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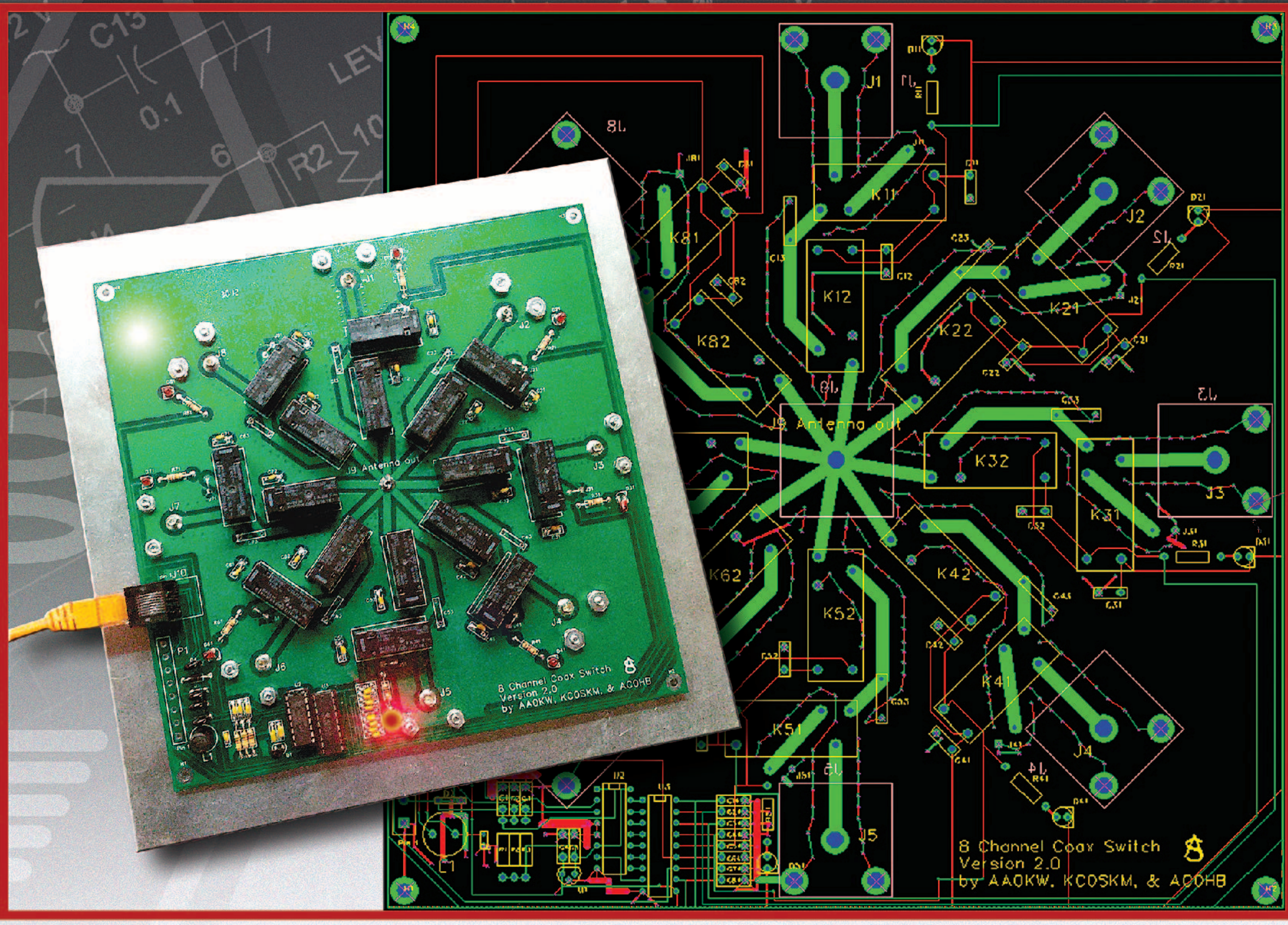
QEX

A Forum for Communications Experimenters

Issue No. 283



ACØHB describes "An Eight Channel Remote Control Antenna Selector." Connect a desired antenna to your radio with only one coaxial cable run to your shack, or select between several radios. This antenna switch achieves better than 70 dB of port-to-port isolation!

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225 Main Street
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QEX (ISSN: 0886-8093) is published bimonthly in January, March, May, July, September, and November by the American Radio Relay League, 225 Main Street, Newington, CT 06111-1494. Periodicals postage paid at Hartford, CT and at additional mailing offices.

POSTMASTER: Send address changes to: QEX, 225 Main St, Newington, CT 06111-1494
Issue No 283

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Subscription rate for 6 issues:

In the US: ARRL Member $24, nonmember $36;

US by First Class Mail:
ARRL member $37, nonmember $49;

International and Canada by Airmail: ARRL member $31, nonmember $43;

Members are asked to include their membership control number or a label from their QST when applying.

In order to ensure prompt delivery, we ask that you periodically check the address information on your mailing label. If you find any inaccuracies, please contact the Circulation Department immediately. Thank you for your assistance.

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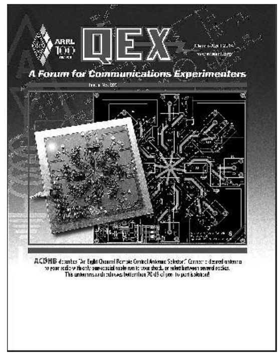

Michael Dzado, ACØHB writes about his "Eight Channel Remote Control Antenna Selector." With a single coaxial cable run to his shack, he is able to select between up to eight antennas. The Collins Amateur Radio Club uses a pair of the selectors to pick between several antennas and four radios. With better than 70 dB of port-to-port isolation he is sure the transmitted signal is going to the desired antenna!

In This Issue

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The American Radio Relay League

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interest 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 the 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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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**.

Larry Wolfgang, WR1B

Empirical Outlook

ARRL Centennial Celebrations

By now I am sure all of our readers are aware of the many ARRL Centennial events taking place throughout 2014. The excitement level around ARRL Headquarters is very high, and it is easy to sense the growing anticipation of many events coming up later this year. The National Convention in Hartford, Connecticut on July 18 through 20, 2014 is causing quite a stir, since many Headquarters Staff members are playing a major role in planning this event. Oh, we participate in National Conventions every year, and there are always several other major conventions that require a bit of extra planning and participation by Staff members. We don't normally have much to do with the overall convention planning, though. That is usually handled by the committee members from the area local to the convention who plan that gathering every year.

There hasn't been a major ARRL Convention in the Hartford area in recent memory. Luckily, we have lots of help from many of the people who organize the New England Division Convention in Boxboro, Massachusetts every other year, as well as a dedicated group who have been putting on the popular "Nutmeg Hamfest/Connecticut State Convention" for a number of years. Even so, the buzz around here from all the additional activity of planning the Centennial Convention is palpable.

Of course State and Section Conventions across the country will be a bit extra special this year. I am sure there is a lot of similar activity to go along with planning all those events. I hope you plan to participate in or at least attend your local hamfests and conventions, as well as travel to one or two of the larger events.

There are a number of special operating activities that you can participate in from the comforts of your own radio shack, as well. I previously mentioned the ARRL Centennial QSO Party as one of those activities. This is shaping up to be a fun on-the-air event, even for the most casual operators. Just spend a bit of time on the air, make a few QSOs, and submit your logs to Log Book of the World. As your logs are processed they are cross checked, and points are awarded based on the confirmed contacts with other ARRL members. You don't even have to keep track of the points values, although I think it is fun to find out what appointments and positions my contacts have in the ARRL Organization. In the Nov/Dec 2013 Empirical Outlook I mentioned that I had been told QSOs with your *QEX* Editor were worth 30 points. In a later refinement of the "Rules" all ARRL Headquarters Staff Members became worth 50 points per contact. So, it's even better than I first thought.

Have you been chasing the portable W1AW operations? I have managed to contact most of them so far, but I've already missed a couple so I am looking forward to the second operation from at least some of the states. It seems to me that the operators tasked with these "Field Operations" have been doing a great job, using the popular modes and having one or more operators on the air at all times. At least I have found it pretty easy to look them up on the DX Cluster and then with some patience to get them in my log. You can check the schedule for which states and territories will be on with the W1AW portable call sign on the ARRL website: **www.arrl.org/files/file/On the Air/W1AW_2014_sked.pdf**. There are two portable operations, each lasting one week, with each state being on two different weeks during the year. As an added bonus, The Hiram Percy Maxim Memorial Station at ARRL Headquarters is operating as W100AW. Find a guest operator or maybe even an ARRL Staff Member operating from one of the studios, and have fun chatting with this special event call sign.

My Christmas Wish List for last year included a Raspberry Pi. Jean, WB3IOS, saw fit to fulfill that wish, so I was thrilled to find a Pi under the tree on Christmas morning. I am still very much at the bottom of the learning curve with this little circuit board, but it is definitely fun to play with. Of course, I've had to add a few "accessories," so I can start to learn about inputs and outputs, to explore some things I can do with the computer. The "Operating Manual" I found on line was two pages of quick-start instructions, so I have also had to go looking for a bit more documentation. I have never used a computer running any version of *Linux*, and it's been quite a few years since I have done much with running a computer from command line entries, so I have some challenges, but I think that adds to the fun. I am sure many of you are reading this and laughing at my lack of knowledge and experience.

I have found several websites describing some fun Amateur Radio applications, and there are several ideas I want to explore. I have managed to install XASTIR (X Amateur Station Tracking and Information Reporting) and get it running. No great accomplishment, since many have already done this. My challenge was figuring out how to put some downloaded maps into the program's Maps directory. That was one of those *Linux* challenges. All of the descriptions I could find assumed that I would already know how to copy files into the root directory.

Eventually I want to acquire a small monitor that runs on 12 V and have a computer system that I can use in the car or while on camping trips. With a modem and radio I'll have an APRS station ready to go anywhere. Oh, the possibilities seem endless as I begin to learn more about this little computer.

If you are reading this and thinking about how much more advanced you are with using your Raspberry Pi, I would like to encourage you to write about some project or application that you have found. I'll bet there are many other readers who would benefit from your expertise. I'll be looking forward to hearing from you!

Michael Knitter, DG5MK

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A Different Type of Software Defined Radio — SDR Based on *Labview*

Software Defined Radio is a good example of hardware and software integration in Ham Radio. Labview*, as a graphical programming environment, keeps away from a lot of program syntax. It frees up the programmer to focus on the system from a holistic point of view. This article shows the development of a SDR based on* Labview *and common hardware frontends.*

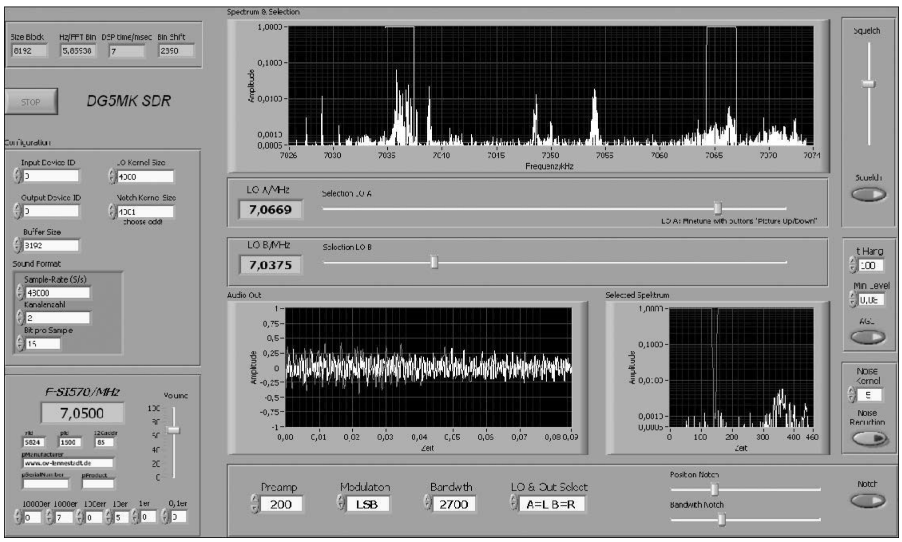

Figure 1 — Screen capture showing the SDR front panel

Some years ago, I started with SDR and was excited about the capabilities and opportunities. With minimal investment, I was able to work all types of modulation while having a very comfortable PC user interface. Frequency spectrums and their usage, image frequencies and modulation suddenly became visible.

