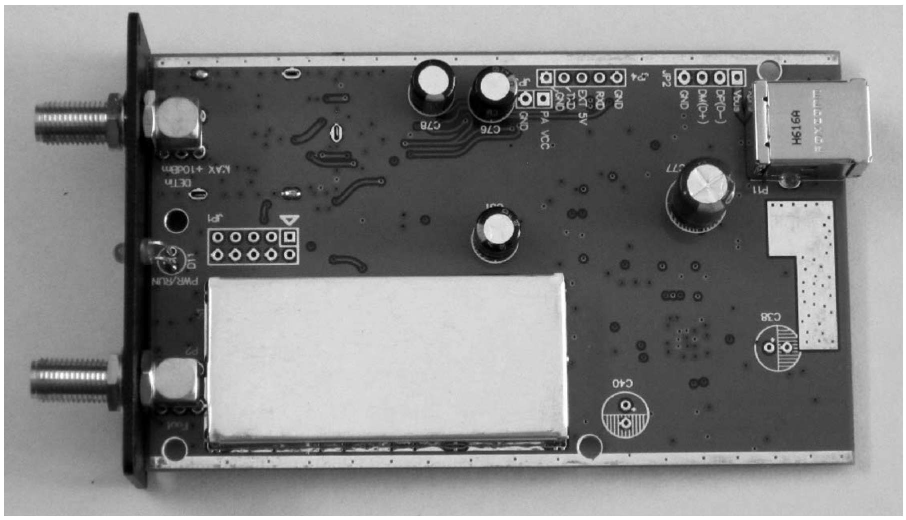

Figure 4 —Top side of NWT70 circuit board. [George R. Steber, WB9LVI, photo.]

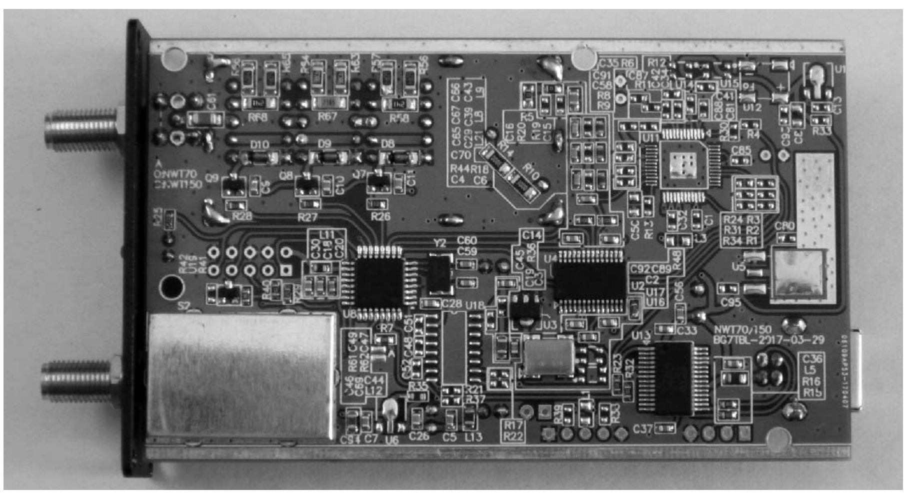

Figure 5 — Bottom side of NWT70 circuit board. [George R. Steber, WB9LVI, photo.]

Table 1.

NWT70 Frequency (MHz) measurements for different bands.

Set:	1.8000000	3.6000000	7.000000	14.000000	28.000000
Measured:	1.8000003	3.6000006	7.000011	14.000022	28.000046

Table 2.

Second harmonic content of NWT70 signal generator for different bands.

Set Frequency (MHz):	1.8	3.6	7.0	14.0	28.0
Second harmonic (dB):	-31.6	-30.6	-29.9	-31.0	-35.4

Table 3.

Power (dBm) at various frequencies measured by NWT70 from 0 dBm output of Marconi 2022D.

Frequency (MHz):	0.05	1.0	5.0	10.0	20.0	30.0	40.0	50.0	90.0
NWT70 (dBm):	0.7	0.3	0.3	0.3	0.1	-0.2	0.0	-0.2	-0.8

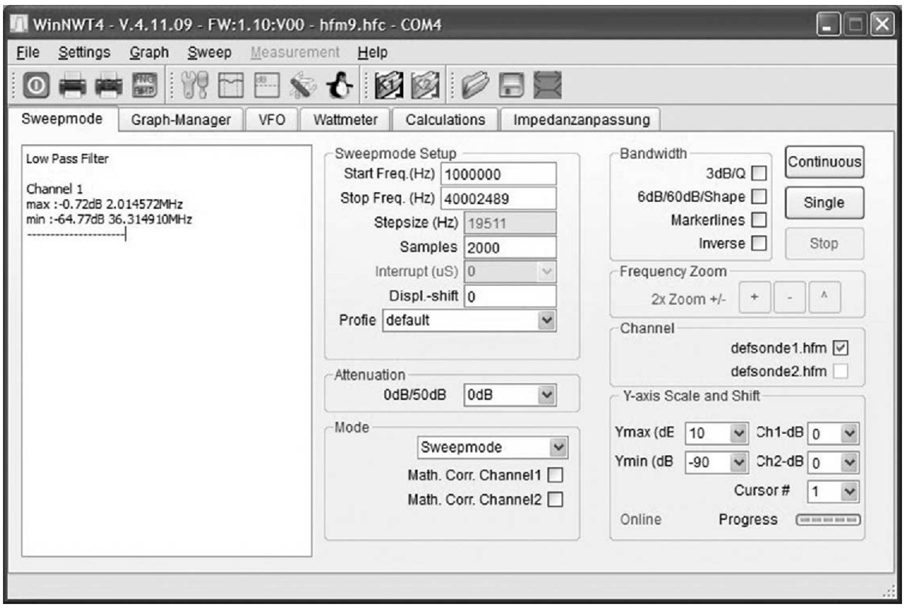

Figure 6 — NWT70 Sweepmode screen.

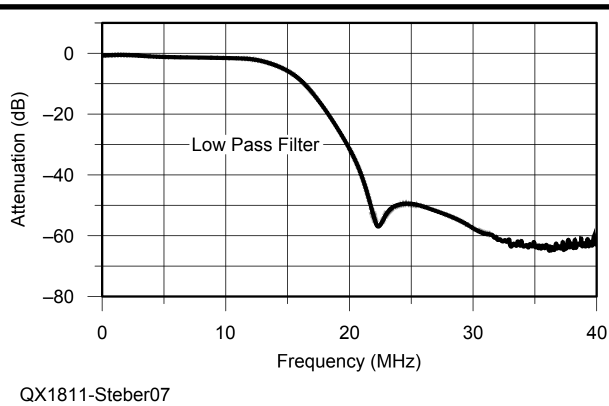

Figure 7 — NWT70 Graphical screen showing low pass filter response.

there were other USB devices installed using this driver on these PCs. The driver on the CD will have to be installed if it is not on the computer. When the PC recognizes the NWT70, it installs a COM port on the computer. A specific COM number is required by the software and can be found by looking in the Windows Device Manager under Ports. Next step was to find and install the software.

NWT70 Software

There are nine folders on the CD. Of these, the NWT folder has the material for the NWT sweep analyzer units. The remaining eight folders are for other products from the company. The NWT folder has the USB drivers, PC software, user manual, application manual, calibration notes for the NWT microwave units, and Linux and Mac-OS software. These manuals are written in a combination of Chinese, English and sometimes German, and are challenging to follow and interpret.

All NWT models are covered under the included software *WinNWT4, V4.11.09* for Windows. *WinNT4* installed under WinXP without problems. Andreas Lindenau, DL4JAL, another German radio colleague, wrote the software. Andreas has also written a document *LinNWT*, available on the web, which describes the software.

For the program to function, the COM port number that was created when Windows installed the port needs to be entered in *WinNWT4* under Settings and Options. The program is then ready to operate. Although fairly easy to use it is complicated by the fact that it covers all NWT units, not just the NWT70. So you must learn to ignore certain features.

The Sweepmode screen is shown in Figure 6. There is also a separate graphical screen to show the results of a scan, see Figure 7. There are five other screens covering Graph management, VFO operation, Wattmeter, Calculations, and Impedance matching. But Sweepmode is the most valuable since it controls the initiation of a scan, the start and stop frequencies, number of samples, mode, and decibel scaling, to mention a few.

Plunging ahead without reading the manual, a low-pass filter was connected between the OUT and IN Ports on the unit and the button labeled 'Single' in Sweepmode was clicked. A nice graph of the filter response was plotted on the graphical screen as shown in Figure 7. That was too easy! With that incentive, let's find out how well this unit really performs.

NWT70 Checkout

A frequency accuracy check was done

using a HP 5315A frequency counter. Table 1 shows set and measured frequencies in MHz for several bands. For this unit, the error rate is approximately 1.666 Hz per MHz. There is a procedure in the software to calibrate the signal generator frequency, but it was not done at this time.

Harmonics of the device's signal generator were checked with the R3361A spectrum analyzer for several bands. Levels of the second harmonic are shown in Table 2.

Harmonics in a tracking signal generator are not a problem with the classic SNA that uses a narrow RBW tracking filter in the spectrum analyzer. But for the NWT70 they will show up as errors in DUT response. In other words, when you set the frequency to say, 1.8 MHz, there is also another signal (harmonic) at 3.6 MHz, albeit -31.6 dB lower, that contributes to the response. Similarly, many of the other harmonics are large enough that they will likewise contribute to the error. We will see consequences of these harmonics in a band-pass filter scan later on.

The input and output impedances of the NWT70 were measured up to 90 MHz. They are shown in Figure 8. The output impedance varies a bit over this frequency range. This should not be much of a problem if the device under test has an input impedance of 50 Ω. Otherwise a 6 dB (or more) attenuator at the output is advised in the interest of accuracy, even though the dynamic range will be reduced.

The input impedance of the NWT70 is close to 50 Ω over the range. This is due to the fact that a simple 51 Ω resistor to ground is used at the input of the AD8307.

The power measurement capabilities of the NWT70 were tested using a standard signal generator, a Marconi 2022D. Power level was set to 0 dBm. Note that it's possible there may be some variation in the 2022D output. Results over the unit's range are shown in Table 3.

It was not clear if the power scale could be calibrated in the software. Other levels of power were tested, and the scale was found to be linear from +10 dBm down to -60 dBm. Note that if the unit is used as a power meter, the maximum input is 10 dBm. Using higher power levels, for example when testing a power amplifier, will require an external attenuator to avoid damage.

Sweep calibration is performed by clicking Sweep and Channel 1 Calibration. It is a simple procedure that requires placing a short between the output and input connectors. The process is easy to follow and takes only a minute or so. It is particularly convenient since it uses the built-in 40 dB attenuator, so there is no changing of external attenuators required.

After calibration, tests were run using a

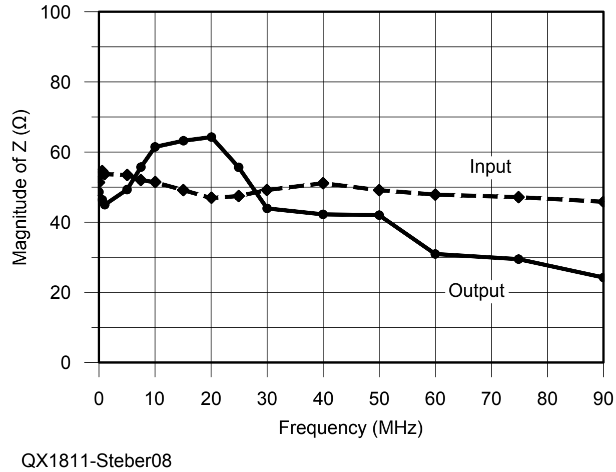

Figure 8— Input and output impedance magnitudes of the NWT70.

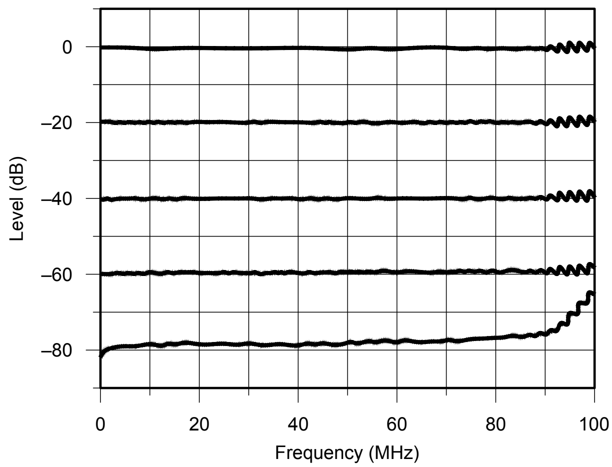

Figure 9— Gain plots using Kay 432D attenuator at various settings. Top plot is 0 dB and decreases in 20 dB steps. Sweep from 1 MHz to 90 MHz.