One day my son called me to help program his LEGO NXT robot. Programming was done with a graphical program called *NXT-G* where everything operates by drag and drop. *NXT-G* showed up to be a dramatically reduced version of National Instruments *Labview*. I was surprised and amazed and an idea started to form up in my brain to create a real time SDR with such a system.

By taking self study lectures on *Labview*, some major advantages showed up compared to a classical text oriented programming environment like C. Many functions, from graphics to DSP, are available by drag and drop. I was more and more able to focus on the overall system and its functionality because the programming details are moved to the background. The block diagram you create is the program itself; the concept of a block diagram should be very familiar to many amateur radio operators.

Front Panel

Figure 1 shows the front panel of the finalized SDR receiver with signals in the 40 m band. Major features are two independent receivers, a Softrock compatible SI570 frequency control, demodulation of SSB and AM signals and the implementation of different kinds of filters. The program is able to use internal and external soundcards from common SDR hardware front ends.

Operating Mode

Figure 2 shows the *Labview* block diagram of the RX program version 1.2.1. A larger version can be found on my web page because it is too large to print clearly as part of this article.[1] It seems to be overwhelming, but notice that this is the whole program. It has to be compared to thousands of lines of code in a standard text oriented program like *C*. You can easily identify the different functional blocks in the enlarged version. Figure 3 shows the building blocks of the SDR program.

[1]Notes appear on page 7.

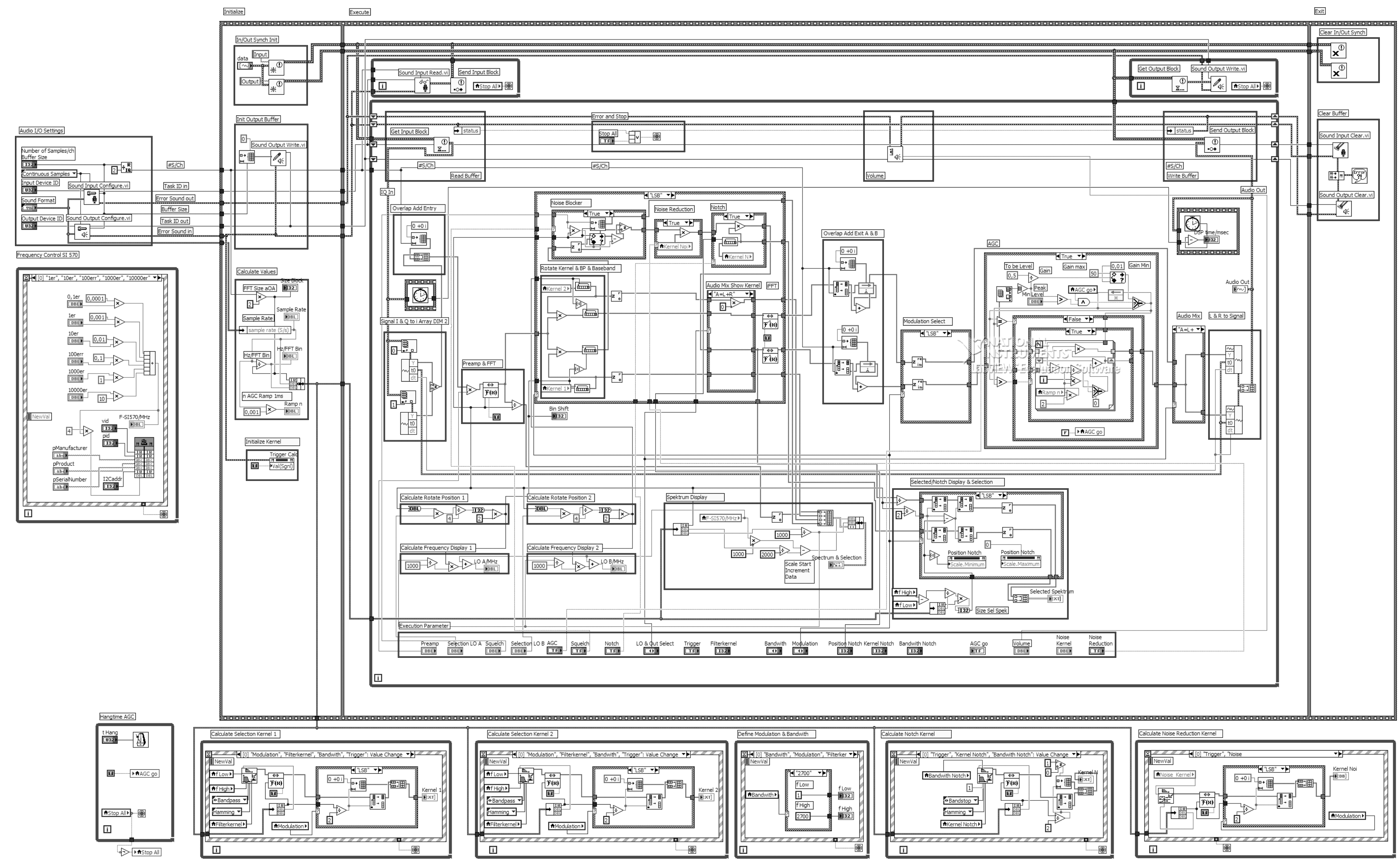

Figure 2 — This is the complete *Labview* block diagram of the RX program version 1.2.1. Obviously this is too small to read, but it will give you an idea of how *Labview* put the pieces together. For a larger version, go to my website: www.dg5mk.de. Click on the British flag on the right side of the page to see the English content.

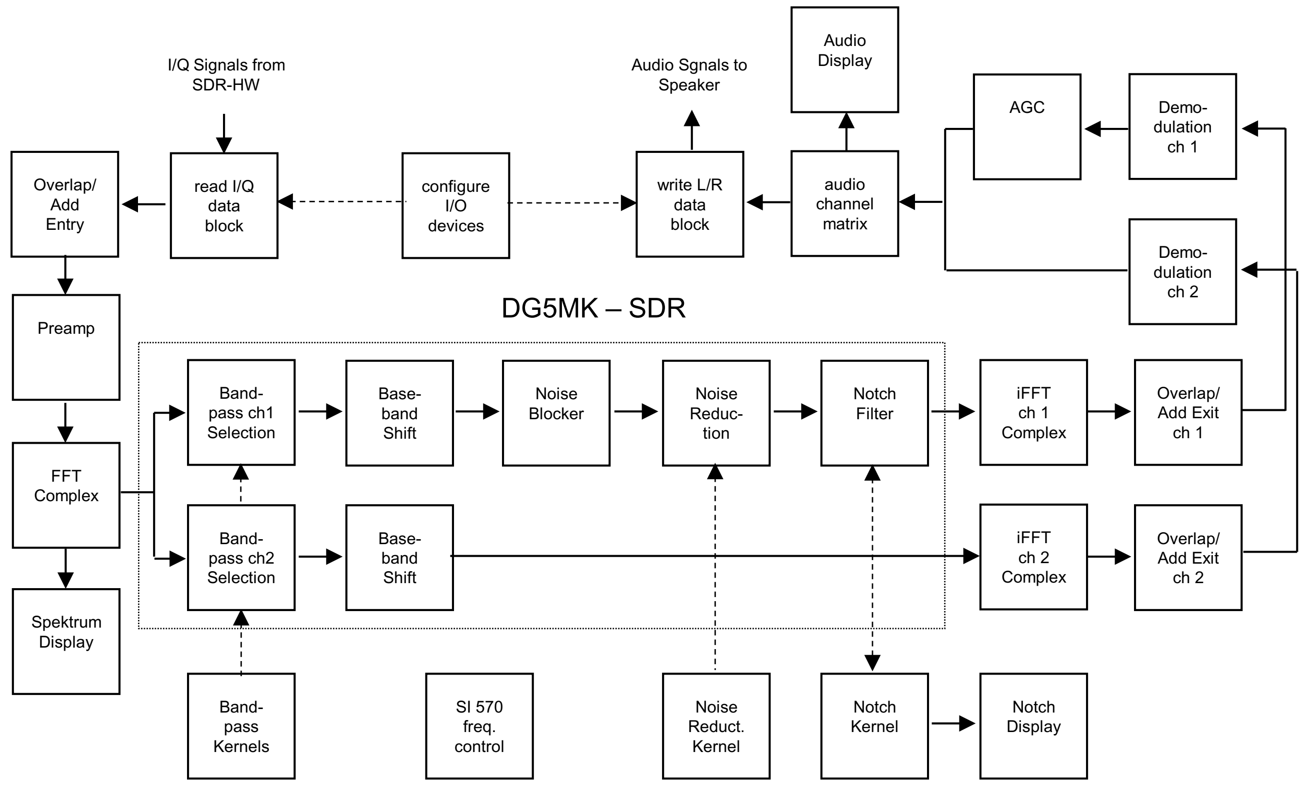

Figure 3 — SDR High Level Block Diagram

I/Q signals from Softrock compatible hardware (for example, FIFI-SDR, Softrock RX/TX, Funkamateur SDR) are processed as a sequence of overlapping blocks of data.[2] Reading and writing of these signals takes place in different program threads. This enables smooth processing with good performance. The whole processing takes place based on complex numbers. Processing digital signals as a sequence of data blocks and doing DSP changes can create major distortions to the signals. Therefore, an overlap/add procedure is implemented. You can learn more about these secrets of DSP in the references.[1, 3]

Following a simple preamplifier (just multiplication of the signals with a number), I/Q signals are transformed into the frequency domain via a fast Fourier transform (FFT). This also allows displaying the signals in a spectrum display. Most of the following DSP routines work in the frequency domain (blocks shown with dashed line data flow). This is a big difference compared with other SDR implementations. The reason to do so is higher program performance, as nearly all filtering is done via multiplication of transformed filter kernels with the signal. The filter kernels themselves will only be recalculated in case of any changes. To enable split operation on different frequencies, I have implemented a simple second receiver line without any complex filtering. Please note the second line in the block diagram.

The first processing of the signal in the frequency domain is band pass filtering. This isolates the target signal spectrum, and later it is shifted into the base band. There is an unusual kind of noise blocker (labelled Squelch in Figure 1) in the main receiver line. It works with spectral subtraction. If you control it carefully, an overlapping noise is killed with very high efficiency. As an alternative to that filter, there is also a classical moving average noise reduction filter. The next block is a notch filter, which is adjustable in terms of frequency range and target frequency. For easier use, the frequency response of that filter is shown in a separate window overlaying the selected target frequency range. The inverse FFT is used to transform the remaining signal back to the time domain. The second part of overlap/add implementation closes this part of the processing.

The demodulation of the signal depends on the selected modulation type. For AM, the demodulation is nothing more than building the absolute value or magnitude of the complex signal. The dc component is eliminated by a following filter. A new way has been used for SSB. Normally either I or Q is shifted in phase by a Hilbert transform and then I and Q are subtracted or added to each other depending on whether you want LSB or USB modulation type. Here SSB demodulation is completely done in the frequency domain within the demodulator blocks in the upper right of Figure 3. The system works by selecting a single sideband via the band pass filter and a shift to base band. Using symmetry in the FFT/iFFT processing, the real part of the complex time domain signal equals the demodulated signal. A suppression of the image frequency takes place as part of the process, again using symmetry in the FFT processing. The details of this DSP processing are beyond the scope of this article. It could be called "FFT SSB Demodulation."

In the main receiver line, the next block is AGC processing with variable hang time. During the design of the program, I found out that AGC is more art than science, as it is a totally non-linear process. Don't expect an easy answer to "How can I …?" in this context.

Next, the demodulated audio signals are displayed and can be mixed to the left and right channels in multiple configurations. The last functional block in the diagram is the output processing where each iFFT data block is sent to the soundcard.

There is also a separate control block for SI570 frequency control implemented via USB. The SI570 chip is used widely in SDR hardware to control the local oscillator. Version 1.2.1 of the RX program also includes I/Q manual imbalance correction, I/Q swap, shutdown sequences to avoid and clear *Windows* audio device errors, and saving of parameters in an XML file. In addition, there is a version 1.3.1 RX TX program, which allows SSB transmitting, SI570 PTT and tune functions in addition to receiving. These new functions are not shown in the block diagram.

Installation

Start with version 1.2.1 of the RX program! If everything works fine, you can always go to the RX TX version 1.3.1 later. A compiled EXE version of the program for *Windows* is available on my website. (See Note 1.) This requires the *Labview Runtime-Engine* (RTE) in order to run.

To use the SI570 frequency control, two additional DLLs are needed: SRDLL.dll from F. Krom, PEØFKO, and a USB DLL from the SDR hardware.[4] FIFI-SDR and Softrock RX/TX with SI570 control from DG8SAQ have been tested successfully with libusb0.dll. Both are included in the zipped archive.

So, there are three simple steps to go live:

1) Download and install the *Labview* RTE 2011 from the National Instruments web page.[5]

2) Download the EXE archive from the DG5MK web page and unzip it to a location on your hard drive. Don't change the structure or content of the folder.

3) Start the program by selecting the EXE file in the folder.

Usage

The usage of the program is simple because all of the parameters are set with meaningful values. All filters are turned off. There is an important remark about stopping the program. This has to be done via the shutdown button. If you do it differently, *Windows* will not release the allocated audio devices. In the worst case, *Windows* must be restarted. I built in some security to avoid that situation, but users are very clever in finding out other ways.

The white arrow in the tools menu is used to start the program again. The Soundformat, Soundbuffer and Dev In/Out controls should be changed while the program is running, but they will only become active after shutdown and restart of the program as all parameters are stored in an XML file.

The sample rate is an important parameter. It defines the spectral width you can work on the display. The soundcard default is 44.1 kHz and should always work. Even if the soundcard supports higher sample rates, that does not mean it always works. Some soundcards do not switch successfully to another sample rate. The background is complex and goes back to the design of the *Windows* sound subsystem. Some soundcards come with small programs that allow changing the sample rate, for example *EMU 0202 USB*. Use that program before starting the SDR.

For the RX program, change the DEV In/Out control to the incoming I/Q signals and the speaker. For the RX TX program, a second soundcard is needed. The simple one located on the mainboard is fine for the transmit channels. For decoding of DRM, PSK, and other modes, a program called *VAC* has to be used to route the channels to another decoding program. Please note that *VAC* 4.10 has some issues, but *VAC* 4.09 works fine. All other parameters can be changed during operation and do not require a program restart.

Right after starting the program there should be stations, noise, and so on in the spectral display. If not, change the level of the preamplifier or the SI570 frequency (also a one time change after starting the program). The stations can be selected by using the slider below the spectral display. Using the right modulation scheme and bandwidth, a demodulated signal should be heard from the speakers. If there are distortions, try to reset the audio system with the corresponding button. The notch filter can be adjusted with the two available sliders. Just try it; it is easy! No explanation about the noise reduction

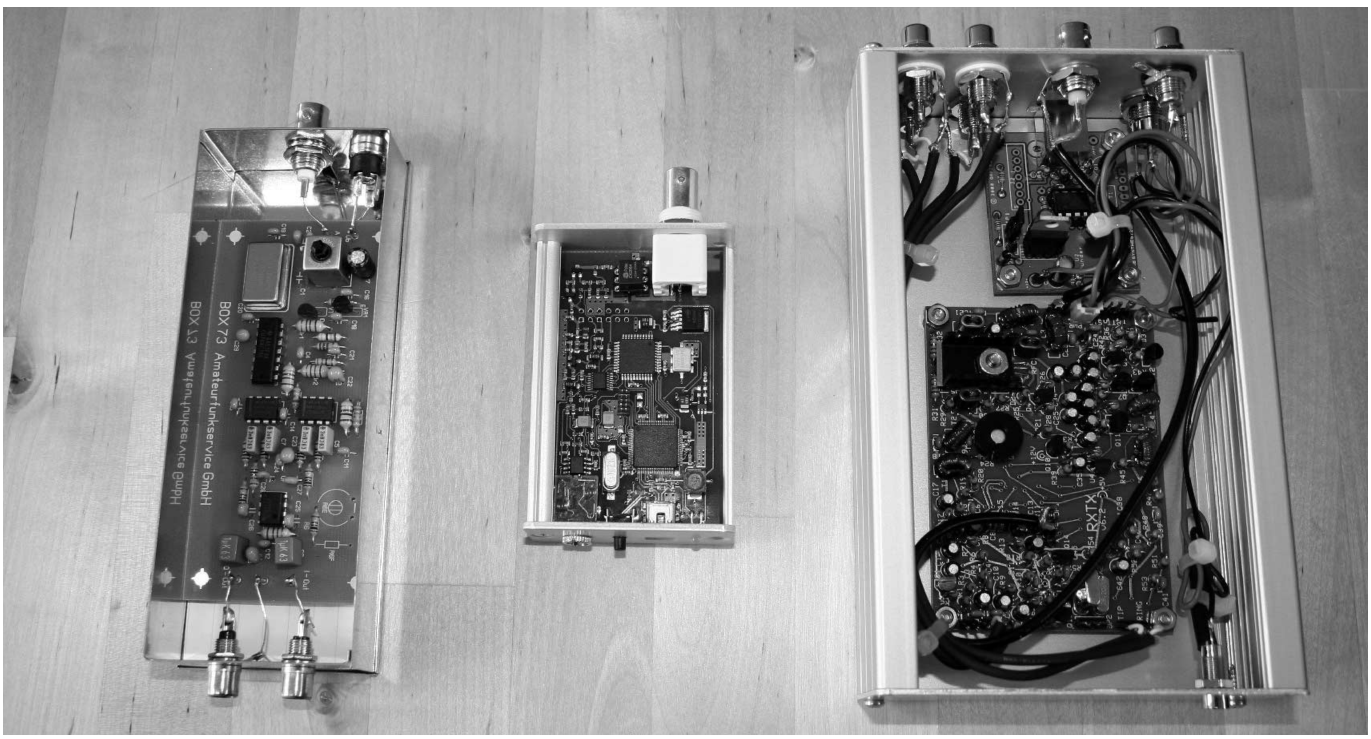

Figure 4 — Photo of the hardware used for the *Labview* SDR system. Items left to right: German Funkamateur SDR, FIFI-SDR and a Softrock 6.2 RX TX with SI 570 oscillator.

filter is needed. The noise blocker instead needs some attention. After switching it on, there is a slider to adjust the level of spectral subtraction. This level is shown by a red line in the spectral display. A level that is too high causes distortion. The right level will effectively kill the noise. With the LO & Out Select control, there are four options to mix the output channels. This will also turn on and off the second receiver line.

With versions 1.2.1 and 1.3.1, there are some additional parameters for correcting I/Q imbalance for phase and level differences. Another important new parameter is the display refresh rate. DSP and display were put to separate threads. The display processing is the single most performance consuming routine in the overall implementation. Change the refresh rate to your needs while looking at the CPU power in the *Windows* task manager! For TX the tune frequency is adjustable. Please also note that the USB parameters in the SI570 section need to fit to your hardware. The RX version 1.2.1 has the FIFI-SDR as the default hardware. The RX TX 1.3.1 version has the Softrock RX TX with DG8SAQ interface set as the default.

Experience and Perspective

The *Labview* SDR program was extensively tested mostly using SSB and AM on 40 m, 80 m, and 20 m with a FIFI-SDR and a Softrock RX/TX 6.2 with SI570 oscillator. The antenna used was an inverted-V dipole. Digital modulation like PSK 31 and DRM with the *DREAM* software was also successfully tested. Figure 4 shows the SDR hardware used.

It is amazing the results that can be achieved with such a software implementation. Without any formal measurement, this system seemed to outperformed a mid-price analog Amateur Radio Transceiver.

The design and the development of the *Labview* SDR provided a lot of fun to me. Maybe in the future I will implement something like direct PSK 31 coding and decoding, or a cross platform audio API like *Portaudio*. It was never planned to build something commercial or even reach the established programs with years of development like *Power-SDR*. The implicit goal was to prove that full SSB modulation and demodulation can be done completely in the frequency domain. Having this goal, I learned worlds about DSP.

If you would like to get more details on the *Labview* SDR, please have a look at the DG5MK web page. For *Labview* you should also look at the articles by Giorgi and Mütterlein.[6, 7] The Labview source code (VI) is also available for download from my web page. Unfortunately, I do not have the time to translate everything from German to English, but the source code is documented in English.

For more information on digital signal processing, please look at the articles by Smith, Smith, and Lyons.[3, 8, 9] A special focus on SDR DSP is described in the articles by Lyons and Youngblood.[10, 11, 12] Most of it is in English.

Fell free to contact me in case of any questions. I would also be interested in your experience with the program.

Have fun!

Michael Knitter, DG5MK, works in an international computer company in sales and distribution. He earned a degree in telecommunication technology from the University of Dortmund. Besides his professional career in a very different area, he never moved away from electronics and radio communications. Michael has been a licensed radio amateur since 2006. Special areas of interest are software defined radio, digital signal processing, filter design, magnetic loop antennas, microcontrollers, C++ and Labview *programming. He loves to sail as a true contrast to modern busy life.*

Notes

[1]Figure 1 is displayed on the author's web page, **www.dg5mk.de**. Click on the British flag on the right side of the page to see the English content.

[2]Funkamateur, BOX 73 Amateurfunkservice, 40 m Einsteiger SDR, **www.box73.de**.

[3]S. Smith, *Digital Signal Processing. A Practical Guide for Engineers and Scientists*, Elsevier Ltd, Oxford, 2002, also available as a free download at **www.dspguide.com**.