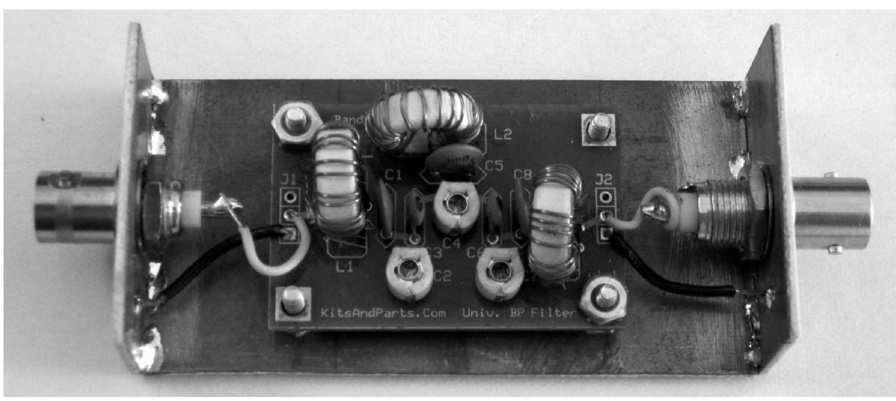

Figure 10 — 14 MHz band-pass filter, shown with the top cover removed. [George R. Steber, WB9LVI, photo.]

Kay 432D attenuator as the DUT. Frequency scans from 1 MHz to 90 MHz were plotted at 20 dB intervals as shown in Figure 9. The results show a fairly respectable dynamic range with the lines being flat and the noise floor intruding around -75 dB.

When testing components like filters, crystals, and other devices, it important that the unit be normalized (calibrated) to include the effects of the fixture and connecting cables. In other words, perform the calibration with exactly the same test setup used for measuring the DUT but with the DUT replaced with a short. This has been done for all the subsequent tests.

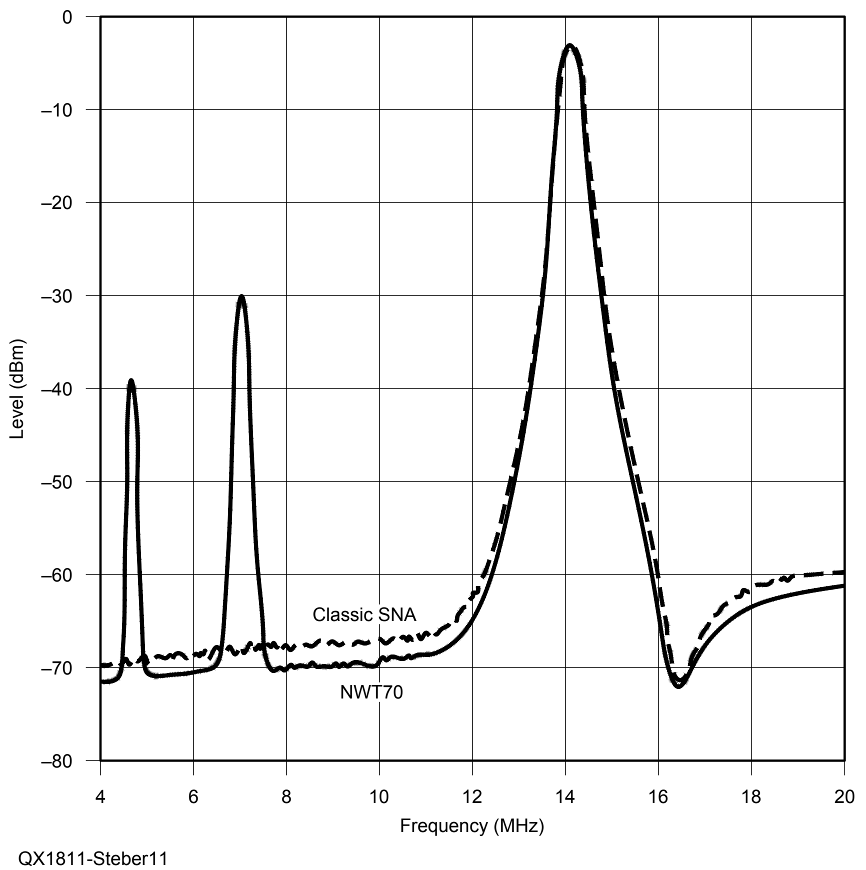

Figure 11 — Frequency responses of 14 MHz band-pass filter comparing classic SNA to simple SNA. See text for more information.

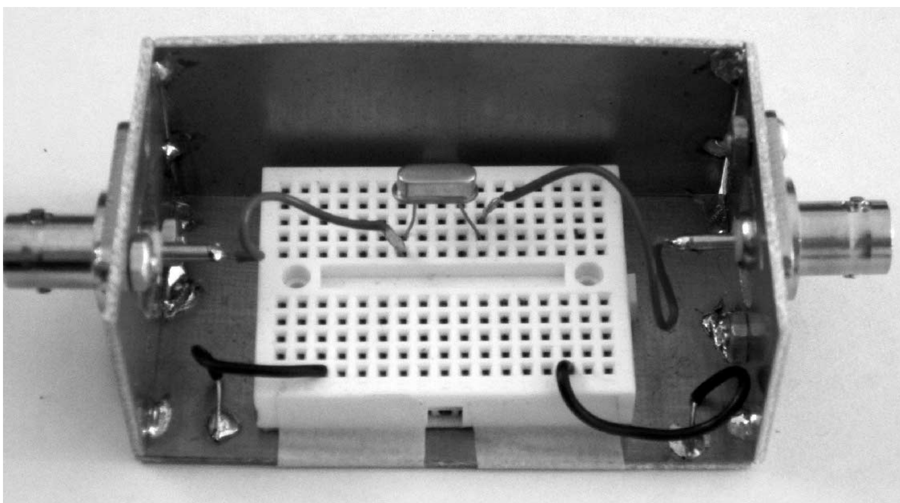

Figure 12 — Test fixture for crystals and other components shown here with top cover removed. [George R. Steber, WB9LVI, photo.]

Testing Filters

A 14 MHz band-pass filter that was tested is shown in Figure 10. Its frequency response obtained with the NWT70 is shown in Figure 11 together with the response obtained using the classic SNA mentioned earlier. The plot with the peaks on the left side is from the NWT70. Those peaks are due to the harmonic responses of the NWT70 signal generator as discussed in the previous section. They are not present in the classic SNA plot. In this case we see the response of the filter due to the first and second harmonics being placed lower in frequency at the ½ and ⅓ frequencies. Otherwise the curves are similar.

Low-pass filters would not be expected to show such extreme errors since the harmonic response would be taken from the filter stop band and their contribution would be small. High-pass filters suffer more since the harmonic response is taken from the filter pass band and is placed in the filter stop band indicating less attenuation.

Testing Crystals

Checking crystals is very easy with the NWT70. It does not suffer as much from the harmonic problem mentioned earlier. A simple test fixture for testing crystals and other devices is shown in Figure 12. It is not completely shielded as it was used for the classic SNA which selectively filters out off-frequency noise. When used with the NWT70 it should be well shielded to avoid raising the noise floor. An example of a crystal measurement is shown in Figure 13. A 30 m band crystal was used here. The classic SNA produced an almost identical plot

Using an SWR Bridge

An SWR bridge can be used with most SNAs for testing antennas and measuring impedance. A Mini-Circuits ZDC-20-3 coupler was used here, as shown in Figure 14. It was setup with the ZDC-20-3 coupled port connected to the SNA Input. The other ports of the ZDC device were connected to the SNA Output and Load. The NWT70 software includes modes for SWR and impedance testing with a coupler. When used for impedance measuring a separate SWR calibration is required.

To verify SWR operation, an antenna model consisting of a series *RLC* circuit ($R = 50\ \Omega$, $L = 6.5\ \mu H$ and $C = 21.1$ pF) was constructed. This is a good way to test the analyzer without using an actual antenna. The resulting SWR plot is shown in Figure 15. The impedance measuring mode of the analyzer works in a similar way. Impedance should be in the range of 20 to 150 Ω for reasonable accuracy.

Summary and Conclusions

This article discussed the differences between a classic SNA and a simple SNA design commonly found on the internet. The NWT70 SNA used in this study, although very low cost, produced results that were surprisingly good. It was found, however, that because of the extreme sensitivity of the AD8307 RF detector input to all signals, the DUT must be very well shielded to achieve a large dynamic range. Placing the DUT in a metal case solves this problem.

A more subtle problem of the NWT70 and its ilk is the lack of selectivity, which can produce errors from the device's signal generator harmonics when measuring band-pass and high-pass filters. That said, in many situations, the NWT70 performed adequately for finding low pass filter responses. It also worked well for finding crystal frequencies and would probably be acceptable for measuring cable attenuation and tuning stubs.

The ability to measure power in dBm is a nice feature. Power attenuators would be required to put the signal in the safe range for the measurement of amplifiers and transmitters. Realize though that the total power measured will include harmonics, unlike a spectrum analyzer.

An SWR bridge can easily be connected to the NWT70 and used for antenna tuning and impedance measurement. The impedance measurement range is limited and there are other instruments on the market that may work better if a large impedance range is needed.

Although much of the software for the NWT70 was in English it was not always clear how to perform certain operations. That said, most commands in German are easy to translate. For instance, *'Abbruch'* means *'Cancel'* and *'Bild spechern'* means *'Save image'*. The written manuals are another matter.

My college German language skills are very rusty, so the written manuals were very challenging for me to read. That said, it should not pose a problem as most people skilled in this art will probably not use the manual.

While the simple analyzer worked reasonably well, the classic SNA analyzer using a tracking generator is more accurate, less critical to use, and has a larger dynamic range without having to worry about the issues mentioned here. Of course the classic implementation costs quite a bit more than the $68 for the simple version tested here.

Experimenting with this simple SNA was

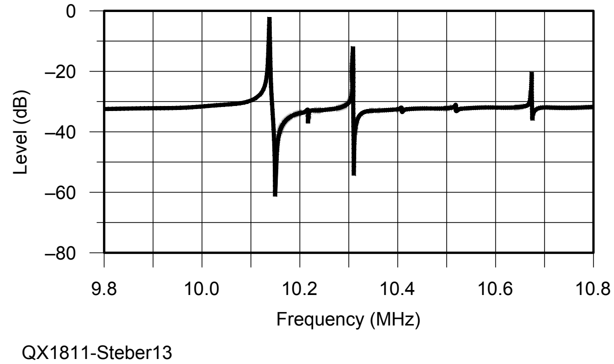

Figure 13 — Frequency response of WSPR crystal. The maximum value is -2.06 dB at 10.138000 MHz, and the minimum value is -61.52 dB at 10.150000 MHz.

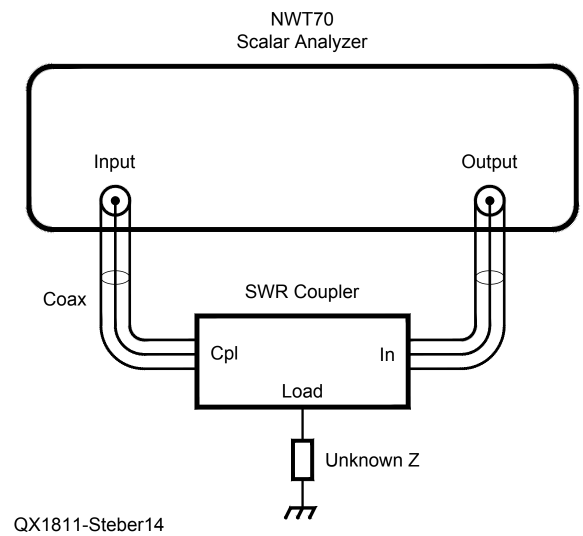

Figure 14 — The NWT70 connected to a Mini-Circuits ZDC-20-3 SWR coupler can be used to measure SWR.