[4]F. Krom, PEØFKO, Softrock DLL, **home.ict.nl/~fredkrom/pe0fko/CFGSR/**.

[5]Search "LabVIEW RTE" to download the runtime engine for Windows x86, **www.ni.com**.

[6]W. Georgi; E. Metin, *Einführung in LabVIEW*. Carl Hanser Verlag, 4. Auflage 2009.

[7]B. Mütterlein, *Handbuch für die Programmierung mit LabVIEW*, 1. Auflage 2007.

[8]D. Smith, "Signals, Samples and Stuff: A DSP Tutorial - Part 1 – 4," *QEX*, March 1998 – September 1998, **www.arrl.org/dsp-digital-signal-processing**.

[9]R. Lyons, *Understanding Digital Signal Processing*, Prentice Hall International, 2004.

[10]R. Lyons, *Quadrature Signals: Complex, But Not Complicated*, 2008, **www.dspguru.com/sites/dspguru//files/QuadSignals.pdf**.

[11]R. Lyons, "Quadratursignale: Komplex, aber nicht kompliziert". Dt. Übersetzung DL6KBF 2011, **www.needles.de/HPSDR/QuadSignals-DE.pdf**.

[12]G. Youngblood, "A Software Defined Radio for the Masses, Part 1 – 4," *QEX,* July 2002 to April 2003, **www.flex-radio.com/News.aspx?topic=publications**.

Michael Dzado, ACØHB

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An Eight Channel Remote Control Antenna Selector

Select between eight antennas or feed one antenna to any of up to eight radios. With better than 70 dB of port-to-port isolation, you can be sure the signal is going where you want it to.

A few years ago, my friend Joe Spinks, AAØKW, and I started building and experimenting with Double Bazooka antennas. I decided to build two 20 m and two 40 m antennas for my antenna farm. My plan was to deploy a 20 and 40 m Double Bazooka facing East-West and a 20 and 40 m Double Bazooka facing North-South. These, along with a vertical antenna and two G5RV antennas I already had mounted, quickly gave me a cabling and switching problem. Also, Joe pointed out that my wife may not appreciate me punching seven more holes in our house and running a sizable bundle of coax across my basement to my station. Even if I could do all that, manual switching wasn't practical between that many antennas. I would be constantly connecting and disconnecting antennas when I wanted to change directions or bands.

That's when I decided to design a remote control antenna selector to select between eight antennas. I am currently a Systems Engineer but have degrees in both Electrical Engineering and Software Engineering. In addition, I have electronic circuit design and printed circuit board layout experience. Joe is also an electrical engineer, and has designed automatic antenna tuners for our company and has extensive RF circuit design experience. So creating and testing a viable design was not a technical concern. The main question then became, what improvements could be made over the existing products. The answer came quickly: Isolation between the selected and other antennas! Wouldn't it be nice to select one antenna and not get interference from another antenna?

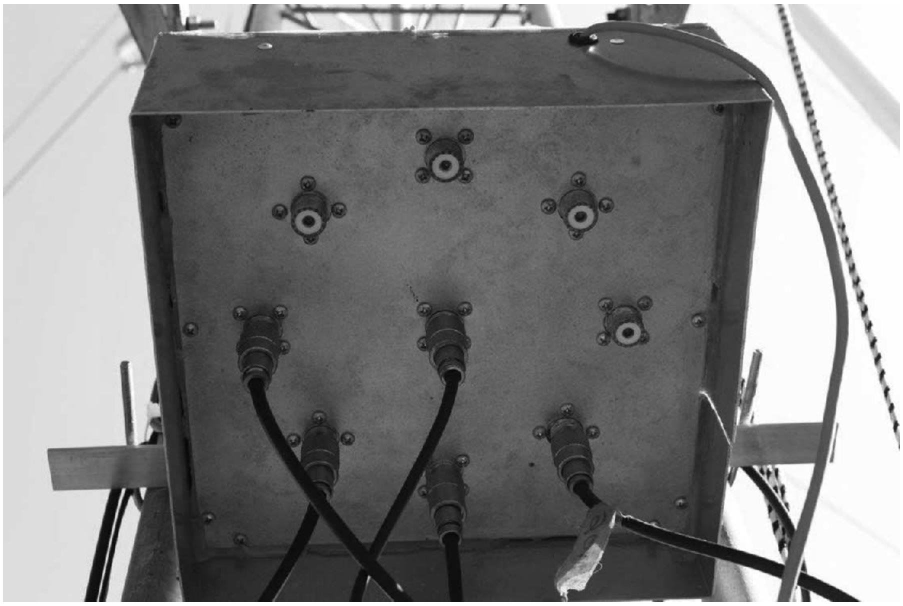


Figure 1 — The remote antenna selector board is mounted on the side of my tower. As shown here, it is selecting between four antennas.

We achieved greater than 70 dB of isolation between antenna ports, as shown in the test plots included with this article. There are several design techniques that helped us achieve this kind of isolation: Two relays were used in each signal path to double isolate the antenna from the radio input. The traces that make up the RF path were designed as a coplanar waveguide. The RF connector placement on the circuit board was tightly controlled to be symmetrical to make the RF electrical paths identical. RF trace routing followed good design practices by limiting angles to 45°. Control traces were routed on the bottom of the circuit board to maximize ground plane continuity. For our switch, we selected control trace widths of 15 mils for better current capacity.

Implementing the Design

I mounted a switch assembly on my tower and connected four antennas to the assembly as shown in Figure 1. My home station consists of an Icom IC-706 MKIIG rig, with an LDG tuner and our remote control antenna selector, as shown in Figure 2. The remote controller is shown in the lower left. Switching between eight antennas is just a matter of twisting the rotary switch. The real story is the isolation between antennas. Figure 3 shows the IC-706 S meter displaying S6 when an antenna is selected. Figure 4 shows the IC-706 S meter displaying blank when an unused port is selected.

One evening in my "Lab," Gregg Lind, KCØSKM, noticed my design and board layout. Gregg and I have been collaborating on a remotely deployable solar powered Weak Signal Propagation Reporting (WSPR) station using a Netduino. Gregg immediately saw multiple applications for our project and had me present it to both of our local clubs, Cedar Valley Amateur Radio Club (CVARC) and Collins Amateur Radio Club (CARC) here in Cedar Rapids Iowa. He quickly took orders from 20 hams that were interested in purchasing our project if we offered it as kit. The kit was then featured during one of our club's annual kit build nights.

The project was a big hit with both clubs. It turns out that I was not the only one grappling with antenna management. Besides, many amateurs like to build kits. Gregg used this success to convince me to write this article to invite collaboration from other Amateur Radio enthusiasts throughout the ham community.

Later, Gregg used our switch to solve a radio and antenna management problem in our club's shack. Our club has multiple radios and multiple antenna options, which require an operator to trace cables behind a huge rack and make the connection manually. Automatic selection of radios and antennas would make the station much more user friendly. The application required two remote controlled antenna selectors, one to select a radio connected to another antenna selector that picks the desired antenna. Since the installation, our station usage has increased dramatically. The first switch selects one of eight radios, the second switch selects one of eight antennas. Figure 5 shows the two RF switch assemblies used in the CARC (NØCXX) station. Currently the assembly is managing the selection of four radios to six antennas.

Design Details

The remote control antenna selector consists of a switch assembly and an optional remote controller assembly.

The switch assembly contains the relays and relay drive circuitry to select between position 0 to 7. The assembly only requires a 13.8 V dc supply and a 3-bit TTL signal to input for the switch selection. This interface allows for a variety of remote control solutions. Figures 6 and 7 show the bottom and top views of the RF switch assembly.

A remote controller assembly was designed as a simple solution for my station. The remote controller assembly consists of an eight position rotary switch and an 8-to-3 digital encoder connected to 2N2222A transistors to drive the switch assembly via a standard CAT-5 eight wire cable. The remote controller Assembly is shown in Figure 8.

As mentioned earlier, one major design consideration was to minimize RF coupling between antenna channels. Therefore, it was essential to follow the rules and principals of good basic RF/Microwave design.

The switch assembly features a coplanar

Figure 2 — This photo of my operating position shows my power supply, LDG Autotuner, and Icom IC-706 MKIIG radio. The antenna selector control panel is mounted behind a block of wood to match the operating position shelf. You can see the selector shaft, with the first of eight LEDs illuminated to show which antenna has been selected. I still have to add labels below the LEDs to help identify the antennas.

Figure 3 — Here is a close-up of the Icom radio display. Notice that with a 40 m antenna connected, the S meter is showing an S6 signal.

Figure 4 — In this close-up of the Icom radio display, there is no antenna connected. In this case, the S meter is showing no signal.

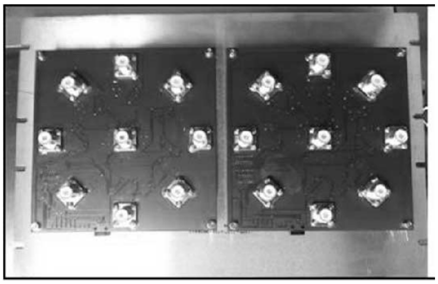


Figure 5 — This photo shows the two RF switch assemblies used in the Collins Amateur Radio Club (NØCXX) station. The assembly is managing the selection of four radios to six antennas in the club station.

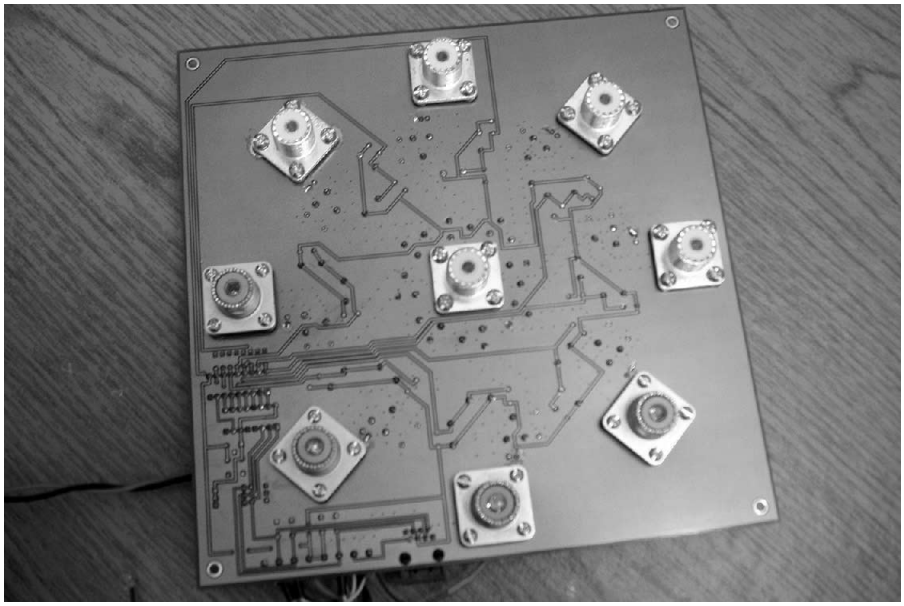


Figure 6 — Here is a close-up of the SO-239 connector side of the antenna selector circuit board.