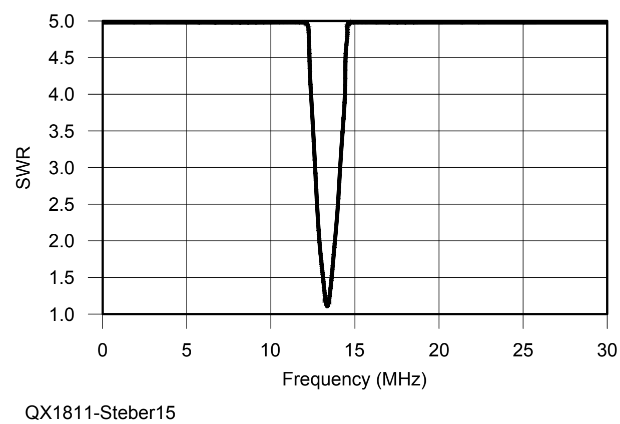

Figure 15 — SWR plot of an antenna simulated by an *RLC* network is resonant at 13.6 MHz.

enjoyable and educational. Hopefully you have learned something about its strengths, weaknesses and applications. An SNA is a nice general purpose instrument for the advanced radio amateur. With any luck you will start measuring with your scalar network analyzer soon.

George R. Steber, PhD, is Emeritus Professor of Electrical Engineering and Computer Science at the University of Wisconsin-Milwaukee. He is now semi-retired, having served over 35 years. George, WB9LVI, has an Advanced class license, is a life member of ARRL and IEEE and is a Professional Engineer. His last article in QEX , "Using a Wide-Band Noise Generator with a Spectrum Analyzer", appeared in the May/June 2016 issue. George has worked for NASA and the USAF and still lectures on various topics at the University. He is currently involved in cosmic ray research and is developing methods to study them on a global basis. When not dodging protons, pions and muons, he enjoys WSPR/JT9 Amateur Radio, racquetball, astronomy, and jazz. Contact George at ***steber@execpc.com*** *with "SNA" in subject line and email mode set to text.*

Notes

[1]G. Steber, "An Unusual Vector Network Analyzer", *QEX*, Sept./Oct., 2007, pp. 11-23.
[2]T. C. Baier, "A Small, Simple, USB-Powered Vector Network Analyzer Covering 1 kHz to 1.3 GHz", *QEX*, Jan./Feb., 2009, pp. 32-36.
[3]Poor Ham's Scalar Network Analyzer, **https://groups.io/g/PHSNA**.
[4]G. Richardson, "An RF Filter Evaluation Tool", *QEX*, July/Aug., 2014, pp. 3-6.

Letters

Satellite Distance and Velocity, (July/August 2018)

Dear Editor,

I found the article by Andy Przedpelski, KØABP, to be quite interesting and inspiring. It is intuitively obvious that velocity can be determined from Doppler measurements, but I had never considered that distance could be determined as well.

The article inspired me to carefully derive the article's equations from basic principles. This exercise uncovered several typos. First, the curves in Figure 1 are mislabeled. 'ABCD' should be labeled in reverse order, 'DCBA'. Next, equation (6) should read,

$$D = \frac{V^2}{\lambda\left[df/dt\right]_{\max}}$$

which now has the correct units. The example that follows equation (6) incorrectly converts 165 m to 363 ft (it should be 541 ft). Finally the units for satellite distance should obviously be kilometers, not meters.

My derivations of Andy's missile miss distance indicator equations revealed that the results are valid for a non-accelerating target traveling on a straight line. While this is obviously not true for an orbiting satellite, it is apparently not a bad approximation for the small observation period of 7 minutes.

I wondered if using the equations for apparent satellite motion would improve the 10% accuracy found in the article. A quick search of the internet turned up the equation for the distance between an earth station and a satellite in a circular orbit. Using the 'search' function in MS Excel to curve fit the article's Doppler measurements, the velocity and distance were determined to less than 0.5% error! This technique also had additional advantage that orbital radius and period were determined to similar accuracy.— *73, John Gibbs, NN7F,* ***ja_gibbs@hotmail.com***.

The Author Responds,

Thanks to John Gibbs, NN7F, I have learned my lesson: check the units, check the math, check the final version. As John correctly pointed out, the Figure 1 curves top to bottom should be D, C, B, A. Equation (6) is also correctly given by John, NN7F. This error was also caught by Herman Birkner, W2FRH. Yes, the m to ft conversion is wrong, and the satellite distance should obviously be in kilometers.

John was too generous calling them typos. They were really my errors! I do have the equations for an orbital path, but they get quite messy and I thought this was enough for somebody to give it a try.— *73, Andy Przedpelski, KØABP,* ***kc0cwk@comcast.net***.

Q - Factor Formulas, (July/August 2018)

Dear Editor,

Please pass on my congratulations to Kai Siwiak, KE4PT, for his excellent technical note in July/August 2018 *QEX*. In my 45+ years career as an electrical engineer, I have never seen such a comprehensive, succinct and well written article on this often misunderstood subject. Well done! *— Cheers, George Georgevits, VK2KGG,* ***georgg@bigpond.net.au***.

Send your *QEX* Letters to the Editor to, ARRL, 225 Main St., Newington, CT 06111, or by fax at 860-594-0259, or via e-mail to qex@arrl.org. We reserve the right to edit your letter for clarity, and to fit in the available page space. "Letters to the Editor" may also appear in other ARRL media. The publishers of *QEX* assume no responsibilities for statements made by correspondents.

Jim Satterwhite, K4HJU

50 Woodcreek Dr., Wimberley, TX 78676; **k4hju@arrl.net**

Ground Ohms Test Set

Measure installed grounding and bonding resistances, as well as earth resistivity.

When first conceived, this article was intended to describe the details of the test set that I designed and built. However, when the project was essentially complete, there was something I discovered that changed the emphasis of this article. I discovered that many Amateur Radio operators do not want to know the competency of their grounds. Therefore this article will be directed largely about the need to measure the resistance of installed grounds.

I am not an expert in the area of grounding. I do know some parts and I have significant experience in the design of successful test and measurement equipment. After many years of thinking about the methods of design for measuring the resistance of Amateur Radio station grounding, I decided to do something about it. From my experience working with telephone outside plant testing instrumentation, I knew that grounding measurements were not trivial. There are complicating factors, not the least of which are ac and dc currents flowing in the earth.

My interest came to a head when I moved into my present home and had a really difficult time achieving what I would consider good grounding, due to the mostly solid limestone just inches below the surface over most of the property. I wanted to know the condition of my grounding. There are a number of competent clamp-on instruments for measuring ground resistance of installations, however they were outside my budget, and I wanted the experience of designing and building such an instrument.

I find it a bit frustrating and somewhat amusing that a number of references on grounding and bonding discuss in detail methods of choosing and installing such systems. I see few discussions of methods to verify that the installation is indeed competent. I suspect that this is due to the difficulty involved in achieving valid measurements. This test set is my attempt to measure such systems competently.

I had a surprise when I got the test set completed, and put the word out in the local ham community that I would be willing to measure their installed grounds. I wanted to do this to verify the test set for myself on a variety of situations. There were and still are no takers. I began to suspect that there was something deeper going on. As much as hams want to measure and quantify things, I wondered why not measure installed grounds. I will address this further below.

While this instrument is capable of the measurement of earth resistivity, the emphasis here will be on measuring installed grounding and bonding resistances.

The Problem

We spend a good bit of time and effort designing and installing grounds for our stations.[1,2] Every situation is different. The very nature of our hobby exposes us to lightning events[3] that can cause considerable damage and potentially loss of life. Therefore, it is important that we apply due diligence to proper grounding and bonding. There is very little discussion about verifying the performance of the grounds that we install. There are some very good references and techniques for designing and installing good grounds. Some installations present difficult challenges. For example, here in central Texas, the presence of solid limestone just inches below the soil surface makes installing ground rods and ground systems particularly difficult and potentially expensive. It seems to me that verification of the installed ground is essential to completing and maintaining the ground system.

As pointed out by some, there are factors of concern that a ground resistance measurement will not tell you. An example is the inductance of a tower and the various transmission lines involved. While this is true; the ground system can be only as good as the dc resistance permits. That is the bottom line. There are a number of issues such as inductance that require due diligence in the design and installation. A low ground resistance is essential. The other factors will not matter if the dc ground is inadequate. Therefore it is very important to verify the dc ground integrity.

In my estimation, the accuracy of the measurement is not as important as getting a good reading in the vicinity of the actual value. I bring this up because of the situation I ran into with the clamp-on induction coils. For high accuracy the ferrite core halves must consistently mate very tightly around the measured conductor. As I will discuss below, due to the way I wound the wires on the clamp-on ferrite core, getting a good mating

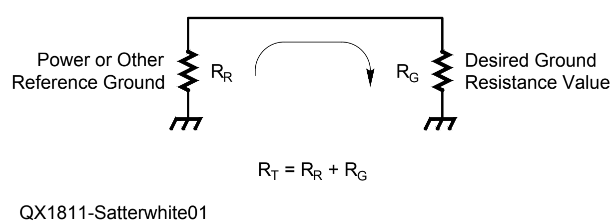

Figure 1 — Measurable ground loop.

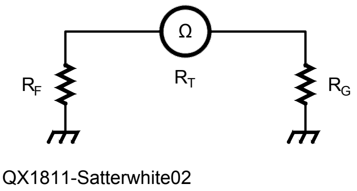

Figure 2 — Measurement using a simple ohm meter.

between the core halves is problematic and affects the accuracy of the measurement.

Soil resistance behavior is anything but uniform and varies significantly with the type of soil, moisture and dissolved salt content. I suspect that it can be non-linear. It can vary from day to day and from month to month.

The competency of the ground system is subject to change over time and environmental conditions. In my opinion the maintenance of the ground system requires periodic inspection and measurement.

The Measurement

When making this measurement, see Figure 1, you are always measuring a loop consisting of some sort of reference resistance and the desired unknown ground resistance. You must know something about the reference ground to infer the value of the desired ground resistance. There are established techniques for using other ground rods in specific locations relative to the primary ground, making resistance measurements and applying an algorithm to those measurements to yield the desired installed ground resistance value. Sometimes this is inconvenient if not impossible. I suggest that there is an alternative: use the power utility system ground as a reference.

Measurement Methods

There are several methods available to measure the resistance of the installed ground. Each has it's advantages and disadvantages.

Ohm meter measurement

The installed ground resistance can be measured with an ordinary ohm meter as seen in Figure 2, however there are some significant considerations. The first is that the ground must be taken loose to insert the meter to make the measurement. Many times this is quite inconvenient. The second is that there is a good possibility that there are dc and/or ac currents flowing in the ground system due to earth currents and power line induction. Even small dc currents will affect the resistance measurement. The power line induction currents can in some cases be dangerous. I have measured as much as 15 amperes flowing in a telephone cable sheath. However, in most cases it is much less than that depending on the ground location relative to power lines. Yet, it is still a concern for reliable measurement.

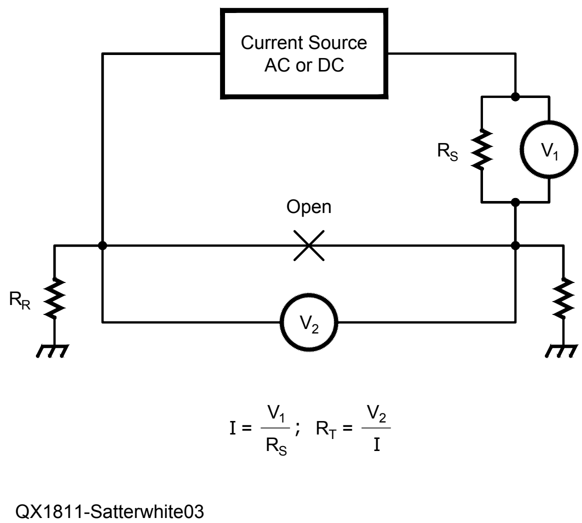

Figure 3 — Measurement using a four-terminal measurement system.

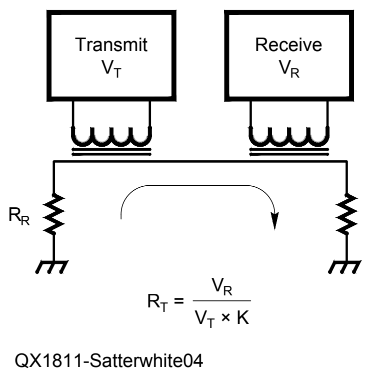

Figure 4 — Measurement using inductive coupling.

Four-terminal ac or dc measurement

In most cases the four terminal ac or dc resistance measurement, see Figure 3, can overcome the problems due to ac or dc earth currents if the measurement configuration is set up properly, and the applied current is significantly higher than the potential earth currents. It still has the disadvantage that the ground connection must be opened to make the measurement and calculations must be made to know the measured resistance. I suppose that it would be easy enough to build a test set around this type of measurement that would make the calculations. That said, if accomplished properly, the four-terminal measurement is probably the most accurate method, since the applied current can be sufficient to make the effects of the potential earth currents insignificant.