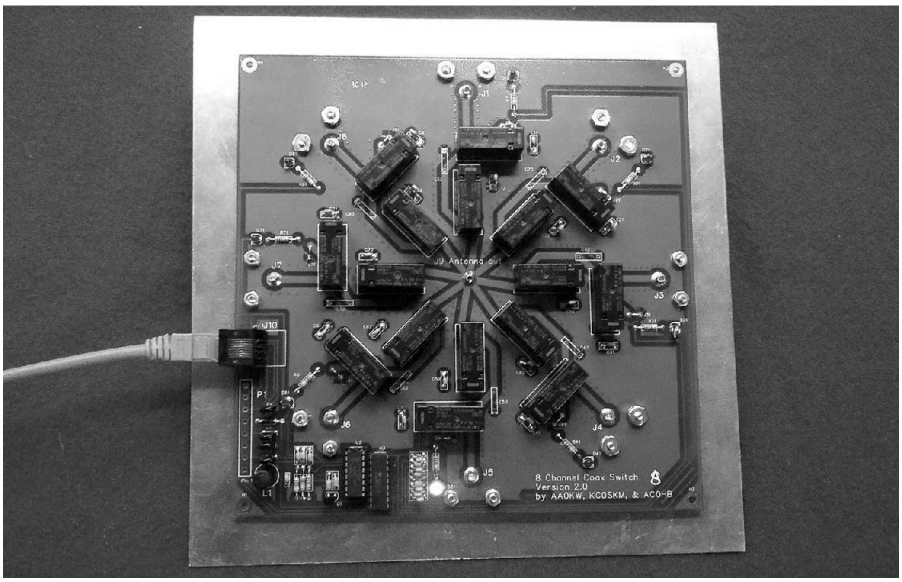


Figure 7 — This photo shows the relays and the circuit board traces that form the coplanar waveguide.

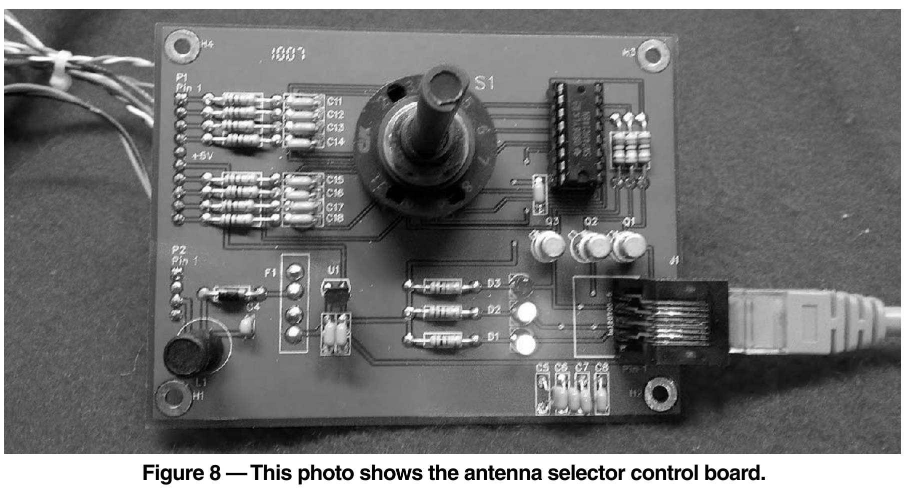


Figure 8 — This photo shows the antenna selector control board.

waveguide design for all RF traces, tuned to a 50 Ω impedance with ground plane stitching that ensures maximum isolation between ports.

Complete impedance (Z_0) matching (50 Ω in to 50 Ω line to 50 Ω out) minimizes return loss and SWR. A coplanar waveguide design was chosen so that the trace impedance on the circuit board could be matched to the input and output impedances. In a coplanar waveguide design, Z_0 is a function of signal conductor width & thickness and a function of the dielectric constant (ε_r) of the material surrounding the signal conductors.

Signal return currents follow the path of least impedance. In high frequency circuits this equates to the path of least inductance. Stitching the ground planes with vias every 0.1 inches or so around each RF trace helps minimize the inductance in the signal return path by virtually creating a waveguide on the circuit board.

The RF connectors were placed symmetrically around the output connector (located in the center of the assembly) to ensure an equal electrical length for each RF path. Typical isolation measured between ports is greater than 70 dB.

Extensive RF decoupling on the power and control lines was added to provide maximum RF decoupling from the control signals. I added transient-voltage-suppression diodes (Transorbs) on all input control lines for good surge protection.

The project begins with a schematic program to capture the logical design. I chose *TinyCAD* for my schematic capture and *Free PCB* for the circuit board layout and trace routing. Both tools are easy to use and use the same net list format.[1, 2]

The switching relay is the heart of this design. We chose an Omron Electronics Inc G6RN-1-DC12, which is a sealed double pole double throw (DPDT) relay with 8 A silver

[1]Notes appear on page 17.

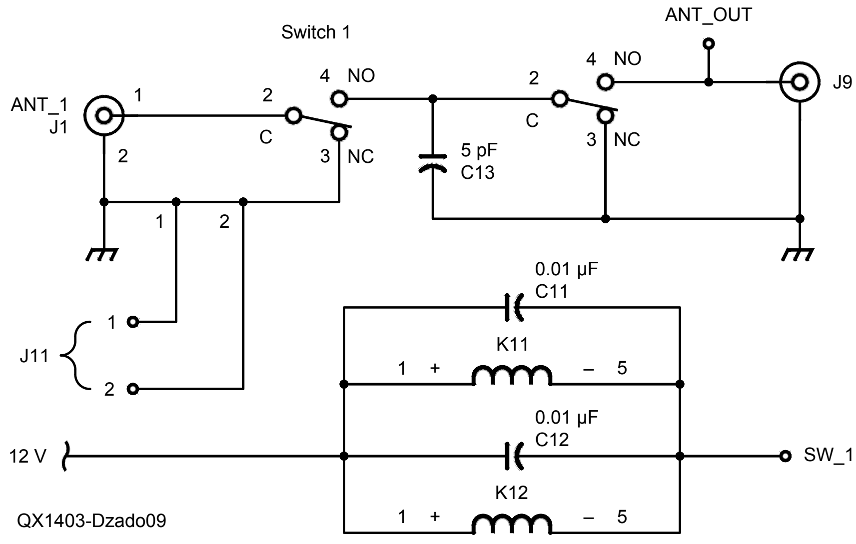


Figure 9 — The schematic of the basic selector switch operation. The normally closed side of the relay grounds the antenna when it is not selected. The second relay serves to isolate that antenna port from the output signal on a selected antenna.

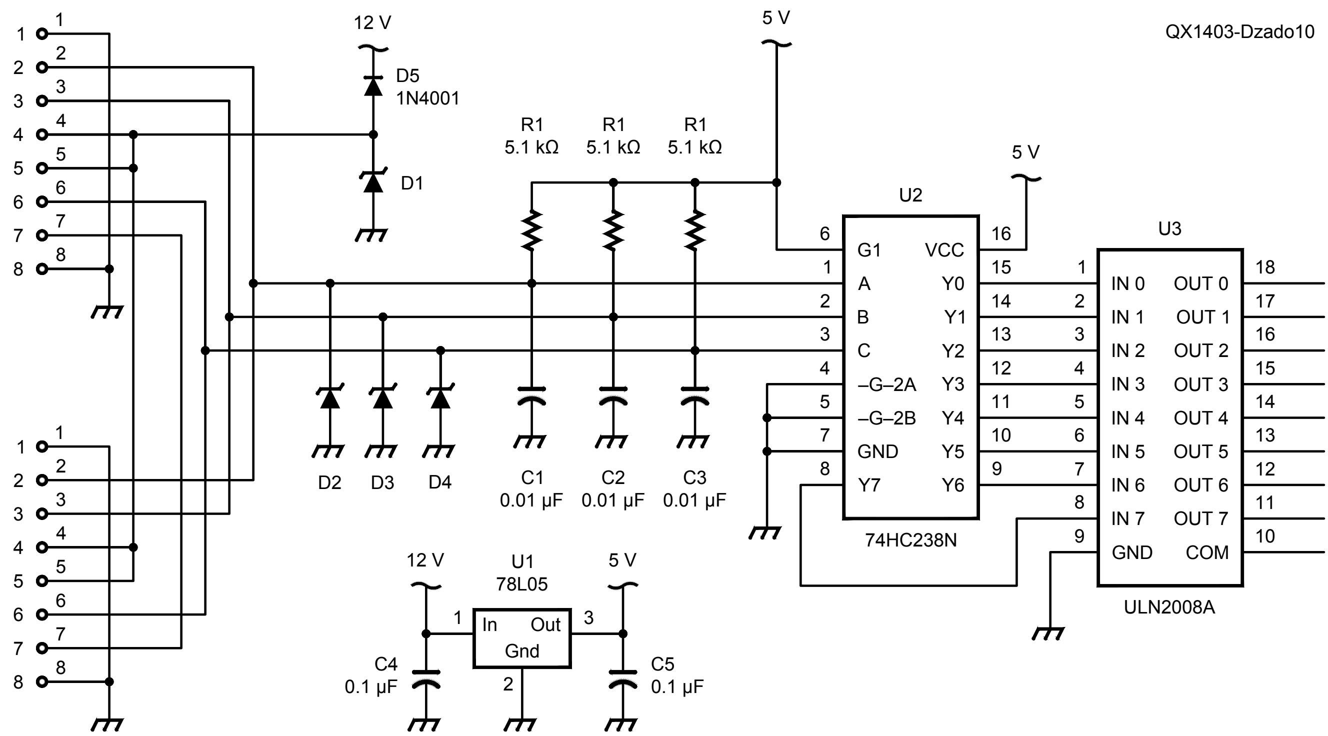


Figure 10 — The schematic of the antenna selector control logic. U2 is a 74HC238 3-to-8 decoder IC and U3 is a ULM2008A high current, 8 pair Darlington transistor array to provide the relay drive current.

over gold contacts. The contact capacity is more than adequate for our design. We chose to double isolate each RF port by using two relays in each path. As shown in Figure 9, the first relay will ground the antenna input when de-energized. The second relay simply isolates the RF port from the output.

The switch control logic incorporates a simple 3-to-8 decoder (74HC238) and a high current, eight-pair Darlington transistor array (ULM2008A) to provide the relay drive current, as seen in Figure 10.

Topologically speaking, a coplanar waveguide design offers better isolation between signals versus a microstrip design. Even though a microstrip design is easier to layout, Joe and I opted for maximum isolation between antennas.

The RF path design starts with the basic power equation to determine the trace width to handle the transmit current.

$$P = I^2 Z_0 \qquad \text{[Eq 1]}$$

then solving for I:

$$I = \sqrt{\frac{P}{Z_0}} \qquad \text{[Eq 2]}$$

Then for a 100 W transmitter into a 50 Ω load we have:

$$I = \sqrt{\frac{100\ \mathrm{W}}{50\ \Omega}} = \sqrt{2}\ \mathrm{A} = 1.4\ \mathrm{A}$$

Using the *AppCAD Coplanar Waveguide Calculator* by Avago Technologies, I set the known dimensions of the circuit board and adjusted the trace width, W, and gap, G, until $Z_0 = 50\ \Omega$, or there about, as shown in Figure 11.[3] I chose the trace width of 115 mils and a gap of 100 mils as a good combination for the 100 W circuit boards. Note that other combinations of trace width and gap will also result in a Z_0 of 50 Ω. For example, a trace width of 125 mils and a gap of 250 mils would also work, but that requires quite a bit more space on the circuit board.

Referring to Table 1 for the circuit board trace current capacity, a 115 mil trace is adequate for this design. You can see that at 100 W, the trace temperature rise will be considerably less than 10°C. In fact, with a power of 1000 W into a 50 Ω load, the current would be 4.47 A, so these traces can handle that power with only about a 10°C temperature rise.

Figure 12 shows the printed circuit board component placement and trace routing.

I used an HP-8753D Network Analyzer to test the completed circuit board. Measurements were taken between all eight channels for port-to-port isolation, insertion loss, and SWR.

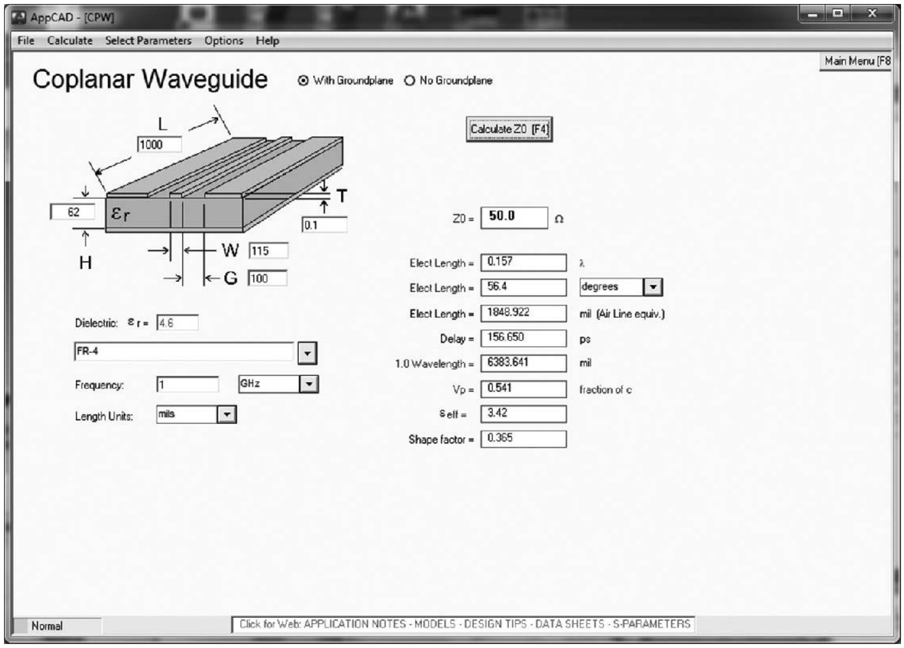


Figure 11 — This is a screen shot of the *AppCAD Coplanar Waveguide Calculator*. I set the following parameters: Circuit Board Material = FR-4, W (Trace Width) = 125 mils, H (Circuit Board Thickness) = 62 mils. Then I adjusted the trace width (W) and the gap between the traces (G) until the calculated Z_0 came to 50 Ω. A width of 115 mils and a gap of 100 mils gave the desired impedance. Other combinations of those dimensions may also result in a 50 Ω impedance, but this combination gave a reasonable trace width for the power handling capability that I wanted.

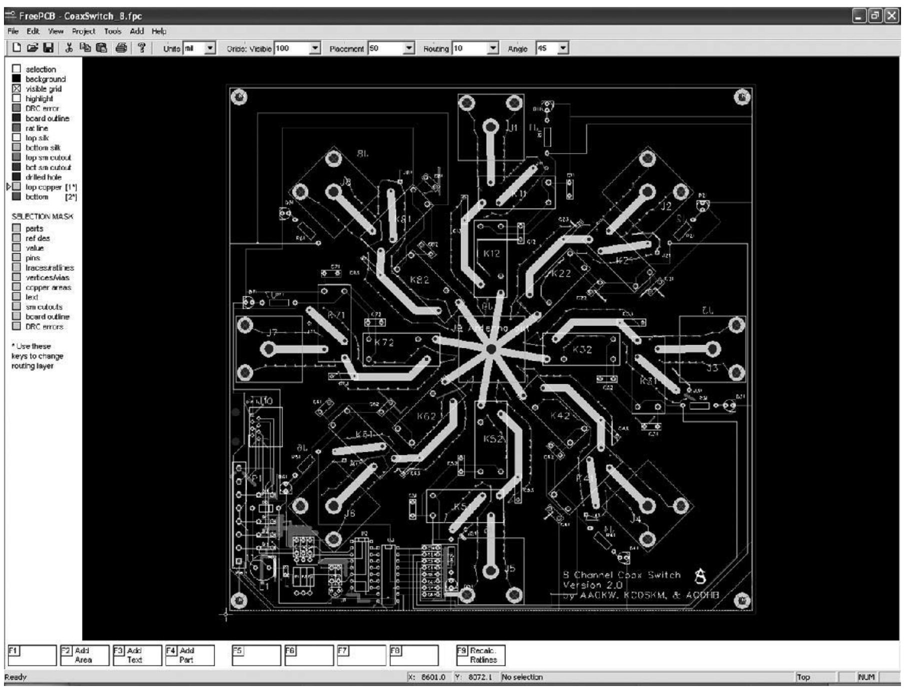


Figure 12 — This is the parts-placement view of the antenna selector circuit board. It also illustrates the coplanar waveguide traces.

Figure 13 shows the test results for port to port isolation between channels 1 and 2. Figure 14 shows the insertion loss for port 1 and Figure 15 shows the SWR for port 1.

The test plots for the remaining ports are almost identical to the port 1 test results. This is due to the coplanar waveguide topology for the RF traces and the symmetrical RF component placement.

Specifications

- **Power Requirements:** 13.8 V dc single power supply at less than 75 mA.
- **Control Line:**
  - Minimum 5-wire connection (2 power, 3 control lines) or
  - Standard CAT-5 cable (8 wire) interconnect between RF Switch and Remote Controller.
- **Switches:** Sealed RF Relays, Contacts are silver over gold for an 8 A contact rating.
- **Status LED indicators:** on both the switch and remote control assemblies.
- **Impedance:** 50 Ω.
- **Connectors:** SO-239 Silver plated Teflon connectors.
- **RF Power:** 1000 W over 2:1 SWR.
- **SWR:**
  - At 30 MHz, < 1.12:1
  - At 50 MHz, < 1.23:1
- **Port to Port Isolation:**
  - At 30 MHz, < –75 dB
  - At 50 MHz, < –70 dB
- **Insertion Loss:**
  - At 30 MHz, < 0.10 dB
  - At 50 MHz, < 0.16 dB
- **Dimensions:**
  - Switch Circuit Board: 8 × 8 inches.
  - Enclosure: 9.5 × 9.5 × 2 inches.
  - Remote Controller: 3 × 4 inches.

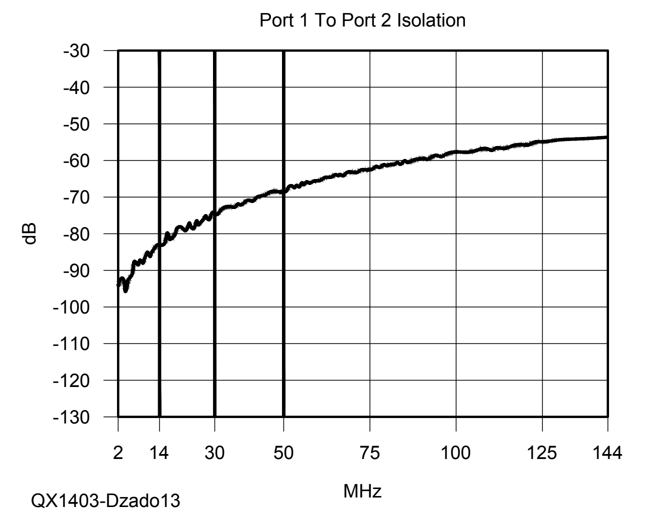


Figure 13 — This graph shows the port-to-port isolation between positions 1 and 2 versus frequency. The test plots for the remaining ports are virtually identical.

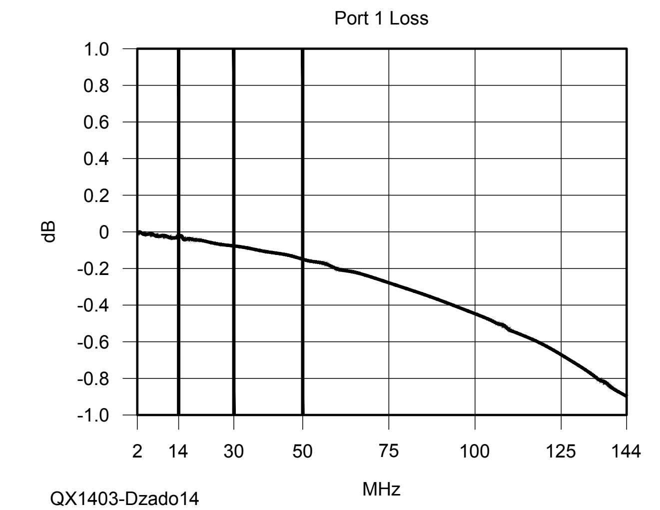


Figure 14 — This graph shows the signal response (S_{11} parameter) versus frequency for port 1. The responses for the other ports are virtually identical.