Inductively coupled ac measurement

In a good many cases, an inductively coupled ac measurement does not require opening the ground circuit, see Figure 4. Also, the effects of dc and ac earth currents are significantly minimized. It is a simple and convenient method of making this measurement. In this case an ac current is induced into the loop formed by two independent grounds, the connecting conductor and the earth. The resulting current in the conductor is sampled by a second clamp-on inductor. The resulting sampled current along with the voltage across the transmit clamp-on inductor are then analyzed to display the measured resistance.

Design Approach

I wanted the test set to measure the installed ground resistance under general field conditions that might include ac and dc earth currents. Also, I wanted to use inductive coupling using easily available snap-on ferrite modules. My system development technique generally uses development of a Windows application to display data and manage the various calculations required in addition to communication to the test set via USB-RS232. In the test set communication, control and data measurement are managed by an Arduino-like microcontroller module.

The measurement hardware performs the desired function and presents data to the microcontroller for analog to digital conversion. A computer provides a display mechanism and tools applications for the development of the Windows application and the microcontroller firmware. The various electronic functions are implemented using functional surface mount modules and insulation displacement wiring. This way the development environment is totally fluid allowing the design process to evolve.

On some systems, once the development is complete, the computer becomes superfluous and the developed system can operate on its own. On this system the netbook or laptop computer is clumsy to use in the field, particularly in bright sunlight. However, it is needed to analyze the data and calculate the results. It might be possible to perform these calculations in the microcontroller, however I have put no effort into that. Ideally, at some point I might replace the computer with an internal Raspberry Pi.

The functional modules, Figure 5, are designed and laid out in Cadsoft Eagle. Insulation displacement wiring, Figure 6, allows flexibility and reliability in construction and maintenance and also the ease of keeping documentation current. Between the insulation displacement wiring and the method of creating a ground plane that I employ, and with care, circuits handling near 100 MHz can be implemented. I say this to give some idea of the viability of using this type of construction.

I wanted the system to make measurements with reasonable accuracy and repeatability over a wide range of resistance values. It turns out that this required a good bit of attention to detail.

Hardware

Figures 7 and 8 show top and bottom views respectively of the test set I designed and built. It consists of an instrumentation box, two clamp-on inductive transducers and a small computer such as a notebook or netbook connected via USB. The USB supplies communication and power for the unit. I decided early on to use surface mount devices (SMD), functional modules, and insulation displacement wiring to develop the test set. Also, I choose to develop a Windows application in National Instruments Lab Windows to manage the operation and analyze and present the data.

Figures 9 and 10 show the front and rear panel views of my test set. Figure 11 shows the clamp-on inductive transducers. The

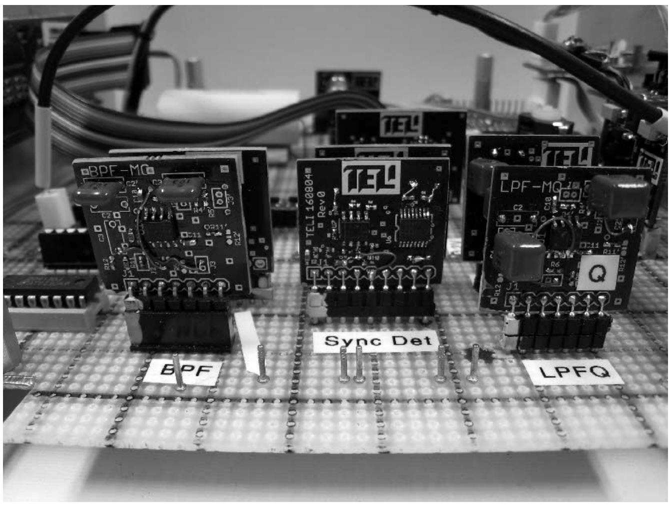

Figure 5 — SMD function modules.

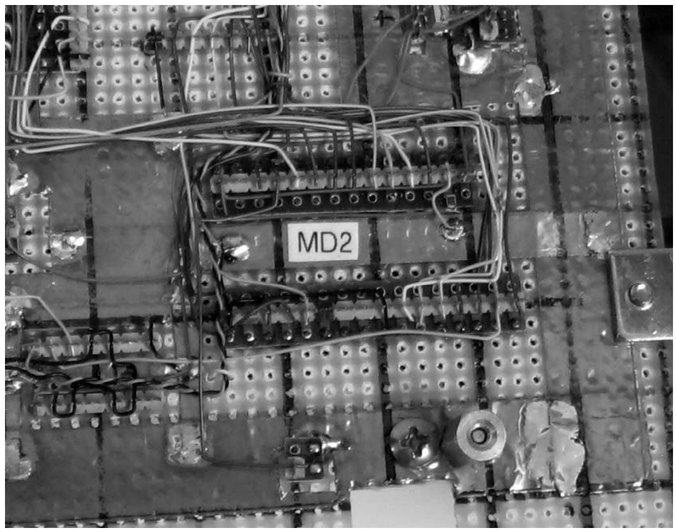

Figure 6 — Insulation displacement wiring.

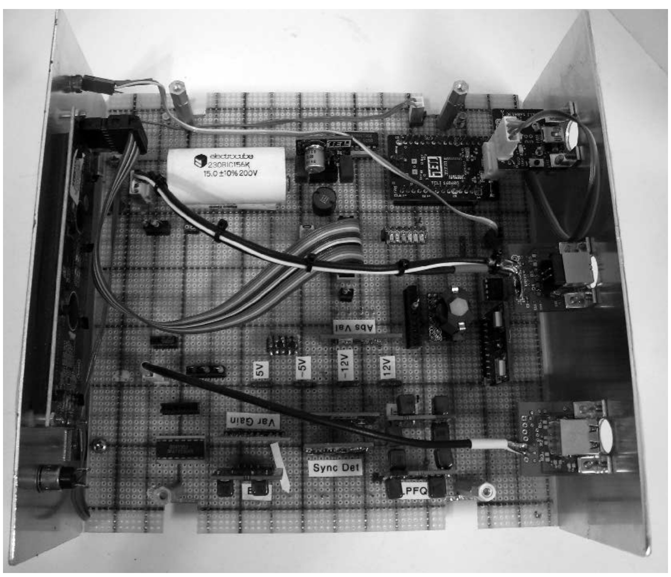

Figure 7 — View of top side.

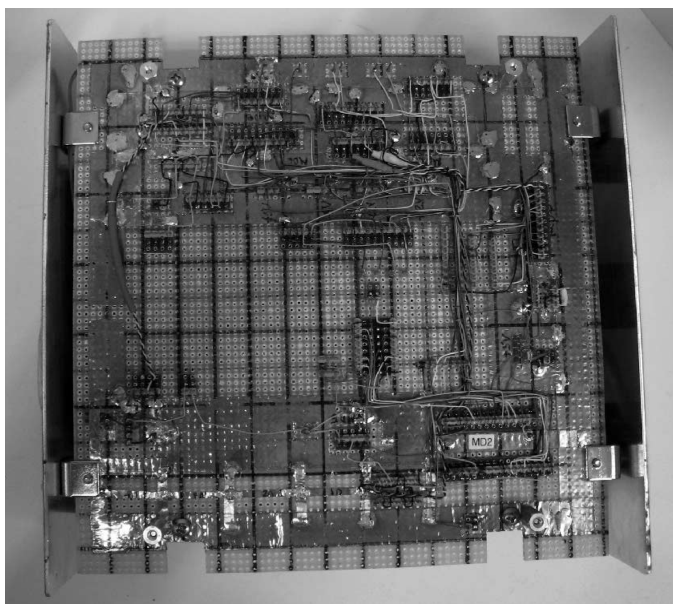

Figure 8 — View of bottom side.

transmit and receive transducers are made by winding turns onto clamp-on ferrite cores. I found that the windings interfere somewhat with the molded-in springs in the clamp-on case. This means that mating of the core halves is not consistent. This can be remedied by using an external clamp to mate the halves more precisely. Without the external clamp, this potentially introduces a variation in the resistance readings of up to 8% from the rather repeatable accuracy of 2% over most of the test set's range. To address this I am considering using 3D printing to make my own core shells where the windings do not interfere with the clamping action.

Figure 9 — Front panel view.

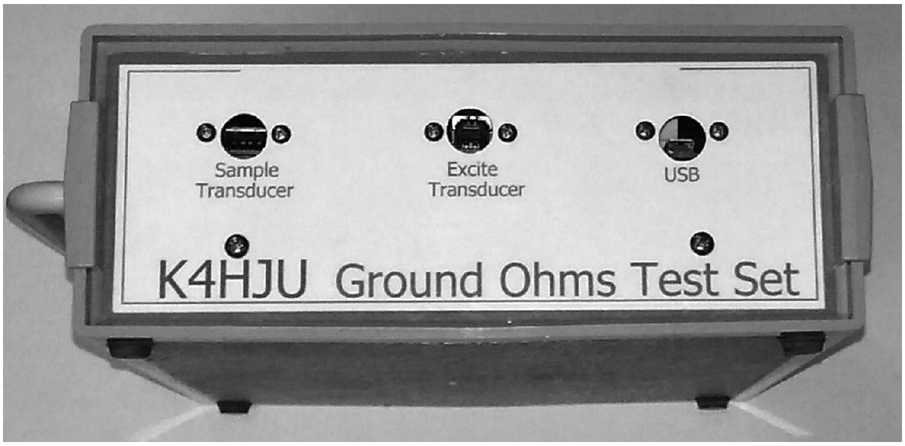

Figure 10 — Rear panel view.

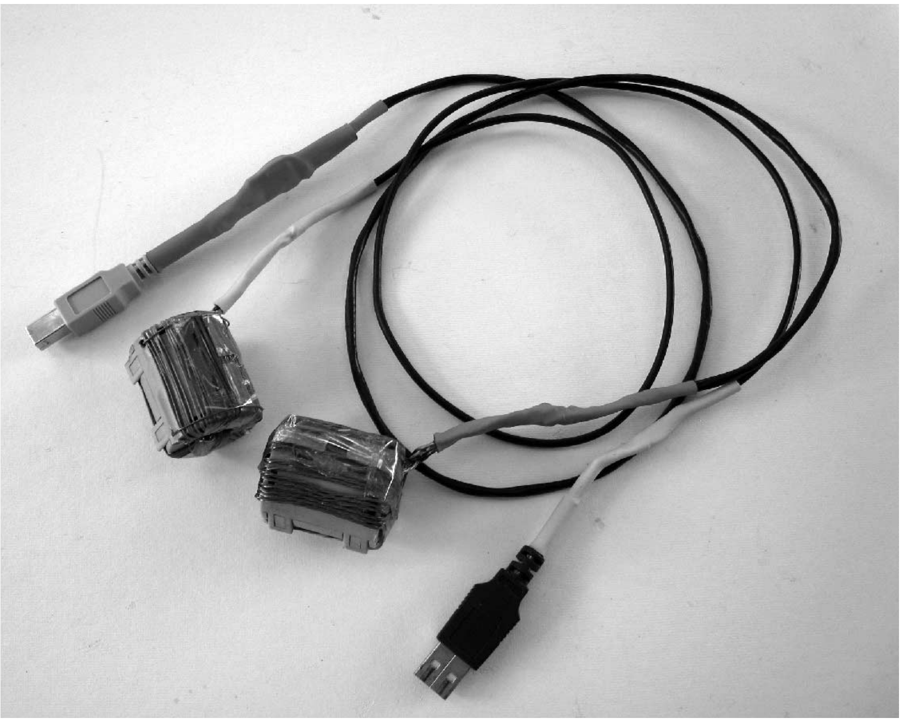

Figure 11 — Clamp-on inductive transducers.

Functional Description

Figure 12 shows a simplified ground ohms system block diagram. A 2048 Hz sine wave is generated and drives a clamp-on inductor to induce a current in the ground conductor that is under evaluation. The induced current is proportional to the resistance of the loop. The induced 2048 Hz current is detected by a second receive clamp-on inductor, and the signal is fed to a 2048 Hz synchronous receiver. The output of the receiver is fed to the microcontroller that digitizes the received signal and sends it to the personal computer (PC). The PC, via a Windows application, analyzes the signals and displays the result on the PC screen and on the front panel of the test set (Figure 9).

The nature of this instrument is that it measures the resistance of a circuit. Therefore in grounding and bonding measurements there are at least two grounding connections being measured. This means that if you don't know something about one of the resistances, all you know from the measurement is the *sum* of the circuit resistances. If one side of the circuit is the electrical power ground, you generally have a low ground resistance to compare the unknown side to and that is helpful.

Figure 13 shows a simplified block diagram of the transmitter portion of the system. The basic frequency reference for the system is a 32,768 Hz crystal oscillator. This signal frequency is divided to yield the 2048 Hz transmit signal, and logically manipulated to give the I and Q clocks at 2048 Hz to drive the receiver synchronous detector. The 2048 Hz transmit signal square wave is converted to a sine wave with a resonant convertor to drive the resonant transmit clamp-on inductor. This induces the 2048 Hz transmit signal into the ground wire.

Figure 14 shows the simplified block diagram of the receiver portion of the system. The signal induced on the ground wire is detected by receive clamp-on inductor, that is operated in a non-resonant mode. The

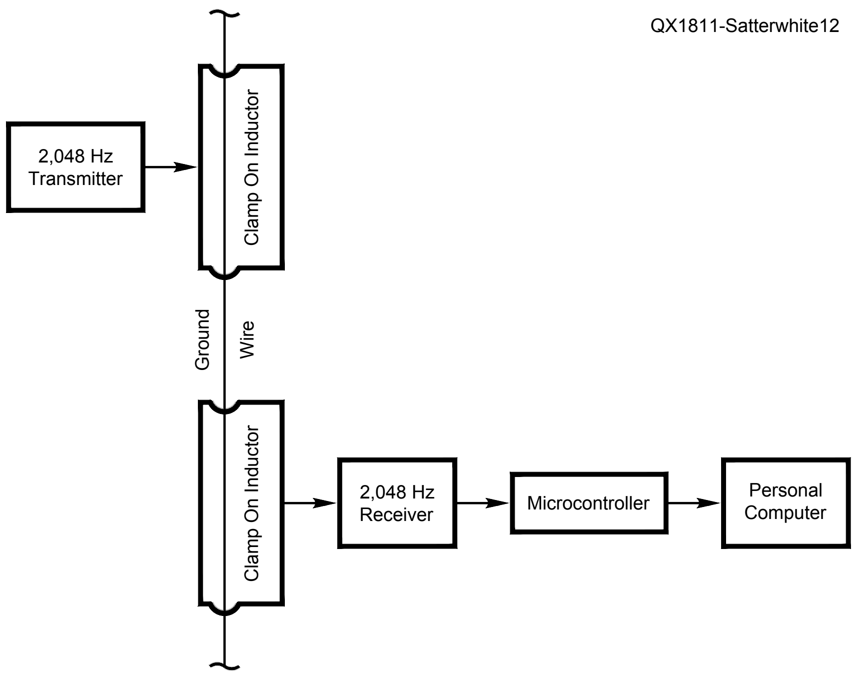

Figure 12 — Simplified ground resistance system block diagram.

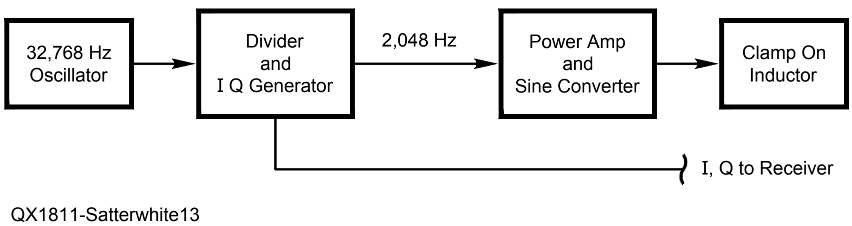

Figure 13 — Simplified ground resistance transmitter block diagram.

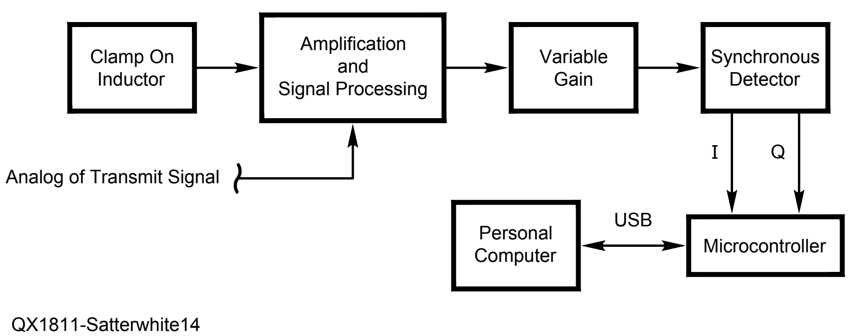

Figure 14 — Simplified ground resistance receiver block diagram.

Figure 15 — Measuring the a ground at the base of a tower.

detected signal is fed to amplification and signal processing circuits to reject unwanted interference signals and select the signals to be processed.

The 2048 Hz signal frequency is chosen to minimize the effects of power line induced noise in the ground current. From studies for the effects of power line induction interference on outside plant telephone circuits, it has been observed that the harmonics of the power line frequency are mostly odd and can be quite large. Assumptions have been made that the even harmonics are sufficiently lower than the odd harmonics to the point they can be ignored.

Figure 15 shows the ground resistance test set in action, measuring the competency of the grounding at the base of a tower.

Conclusions and Recommendations

What I have learned most from this exercise is that developing a culture of Amateur Radio operators measuring their installed ground resistance is important. While my design might never be in production, there may be some hams who will build this instrument or something similar. In any case, I recommend that clubs, organizations, those hams with antenna farms obtain or purchase a competent tester for use by their members. In these cases the cost is small compared to the investment in towers, antennas and electronic equipment.

ARRL member Jim Satterwhite, K4HJU, was first licensed in 1956 as KN4HJU, and later that year as K4HJU. He now holds an Amateur Extra class license and a General Radiotelephone Operator License. Jim is a registered Professional Engineer (Ret) in North Carolina. He is an iNarte Certified EMC Engineer (Ret) and a Life Senior Member of IEEE. He received a BEE degree from the University of Florida in 1965 and an MSEE degree from Purdue University in 1966. Jim has been involved in electronics research and development for more than 60 years, including 12 years as a member of the technical staff at Bell Labs and 34 years as a research and development engineer with Teltest Electronics, the company he founded in 1982. He holds a number of patents and patent applications. While in high school he built an AM transmitter from the ARRL Handbook. Jim enjoys developing electronic systems and is much more comfortable with MathCad or a soldering iron than a microphone. "There is no greater teacher than the lab." His Amateur Radio interests have included HF radio, slow scan TV, developing an antenna tuner using a unique power meter, and antenna analyzer design.

Notes

[1]Understanding Ground Resistance Testing AEMC Instruments publication, **www.aemc.com/techinfo/techworkbooks/Ground_Resistance_Testers/950-WKBK-GROUNDWEB.pdf**.

[2]H. Ward Silver, NØAX, *Grounding and Bonding for the Radio Amateur*,ARRL, 2017.

[3]Ken Rand, "Lightning Protection & Grounding Solutions for Communication Sites", PolyPhasor, Jan. 2000.

Dave Leeson, W6NL

15300 Soda Springs Rd., Los Gatos, CA 95030; **leeson@earthlink.net**

The Story of the Broadband Dipole

A dipole can be broadbanded by a number of techniques including by matching with resonant sections of transmission feed lines.

The 80-meter amateur band covers a very large fractional bandwidth. I was looking for a way to avoid retuning my 80-meter inverted V on a 140-foot guyed tower, since the ends are out of reach even if they are folded straight down. I found that lumped-element matching networks or the often-referenced Bazooka dipole and other forms of thicker conductors were neither practical nor effective for me at the lower HF bands.[1]

In a November 1997 posting on the TowerTalk mail list, I remarked[2],

> "There's an interesting method of broad banding antennas mentioned in various ARRL publications by Frank Witt, AI1H. It uses the transmission line length you are going to need anyway to match at two frequencies in the band (say 3.52 and 3.80 MHz). It's also mentioned in a text I use teaching a graduate microwave course at Stanford, and I have been wanting to try it."

In particular, I was referring to a September 1993 article[3] by Frank Witt, AI1H, that graced the pages of *QST*. Witt uses two cascaded resonant lengths of transmission line, one a multiple of half wavelengths and the other a quarter wavelength, to create a broadband impedance match for a simple wire dipole, as reproduced in Figure 1. Witt says:

> "The antenna system described here is simpler than any of its predecessors and has the following features: A 2:1 SWR or better is achieved over all or most of the 80-meter band. Antenna length and appearance are the same as those of a conventional half-wave dipole … The losses due to broadband matching are acceptable. [and] The dipole antenna itself is not broadband; the system uses a broadband *match*."

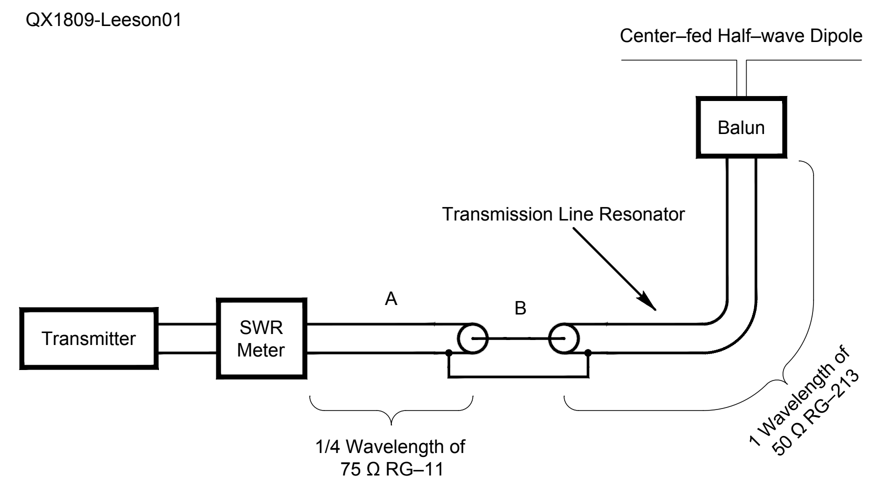

Figure 1 — Broadband feed line match from Witt, AI1H.

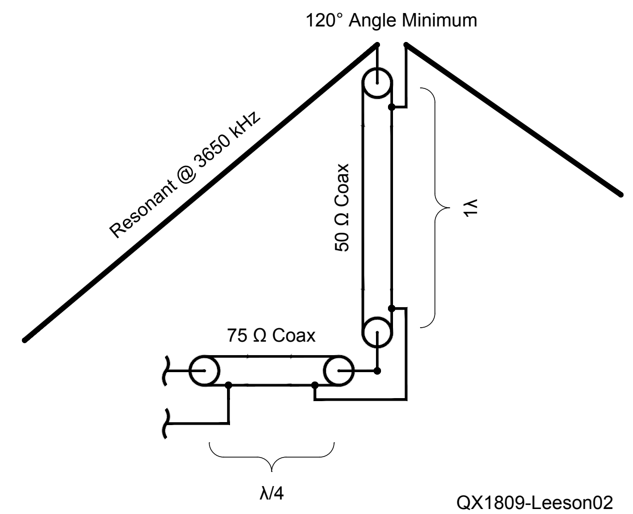

Figure 2 — Inverted V with broadband feed line match.

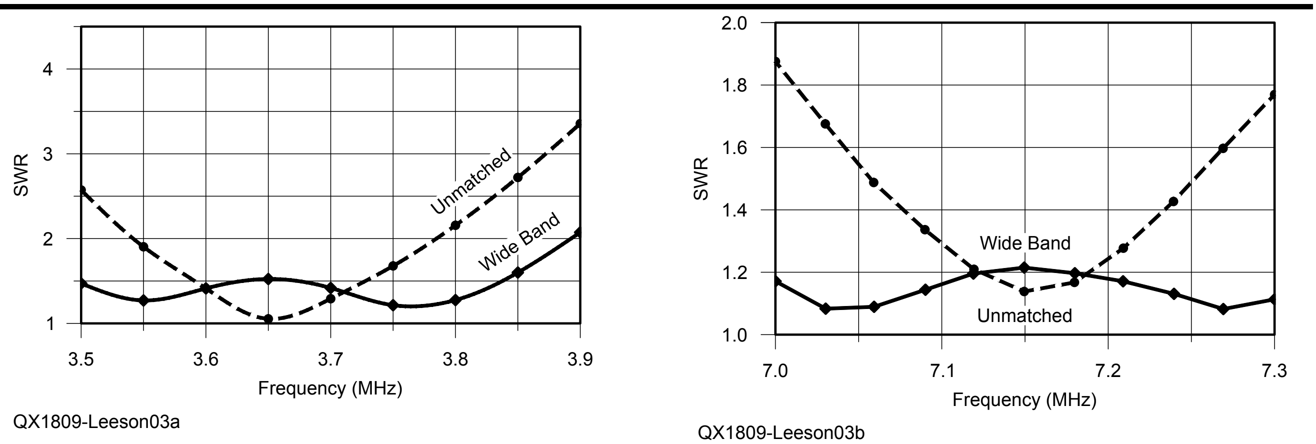

Figure 3 — Calculated SWR in the (A) 80-m band and (B) 40-m band.