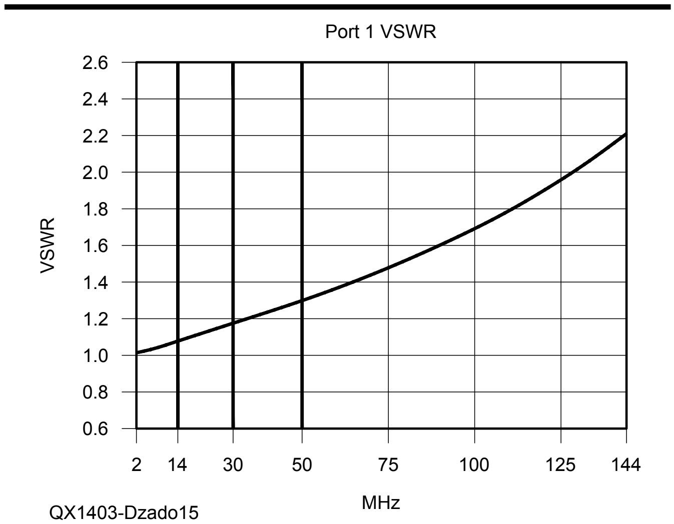
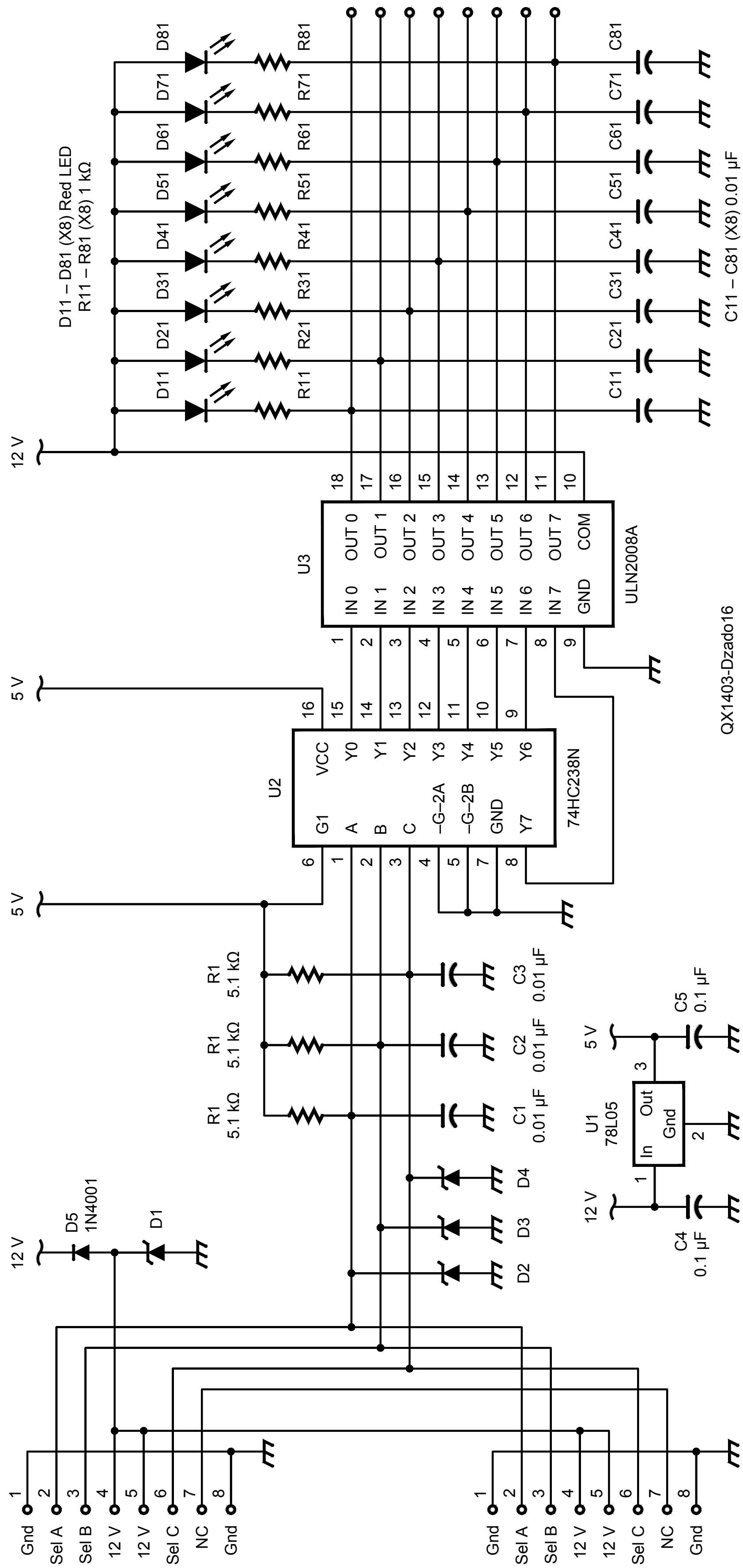


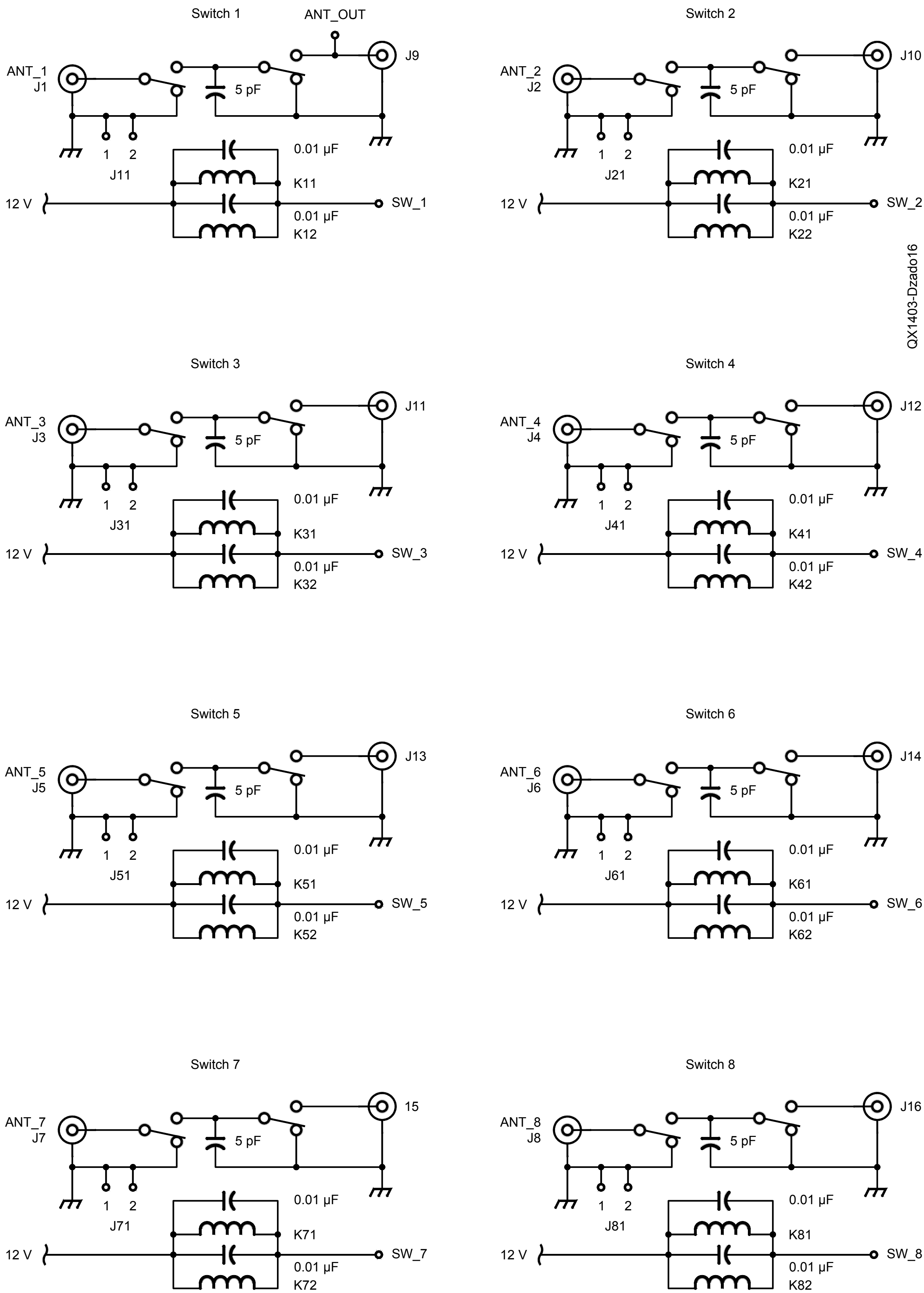
Figure 15 — Here is the SWR response versus frequency as measured for port 1. The other ports have the same response.

Table 1
Circuit Board Trace Widths

Temp Rise / *Width (mils)*	*10°C* / *Max Current (A)*	*20°C*	*30°C*
10	1	1.2	1.5
15	1.2	1.3	1.6
20	1.3	1.7	2.4
25	1.7	2.2	2.8
30	1.9	2.5	3.2
50	2.6	3.6	4.4
75	3.5	4.5	6
100	4.2	6	7.5
200	7	10	13
250	8.3	12.3	15

Figure 16 — This schematic shows the control wiring and the antenna selector wiring for all eight antenna positions.




Switch 1
ANT_OUT
ANT_1
J1
J9
5 pF
1
2
J11
0.01 µF
K11
12 V
SW_1
0.01 µF
K12
Switch 2
ANT_2
J2
J10
5 pF
1
2
J21
0.01 µF
K21
12 V
SW_2
0.01 µF
K22
QX1403-Dzado16
Switch 3
ANT_3
J3
J11
5 pF
1
2
J31
0.01 µF
K31
12 V
SW_3
0.01 µF
K32
Switch 4
ANT_4
J4
J12
5 pF
1
2
J41
0.01 µF
K41
12 V
SW_4
0.01 µF
K42
Switch 5
ANT_5
J5
J13
5 pF
1
2
J51
0.01 µF
K51
12 V
SW_5
0.01 µF
K52
Switch 6
ANT_6
J6
J14
5 pF
1
2
J61
0.01 µF
K61
12 V
SW_6
0.01 µF
K62
Switch 7
ANT_7
J7
15
5 pF
1
2
J71
0.01 µF
K71
12 V
SW_7
0.01 µF
K72
Switch 8
ANT_8
J8
J16
5 pF
1
2
J81
0.01 µF
K81
12 V
SW_8
0.01 µF
K82

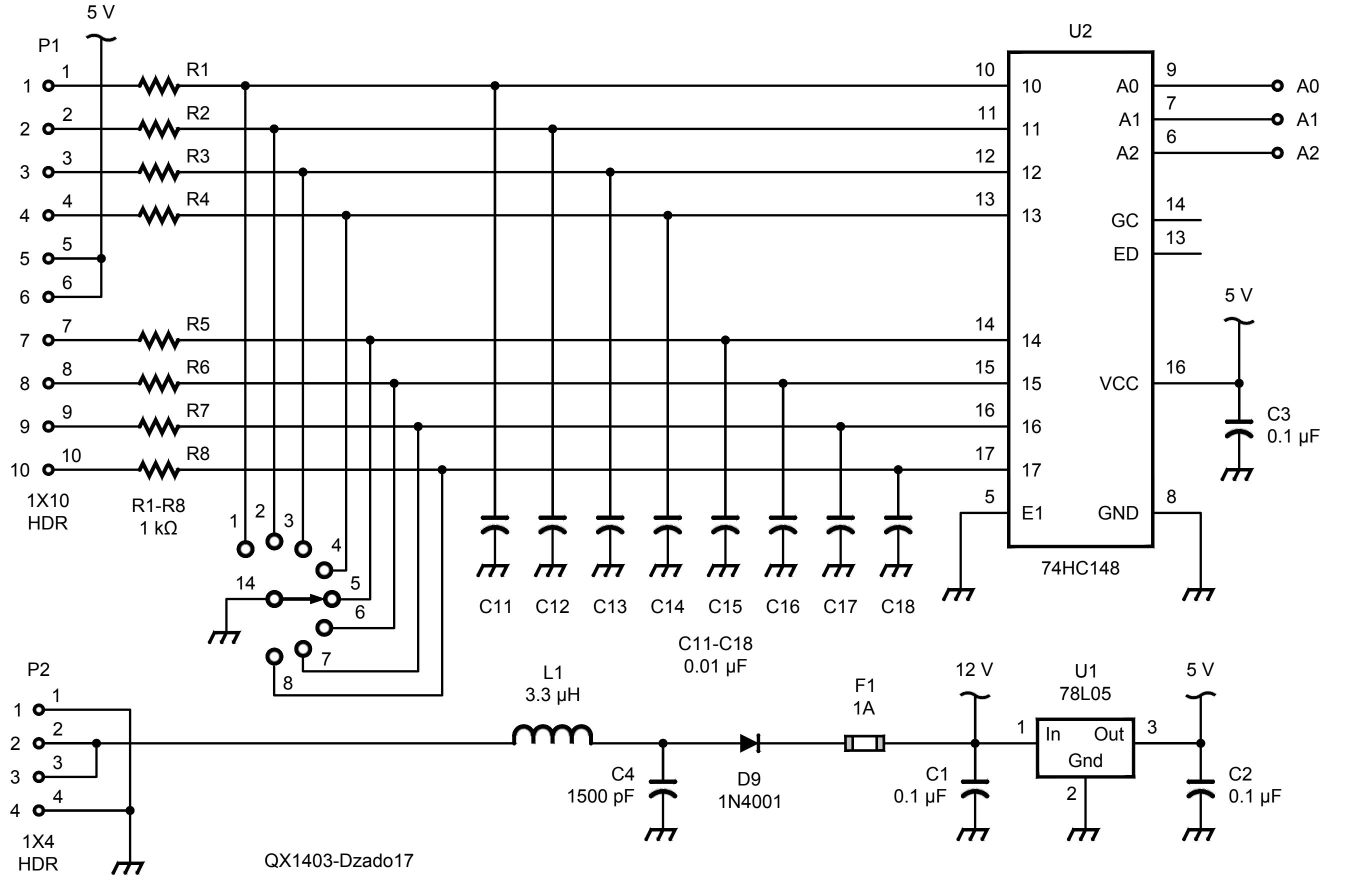


Figure 17 — Here is the antenna selector switch remote control wiring.