Witt's idea seemed promising, especially after analysis using a spreadsheet and Smith chart software. This simple configuration would not arc or melt at high power, nor tangle or break with wind and ice loading. I found that this general concept was known[4] as early as the 1940s. If an inverted V is used, the angle must be shallow to keep the radiation resistance higher than the feed line characteristic impedance. Figure 2 is a sketch of the inverted V and feed line matching network from a talk I gave[5] in October 1998. From my post on TowerTalk:

> "Here is a summary of an 80-m inverted V I put up that matches 1:1 at the CW and SSB frequency. The inverted V is designed using AO to resonate at 3.67 MHz (geometric mean of two frequencies of interest), and to have enough height and included angle to have a resistance at resonance of 70-75 ohms. The apex is at 120 ft, and the angle is about 120 degrees. I use the AO optimizer, then knock off 1% length from experience.
>
> Now for the interesting part. I measured a full wavelength of 50-ohm coax using the MFJ-259 antenna scope. Using Belden 8214, it's about 205 feet long, connects to the antenna 1:1 balun and comes down to the ground. Next is connected a 1/4 wavelength of 75-ohm coax.
>
> "At the transmitter end of the 75-ohm quarter wave, the match is perfect (less than 1.1:1) at 3520 and 3800 [kHz]! This should work for a horizontal dipole, or any antenna with impedance in the 75-ohm range.

To see if this was a fluke, I put up a 40 m dipole at 50 feet for Sweepstakes. It uses a half-wave of 50-ohm coax followed by the quarter-wave section of 75-ohm coax, and it matched fine at 7.0 and 7.25 MHz."

The calculated SWR response for the 80 m antenna is shown in Figure 3A, and the SWR response for the 40 m antenna is shown in Figure 3B, from my 1998 presentation.

L. B. Cebik, W4RNL, posted a confirming comment the following day, in which he provided 80-m calculated results for various configurations, based on dipoles rather than the inverted V.[6] Cebik later posted a more thorough analysis on his web page.[7] His principal result, the graph of an analysis of the effect of differing half-wave line-length multiples, is also found in a talk[8] by Jim George, N3BB, and is reproduced in Figure 4.

This antenna has proven to be a useful addition to the toolkit of station design and construction. We made use of several such dipoles in our successful contest station[9] in the Galápagos Islands, HC8N.

In experimenting with a 40-m dipole and this feed system, I found that one could compensate for antenna mistuning by adjusting the length of the 50 Ω transmission line. This is in essence creating a series-section transformer to match an arbitrary impedance.[10] An extended analysis of the transmission-line matching scheme is found in a paper[11] by E. J. Shortridge, W4JOQ. However, it is generally easier to adjust the length of the dipole legs themselves, rather than change the transmission line lengths and/or impedances. Once the required multiple

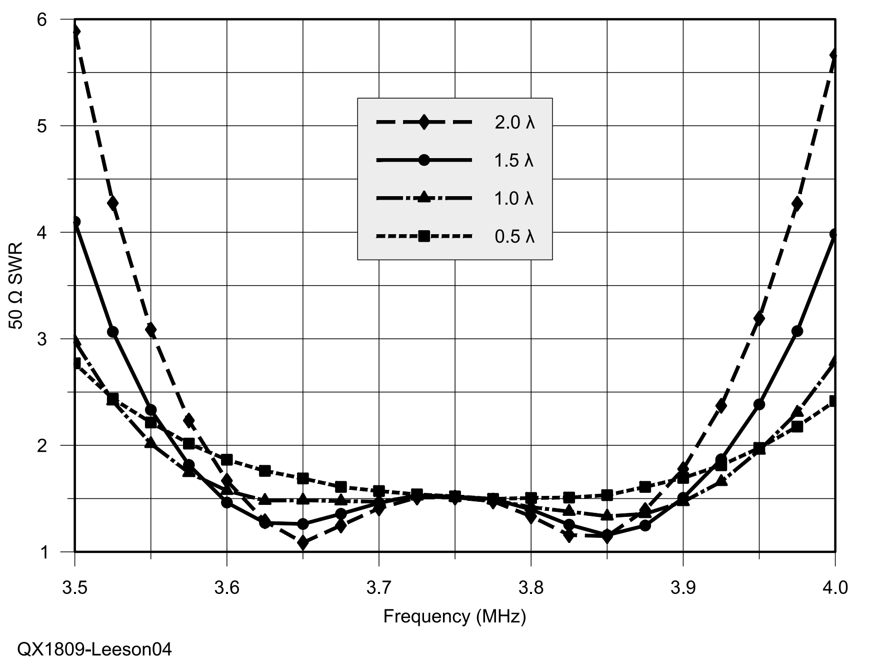

Figure 4 — Wide-band match calculated by Cebik, W4RNL.

half-wave 50 Ω transmission line is in place, an antenna impedance bridge can be used at the feed end on the ground to cut the antenna to resonance at the chosen center frequency.

History of Broadband Matching Alternatives

Broadband impedance matching[12] has drawn particular interest since the development of microwave radar in the Second World War. The classic Fano paper[13] that defined limits appeared as a thesis in 1948, and was published in 1950. The specific techniques of broadband antenna matching have been explored over the years.[14] The continuing interest in dipole matching bandwidth is evidenced by a recent paper[15] on the subject.

In general, a dipole antenna can be rendered more broadband by increasing its diameter to reduce the reactance variation with frequency — but at the lengths of HF antennas this is not a great improvement even if a wire cage is employed — or by compensating the feed point reactance variation with some form of matching network, preferably one that does not introduce excessive loss. A useful simple model for the driving impedance of a half-wave dipole antenna near resonance is a quarter-wave transmission line loaded by a resistance corresponding to the radiation resistance.[16] The reactance is not zero at an exact half-wave length, and is typically zero at a length slightly less than one half wave. The exact reactance at a half-wave and the length for zero reactance both depend on the diameter and shape of the dipole. There are more precise lumped-element models in recent literature.[17] Two comprehensive summaries of broadband antenna models and impedance matching are by S. Stearns, K6OIK.[18]

The search for a broadband 80-m dipole antenna has a long history, culminating in a series of articles[19] by Frank Witt, AI1H. Beginning with the antennas seen in early radio, a wide range of broadbanding concepts has been explored. Here are some major categories:

- Cage, parallel wire, fan or bow-tie dipoles that have a large equivalent diameter or that approximate conical dipoles[20]
- "Bazooka" dipoles with coaxial radiating elements[21]
- Multiple dipoles with staggered resonances[22]
- Dipoles with a coupled resonator "open-sleeve" element[23]
- Dipoles with lumped reactive matching networks[24]
- Dipoles with coaxial radiating and matching elements[25]
- Dipoles with resonant feed line matching, as outlined in this article.

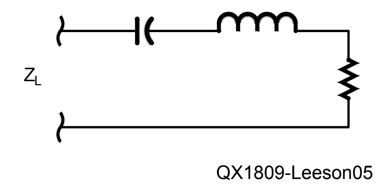

Figure 5 — Dipole equivalent circuit.

Collinear half-wave dipoles have higher feed-point impedance and could provide wider bandwidth. A simpler version is the Extended Double Zepp, but in either case the pattern may be narrower than desired and the extra physical length may be undesirable.[26]

The "open sleeve" or coupled dipole approach is similar in concept to the technique of using a Yagi director close-coupled to the driven element in the "OWA" Yagi designs that produce a wide SWR curve.[27]

There are mechanical and reliability problems associated with many of the broadbanding schemes. Cage and multiple dipole configurations have a tendency to become twisted in the wind, are difficult to construct and install, take up more space, and have greater visual impact than a simple dipole. Coaxial radiating elements require sealing from the weather, are heavy to support and are not necessarily strong enough to avoid stretching under load. In addition, some of the published designs don't have sufficient broadband response, while others are broadband mainly because of losses in the matching network.

The advantage of resonant feed-line matching lies in its simplicity and the fact that there are no additional losses at the

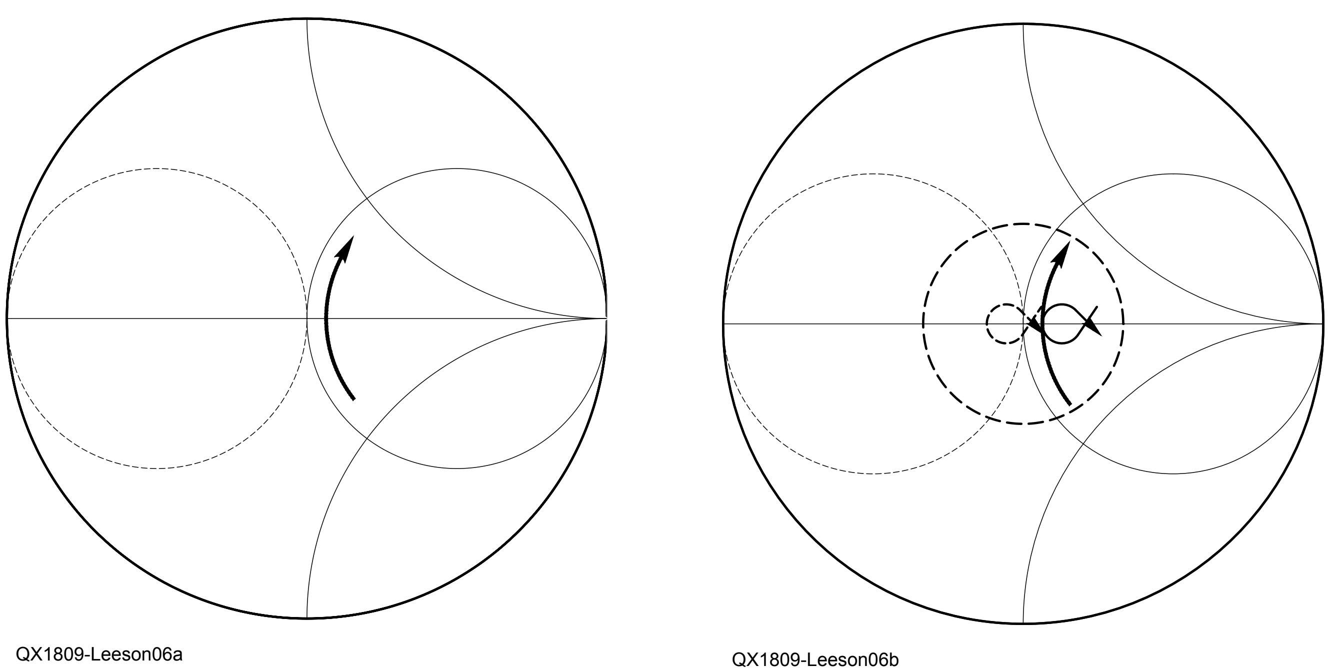

Figure 6 — Smith chart shows impedance vs. frequency, (A) on left at antenna terminals, and (B) after multiple half waves.

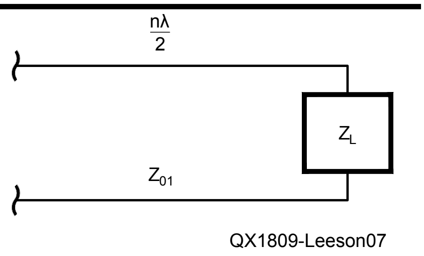

Figure 7 — Multiple half-wave transmission line.

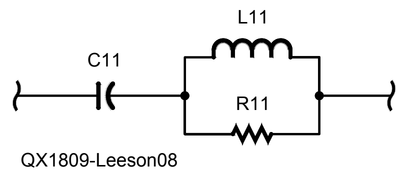

Figure 8 — Dipole equivalent circuit, from Tang, et al.

antenna itself. For the specific application in the 80-m amateur band, this approach has proven to be effective and reliable.

Visualizing How This Works

Circuit theory and a Smith chart representation can help to understand how resonant matching of resonant elements works.[28] This technique is particularly applicable in the case of a load impedance that has the characteristics of a series *RLC* network. This is typical of antennas, and is shown schematically in Figure 5.

A Smith chart is a polar plot of complex reflection coefficient. The impedance plot of this form of load, in this case for $R > Z_0$, is shown in the Smith chart plot of Figure 6A, with the arrow in the direction of increasing frequency. Placing a parallel resonant circuit across the terminals of the dipole, or feeding it through a single or multiple half-wave line of suitable Z_0, results in a trace that curls back on itself as frequency varies.

The operation of the matching network is to close up the arc of the impedance plot as shown in the Smith chart of Figure 6B, resulting in a smaller range of SWR. In particular, the feed line of Figure 7 is electrically shorter at the low-frequency end, and longer at the high-frequency end, thus achieving the desired closing of the impedance curve.

A final step is to use a matching network such as an *LC* network, its transmission line equivalent or the series section network to bring the smaller impedance plot to the center of the Smith chart. As sought in filter design and current ultra-wideband antenna work, a trace that curls around the center of the Smith chart represents a broadband low SWR.[29]

This type of resonant matching is not widely known, but is highly effective in the case of impedances of the form of a loaded series resonant circuit. More than one half-wavelength can be used if the line impedance is not optimum, and a short addition or subtraction can be made from the line length to "center" an impedance plot that is not symmetrical. For the specific impedance plot shown here, we require $Z_{01} < R_L$. More detailed descriptions of this matching concept have been available in the technical literature for quite some time.

Dipole Models of Improved Accuracy

Tang, et al.[30], present some more accurate equivalent circuits for a dipole near its half-wave resonance. These may possibly be worth exploring, but the simple three-element equivalent circuit, as shown in Figure 8, seems adequate for this application.

In this model, the series capacitance is,

$$C_{11} = \frac{27.82\times10^{-12}\,h}{\{\ln(2h/a)-1.693\}} \tag{1}$$

where *h* is the half length of the dipole and *a* is the radius. The inductance is,

$$L_{11} = \frac{1}{\omega_0^2 C_{11}} + C_{11}R_{ao}^2 \tag{2}$$

and the resistance is,

$$R_{11} = \frac{L_{11}}{C_{11}R_{ao}} \tag{3}$$

where ω_o is the resonant frequency and R_{ao} is the radiation resistance at resonance.

For a 3.65 MHz dipole using #10 AWG wire, the values are, C_{11} = 61.1 pF, L_{11} = 31.4 µH and R_{11} = 7.13 kΩ.

This model can be used with a Smith chart to experiment with various wide-band matching schemes. In general, the amateur bands are sufficiently narrow that the improvement available from a better model is not likely to have much effect, and in any event the Witt broadband transmission-line matching approach provides a very simple and useful antenna.

Dave Leeson, W6NL, was first licensed in 1952. He enjoys the friendship, technology and contesting aspects of ham radio, having been part of winning radiosport efforts from North Africa, the Caribbean, the Galápagos and California. He is the author of the ARRL book Physical Design of Yagi Antennas, and was a designer of the IARU International Beacon Network. He drove a racecar, retiring in 1979 to return to ham radio after back-to-back national championships. Professionally, he received degrees from Caltech, MIT and Stanford, is the author of widely cited journal papers on oscillator phase noise and nonlinear circuits, and is a Life Fellow of the IEEE. From 1968 to 1993, he was the founding Chairman and CEO of California Microwave, Inc., retiring with a staff of two thousand. Since 1994, he has been a Consulting Professor at Stanford, and is the faculty advisor of the Stanford Amateur Radio Club, W6YX.

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David Birnbaum, K2LYV

8515 Acorn Ridge Ct., Tampa, FL 33625; **dbirnbau@gmail.com**

Direct Calculation of Antenna Tuner Losses

Built-in and external antenna tuners are convenient, but can add to losses.

Modern ham radio transceivers commonly include some sort of automatic antenna tuner that automatically adjusts its component values to provide a close to 1:1 SWR match. The most common form of these internal tuners is an L-network, most commonly with series inductors and shunt capacitors that can be switched to either the transceiver or antenna side of the inductor.

It is well known that an L-network composed of two reactive elements can match any two impedances at a given frequency. For a given load impedance and frequency one can calculate the exact values of the components of a network that will make an exact match. It is also straightforward to calculate the corresponding currents and voltages in the network. However one must also consider the possible losses in the tuner components when the impedance of the antenna is far from the nominal 50 Ω. In many cases while the tuner can make a near-exact match, there are either large losses in the tuner and/or large voltages on some of the components. These stresses place limits on the range of impedance values that a given set of components can handle.

It is a relatively straightforward exercise to calculate the voltages and currents in the components of a tuner for a given load impedance. Such a calculation can show the potential for destructive stresses in the tuner.

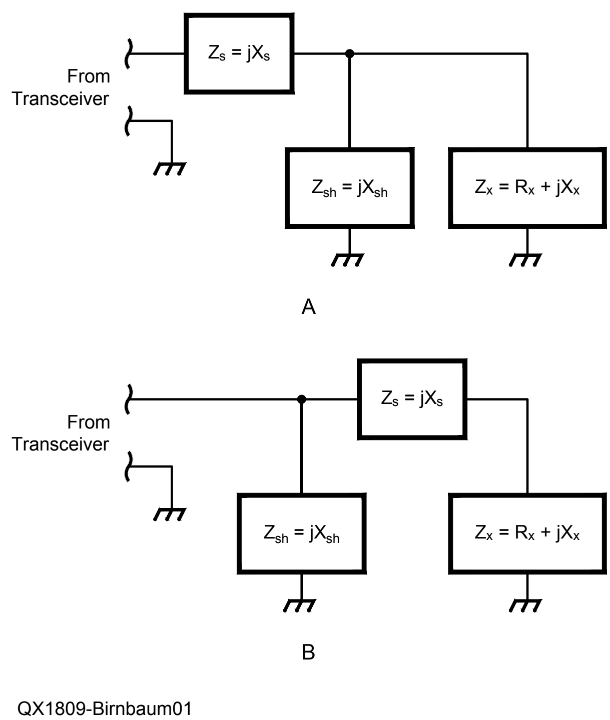


Figure 1 — The L-network can match loads with the real part (A) larger than 50 Ω, or (B) smaller than 50 Ω.

Figures 1(A) and (B) show two different configurations of an L-network. They differ in the placement of the shunt element. In Figure 1 Z_x represents the complex load impedance, Z_{sh} the shunt impedance and Z_s the series impedance. In general, each element can be either a capacitor or an inductor.

The key insight into the design of the L-networks of Figure 1 is that only the shunt element changes the value of the resistive part of the impedance presented to the transceiver.

For the network in Figure 1(A) the parallel combination of the load impedance and the shunt element will have a resistive and reactive component. The addition of the series element will modify only the reactive part of the parallel combination. Hence the parallel combination must have a resistive part of 50 Ω.

Similarly, for the network in Figure 1(B) the series element adds or subtracts from the reactive component of the load. Therefore only the shunt element can modify the series combination so that the resistive part of the resulting impedance is 50 Ω. At the same time the shunt and series element values must make the reactive component presented to the source equal to zero.

For the L-network configuration it is straightforward to use the rules for series and parallel combinations of impedances to compute the impedance presented to the source by the combination of the L-network and the load. We must consider the complex nature of the elements. For those not familiar with math using complex values, see Section 2.10 of *The ARRL Handbook*.[1]

The input impedance of the network in Figure 1(A) is,

$$Z_{in} = Z_s + \frac{Z_x Z_{sh}}{Z_{sh} + Z_x} \quad (1)$$

where the impedances are complex values. For the moment we will assume that all inductors and capacitors are ideal, that is, they have only reactance. Later we will correct the results to take into account the resistive components. Since the typical components used in building antenna tuners have low losses this correction will be small.

If we write the impedances as a combination of real and complex parts we have,

$$Z_{in} = jX_s + \frac{jX_{sh}\left(R_x + jX_x\right)}{R_x + j\left(X_x + X_{sh}\right)} \,. \quad (2)$$

We now expand the expression in Eqn. (2) and separate the real and imaginary parts,

$$\mathrm{Re}\left\{Z_{in}\right\} = \frac{R_x X_x^2}{R_x^2 + \left(X_x + X_{sh}\right)^2} \quad (3)$$

and

$$\mathrm{Im}\left\{Z_{in}\right\} = \frac{R_x^2 X_x}{R_x^2 + \left(X_x + X_{sh}\right)^2} + X_s \,. \quad (4)$$

Since the real part of Z_{in} depends only on the value of the shunt impedance (R_x and X_x are known), we can solve Eqn. (3) for the value of shunt impedance that makes the real part equal to 50 Ω. Then we use this value of the shunt impedance to solve Eqn. (4) for the value of the series impedance that makes the imaginary part of Z_{in} zero.

The algebra is easier if we solve for the susceptance of the shunt element, $B_{sh} = 1/X_{sh}$. While you can solve the equations by hand, there are computer algebra programs that can solve equations symbolically. *Maxima* (**maxima.sourceforge.net**), *Mathcad*, *Octave* and *Python* have symbolic algebra capabilities. Solving first for the value of the shunt impedance to make the equivalent resistance 50 Ω gives,

$$B_{sh} = \frac{X_x \pm \sqrt{\frac{R_x}{50}\left(R_x^2 + X_x^2 - 50R_x\right)}}{R_x^2 + X_x^2} \,. \quad (5)$$

The corresponding value of the series element that reduces the imaginary part of the input impedance to zero is,

$$= \frac{1}{B_{sh}} + \frac{50}{R_x} - \frac{50}{B_{sh} R_x} \,. \quad (6)$$

Solving for the equivalent impedance of the circuit of Figure 1(B) is simpler if we work with the admittance rather than with the impedance,

$$\frac{1}{Z_{in}} = Y_{in} = jB_{sh} + \frac{1}{Z_s + Z_x} \,. \quad (7)$$

We want the real part of Z_{in} to be 50 Ω and the imaginary part to be zero. This is equivalent to requiring that the real part of B^{in} is 1/50 S and that the imaginary part of Y_{in} is zero.

$$\mathrm{Re}\left\{Y_{in}\right\} = \frac{-R_x}{R_x^2 + \left(X_x + X_s\right)^2} \quad (8)$$

and

$$\mathrm{Im}\left\{Y_{in}\right\} = \frac{1}{X_{sh}} + \frac{X_x + X_s}{R_x^2 + \left(X_x + X_s\right)^2} \,. \quad (9)$$

We first solve for the value of the series element that makes the real part of Y_{in} = 1/50 S,

$$X_s = -X_x \pm \sqrt{R_x\left(50 - R_x\right)} \,. \quad (10)$$

Now substitute this into the expression of the imaginary part of B_{in} and solve for the value of X_{sh} that makes the imaginary part zero,

$$X_{sh} = \frac{\sqrt{R_x\left(50 - R_x\right)}}{50 R_x} \quad (11)$$

Notice that since the term (50 - R_x) appears under the square root, the configuration of Figure 1(B) can be used only when the real part of the load is less than 50 Ω.

The solutions in Eqns. (3), (4), (10) and (11) have assumed ideal reactive components. If we are to calculate the losses for the real components we need to take the finite Q of the components into account. For the most general case this makes the equivalent solutions for equations extremely unwieldy. However for most real components the Q values are quite high and hence the resistive parts of the impedance are small compared to the reactive parts. We can therefore calculate the currents (and voltages) using the lossless component values and then use the Q values to calculate the resistive part of the elements and use that to calculate power losses in each component.

In the examples that follow I assume that the input power is 100 W. All losses scale as the square root of the actual power used, that is, for 1000 W the power losses will be $\sqrt{10}$ larger than those given here.

For 100 W into a 50 Ω load the source must provide $I_{in} = \sqrt{2}$ A, or 1.414 A. In the Figure 1(A) configuration, all of this current flows through the series element. The current then divides between the shunt element and the load, where the current in the load is,

$$I_x = I_{in} \frac{Z_{sh}}{Z_{sh} + Z_x} \quad (12)$$

and the current in the shunt element is given by:

$$I_{sh} = I_{in} \frac{Z_x}{Z_{sh} + Z_x} \,. \quad (13)$$

For the network configuration of Figure 1(B), the current divides between the shunt element and the combination of the series element and the load. The two currents are,

$$I_s = I_x = I_{in} \frac{Z_{sh}}{Z_{sh} + Z_x + Z_s} \quad (14)$$

and

$$I_{sh} = I_{in} \frac{Z_x + Z_s}{Z_{sh} + Z_x + Z_s} \,. \quad (15)$$

The loss is I^2R where R is X/Q where Q is typically 1000 for a capacitor and about 200 for an inductor. We use the rms value (absolute value if I is complex) of the current. For most real components, since the Q of the capacitors is much higher than that of the inductors, most of the loss will be in the inductor. However the voltages across the capacitor may become large enough to cause breakdown and arcing.

As an example, consider matching a short vertical antenna 21.5 ft tall with a 2 ft top hat at 7.1 MHz. An NEC simulation gives an impedance of 27.9 - *j*430 Ω. Using the configuration of Figure 1(A) where the shunt element is a capacitor, we substitute the values of the antenna impedance into Eqns. (3) and (4). The solution is X_{sh} = -56 Ω corresponding to a capacitance of 399 pF and X_s = 455 Ω corresponding to an inductance of 10.2 μH. For a coil Q of 200 the loss in the coil for 100 W input is 7.5 W, and the loss in the capacitor is 0.1 W.

Now consider the same antenna at

3.5 MHz where the simulation predicts an impedance of 21.4 - j1420 Ω. Again using Eqns. (3) and (4) we get a match with a shunt capacitor of 1051 pF and a coil with 66 µH inductance. Now the loss in the inductor is 25 W for 100 W input and the loss in the capacitor is 0.1 W. The 25% loss of power is important, but the high thermal stress on the inductor must be considered. Unless the inductor is constructed of heavy duty components, it is likely to fail. Using modes with lower duty cycles (CW or SSB) will reduce the thermal stress, but high duty-cycle modes, like most digital modes, would require reduction in power to keep the inductor from failing.

Several modern computer languages, like Python and Octave can handle complex numbers as part of their standard syntax. The widespread (and free) availability of Python and Octave make it easy to implement a computer program that calculates the component values for an L-network and the expected losses (and voltages) for a given antenna load. An example of such a Python program to compute the component values and power losses available on the **www.arrl.org/qex-files** web page.

While a similar analysis can be applied to the commonly used T-network tuner the solution is not as simple as the one for the L-network. There are three components that can be varied and hence there are multiple combinations that will provide a match. In order to find an optimal solution, a third parameter must be introduced. An example of such a parameter set could be the combination of components that provides a match and simultaneously has the lowest power loss in the tuner components. However to solve such a problem requires solving a set on non-linear complex algebraic equations. Such a problem is best solved by using an optimization technique to minimize the power while making a match, rather than through direct analytical solution

One of the arguments made to prefer T-network tuners over L-network tuners is that sometimes large values of the components are needed to make a match with the L-network. This argument is relevant when using variable capacitors and inductors, especially if they must also have high voltage ratings. Using modern technology, a larger range of component values can be created with a set of digitally controlled switches to combine individual capacitors and inductors. Using a binary coding method a large range of values can be covered using a modest number of individual components. For example a set of 8 inductors can produce any inductance between 0.25 and 63.75 µH in steps of 0.25 µH. Similarly, a wide range of capacitance values can be generated. Indeed, modern automatic antenna tuners, including those built into many transceivers, are designed using such a binary coding method. However given the space limitations in modern transceivers, the components are often smaller than ideal and hence are limited in their ability to handle the stress of high voltage on the capacitors or thermal stress on the inductors. This is most likely the reason that the built in tuners in most transceivers are limited to handle an SWR of 3:1 or less.

Dave Birnbaum, K2LYV, holds an Amateur Extra class license. His previous calls were KN2LYV, W3GFC, WØZX and W2IIV. Dave has a BEngr in Engineering Physics from Cornell University and MS and PhD in physics as well as an MBA from the University of Rochester. After teaching physics at the university level for several years, he transitioned to the General Railway Signal and then to Xerox Corp where he worked on digital imaging for color xerography and ink jet printing. Dave retired from Xerox as an Engineering Fellow in 2003. He holds 15 patents. In addition to ham radio, he plays the trombone in a few local concert bands. David is a member of the Tampa Amateur Radio Club and especially enjoys digital modes and CW. He has earned the WAS and DXCC awards. Dave is a Technical Specialist in the West Central Florida Section.

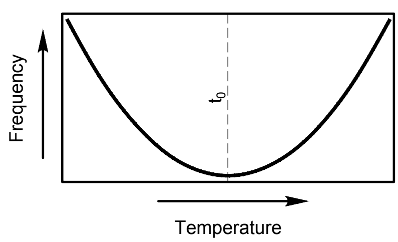
Notes

[1] *The ARRL Antenna Book, 2017 Edition,* Available from your ARRL dealer or the ARRL Bookstore, ARRL item no. 0628. Telephone 860-594-0355, or toll-free in the US 888-277-5289; **www.arrl.org/shop**; **pubsales@arrl.org**.

Technical Notes

Theory and Practice of High Stability Crystal Oscillators

There are applications in the modern world where stable frequency sources are required. This implies a high-Q frequency determining element. An obvious candidate is the quartz crystal. While it is satisfactory at stable temperatures it exhibits a frequency shift when the temperature is varied, as it is in military applications, for instance. This is shown in Figure 1 for a typical crystal. The curve is a second order polynomial. This frequency shift can be partially compensated for by external temperature compensating circuitry (TCXO), but there are other problems. While Figure 1 shows the ideal frequency *vs.* temperature dependence, in practice there are possible anomalies and spurious responses. This is shown in Figure 2, which is an actual measurement of an 89.9 MHz crystal. The figure shows that some crystals exhibit a spurious response at some temperatures.



QX1811-Przedpelski01

Figure 1 — The curve for a typical crystal temperature response is a second order polynomial.

The best solution is to use a crystal in a temperature controlled oven at a temperature somewhat above the highest expected ambient. A crystal should be then selected with a turn-over temperature, t_0, equal to the oven temperature. At that temperature $df/dt = 0$ and is small for small departures from the set value. In addition, the crystal should not have any undesired characteristics over a much smaller temperature range.— *Best regards, Andrzej (Andy) Przedpelski, KØABP.*

Handheld Transceiver Earpiece-Microphone Modification

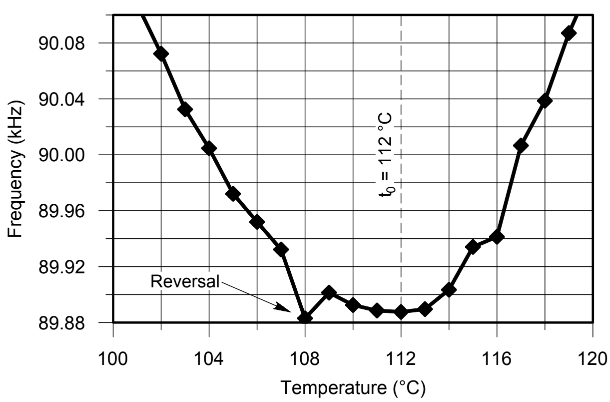
Handheld transceiver earpiece accessories don't always fit every ear. This modification to a proprietary headset allows for earpiece selection, improving overall usability. I recently purchased a Yaesu SSM-512B "VOX Earpiece Microphone" for my Yaesu FT-65R/25R dual band transceiver. It includes an over-the-ear earpiece and a small clip-on housing or dongle that contains the microphone, PTT/VOX and PTT switches, all connected by cords (Figure 3). The transceiver plug is a pair of two different-sized phone plugs — each tip-ring-sleeve (TRS) or 3-conductor stereo-style — molded together. The 3.5 mm connector is for SP (speaker) and the 2.5 mm connector is for MIC (microphone). The separation between the two plugs appeared to be about 8 mm. The MIC/SP connectors on the Yaesu FT-65R/25R are standard miniature phone connectors, but the SSM-512B right angle, paired, molded connector that mates with the transceiver appears unique and is one of the main advantages of the SSM-512B headset.

The earpiece fit my ear poorly. It uses a small disk-shaped earphone/speaker secured with an ear hook made of soft rubber. The thin cable from the speaker routes through the ear hook, and appears behind the ear when worn. The ear hook was far larger than my ear and in no way captured it. The whole contraption rested precariously on top of my ear, and the small speaker continually fell out of my ear under normal use. No amount of adjustment seemed to help.

I modified the headset by cutting the earpiece cable at its half-way point and installing a female 3.5 mm stereo cable connector (available from various source; **www.allelectronics.com**, **www.mouser.com**, **https://www.performanceaudio.com/**) so that I could plug in any earpiece that had a compatible plug. Installation of this female phone connector took some extra care. The earpiece cable was very small, about 2 mm diameter, with a soft rubber outer cover enclosing two, thin Litz wire conductors. Carefully using a small hobby knife, I was able to remove the outer soft rubber and expose short lengths of the inner wires. Each wire readily accepted tinning using a medium-powered soldering iron, which burned off the wire coatings and fiber strands.

The internal conductors were color-coded via the insulated coating on the wires. I used a multimeter to ring out the connections to the SP plug. I found that red was the "Sleeve", and the copper (or clear) wire was the "Tip".

I soldered the color-coded wires to the appropriate "Tip" and "Sleeve" contacts, but



QX1811-Przedpelski02

Figure 2 — An actual measurement of an 89.9 MHz crystal shows a spurious response at some temperatures.

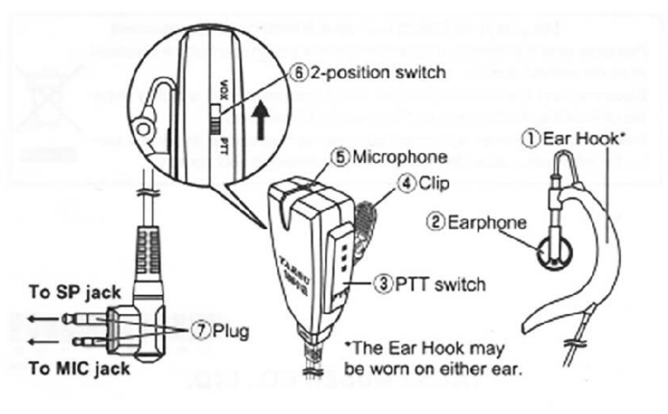


Figure 3 — Ear hook and PTT as depicted in the SSM-512B User Guide.

Figure 4 — The handheld transceiver (upper left) and modified SSM-512B (upper right) with several tested head sets and ear pieces.

left the "Ring" connector contact empty. The connector had a clamping part that held onto the cable, but for extra measure I added a small dollop of silicone rubber sealant (RTV) into the end of the finished connector where the cable enters the housing. This also solved the problem that the cable was much smaller than the connector housing opening.

For a test, I connected the modified headset cable to my Yaesu FT-65R transceiver and plugged in several different headset-earpieces (Figure 4). These included ordinary stereo-head headphones, an ordinary stereo ear bud set, a Scan Sound LBUD3 single earpiece (**https://www.scansound.com/earphones/single-ear-stereo-earphones**), and a single-ear "surveillance" acoustical tube earpiece (**https://midlandusa.com/ https://www.rei.com/**).

I added a 3.5 mm stereo plug to the earpiece cable and an ear mold (**www.amazon.com**) to hold it in my ear. Of course, the stereo dual ear devices responded in just one ear. Both the LBUD3 and acoustical tube earpieces worked best and were roughly equivalent in performance, although the LBUD3 was noticeably louder. Both fit my ear perfectly and neither of the single earpieces dislodged from my ear under normal use yet were comfortable when worn for several hours.— *Best regards, Scott Roleson, KC7CJ.*

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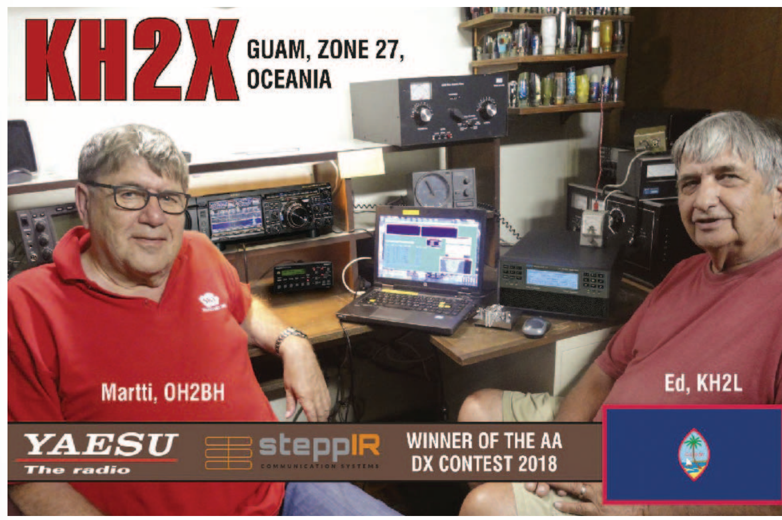
Congrats!

Congratulations to Martti Laine, OH2BH for winning the All Asian DX Contest 2018 – single operator, high power.

Martti operated in Guam as KH2X at the ham station of Ed Poppe KH2L, who fearlessly undertook the task of installing the 3 Element SteppIR Yagi with 40m/30m loop dipole in the midst of typhoon conditions!

Above: John Mertel WA7IR, CEO of SteppIR Communications, shown with SteppIR Advisory Board member and world-renowned DX'er Martti Laine OH2BH at Tokyo Ham Fair 2018



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