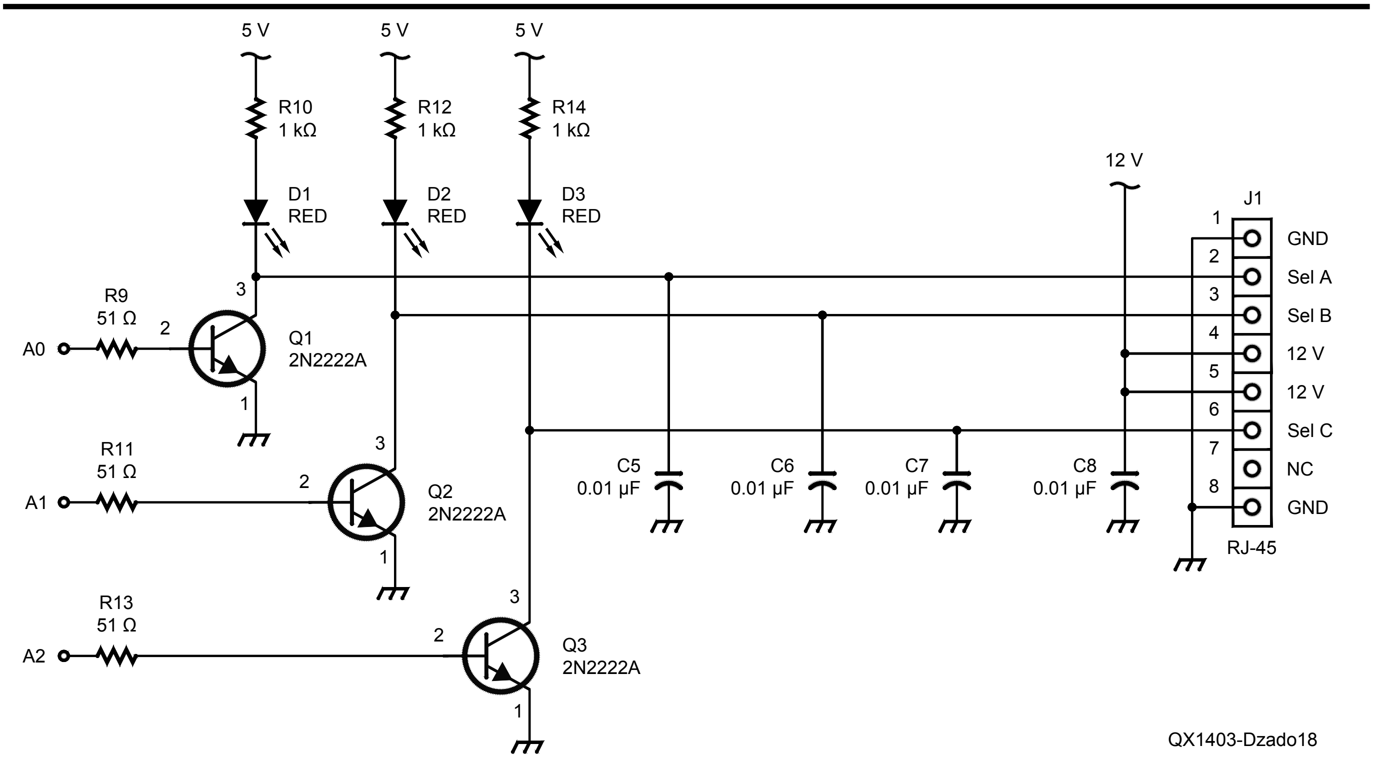


Figure 18 — This schematic diagram shows the remote control wiring output to the antenna selector switch.

Mike Dzado, ACØHB, obtained his Amateur Extra Class License in 2007. Mike has over 35 years combined experience as an Electrical and Software Engineer, which includes CAE/CAD experience. Mike started his career as an Electronic Technician with the USAF and currently is a Senior Systems Engineer for Rockwell Collins Air Transport Large Display Systems. Mike holds an Associate Degree in Meteorological Equipment from the Community College of the Air Force (1978), Bachelor of Science in Electronic Systems Management from Southern Illinois University (1980), Bachelor of Science in Electrical Engineering from the University of Utah (1984), and a Master of Science in Software Engineering from the National Technological University (1994). Mike's career experience includes Power Supply Design, Communications Protocol Software Development, Computer Aided Engineering Design and Analysis Tool development, and Software Defined Radio development.

Mike's other hobbies include weather spotting, and designing and building electronic kits for fellow club members. He is a member of the Cedar Valley Amateur Radio Club (CVARC) and Collins Amateur Radio Club (CARC) See his website, **www.AC0HB.com** *for some of his current projects.*

Notes

[1]Learn more about *TinyCAD* and download the program free at: **tinycad.en.softonic.com**/.

[2]For more information about the *Free PCB* program and to download the installation files, go to: **www.freepcb.com/**.

[3]You can download the *AppCAD* design software free from the Avago Technologies website. Go to **www.avagotech.com/pages/appcad**.

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Actual Measured Performance of Short, Loaded Antennas — Part 2

With the help of many friends over many years, the author studied HF monopoles used as verticals, mobile antennas and in pairs as elements of beams and dipoles.

What are the Bottom Line Numbers?

In this second part of the article, I present the actual measured results for our Series 1 and Series 2 tests. The field strength numbers throughout the Tables are comparisons to a perfect, zero-loss ground-plane antenna with a ¼ λ resonant vertical monopole. This is the "zero" point or benchmark. As you read the charts, keep in mind that the least negative field strength number is the most desirable, because it represents how much weaker the test antenna is than a perfect monopole/ground-plane antenna on that frequency. These tests were conducted in Fletcher, North Carolina and in Harlingen, Texas, and repeated many times over several years. The deviation was very small. These are the averaged numbers from dozens of Series 1 and Series 2 runs.

Each run through Series 1 and Series 2 resulted in nearly 300 measurements. When excursions to other bands occurred, the number of measurements increased proportionally. The two programs resulted in many thousands of measurements. Field intensity readings were converted to decibels and all data was collected, entered into the computer and printed out each day by Arch Doty, K8CFU/W7ACD.

Series 1

How the Position and Q of the Coil in a Shortened Monopole Affects Efficiency

In this series, the test antenna was a fixed length of 8½ feet. Starting with the loading coil at the very top of the mast, a balanced horizontal capacitance above the loading coil was adjusted for resonance. Field strength and all other measurements were collected. Then the coil was moved down 24 inches, the antenna was adjusted for resonance again, and all data collected. Then, down 24 inches more, then another 24 inches, and finally the coil was installed at the base. Figure 22 illustrates the variations in antenna configuration for these tests.

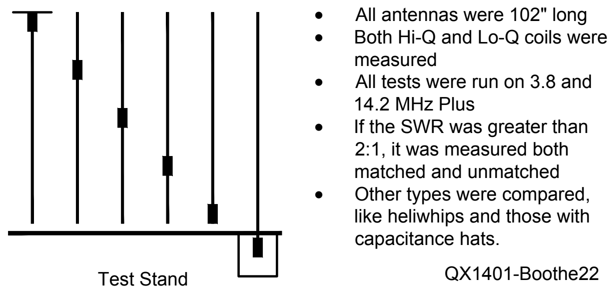

Figure 22 — This drawing illustrates the Series 1 test antenna configurations.

This was done using high-*Q* coils and then low-*Q* coils on both 14.2 MHz and 3.8 MHz, with occasional excursions to 1.8 MHz through 21 MHz to insure the trend was uniform on all the lower ham bands. Besides the base loading position, one additional configuration was added. That was where the loading coil was *below* the test stand in a shielded box, to simulate some of the commercial autotuner and "in the trunk" mobile installations as well as fixed monopoles, base loaded with shielded tuners.

All the high-*Q* coils in our tests were made using #12 or #10 silver tinned copper air core coil stock with spaced turns, either 2, 3, or 4 inches in diameter. Our low-*Q* coils were all either #20 enameled copper (1.8 MHz and 3.8 MHz) or #18 enameled copper (7.2 MHz to 21 MHz). They were close wound on either a PVC or paper phenolic form. Coils for 1.8 and 3.8 MHz were 7/8 inch in diameter, while the 7.2 and 10.1 MHz coils were 5/8 inch in diameter, and those for 14.2 to 21 MHz were 3/8 inch in diameter. Table 1 shows the results on 20 m and Table 2 shows the results on 80 m.

[1]Notes appear on page 31.

Table 1
Series 1 Bottom Line Results
20 Meters, 14.2 MHz

Antenna Configuration	*Resistance at Resonance (Ω)*		*2:1 Bandwidth (kHz)*		*Field Strength (dB) Below Reference Antenna*	
	High-Q	*Low-Q*	*High-Q*	*Low-Q*	*High-Q*	*Low-Q*
¼ λ No Coil	53		690		−2	
102” Coil At Top	42	42	340	456	−2.8	−2.8
102” Coil At 72”	36	35	353	478	−3.3	−3.3
102” Coil At 48”	31	30	361	509	−4.7	−4.7
102” Coil At 24”	27	27	349	490	−5.9	−5.9
102” Coil At Base No Match	23.5	23.5	----	----	−7.5	−7.5
102” Coil At Base Matched	50	50	390	580	−6.5	−6.5
102” Coil Shielded at Base No Match	20.5	20.5	-----	------	−14.2	−14.2
102” Coil Shielded at Base Matched	50	50	382	572	−12.2	−12.2

Table 2
Series 1 Bottom Line Results
80 Meters, 3.8 MHz

Antenna Configuration	*Resistance at Resonance (Ω)*		*2:1 Bandwidth (kHz)*		*Field Strength (dB) Below Reference Antenna*	
	High-Q	*Low-Q*	*High-Q*	*Low-Q*	*High-Q*	*Low-Q*
¼ λ No Coil	74		105		−3	
102” Coil At Top	43.6	43.5	12	25	−8.5	−8.6
102” Coil At 72”	41.5	41.4	14	30	−11.3	−11.4
102” Coil At 48”	40	40	17	32	−14.2	−14.3
102” Coil At 24”	38.6	38.5	20	34	−19.3	−19.3
102” Coil At Base	38.3	38.2	25	36	−24.5	−24.5
102” Coil At Base Shielded	38	38	25	38	−32.6	−32.6

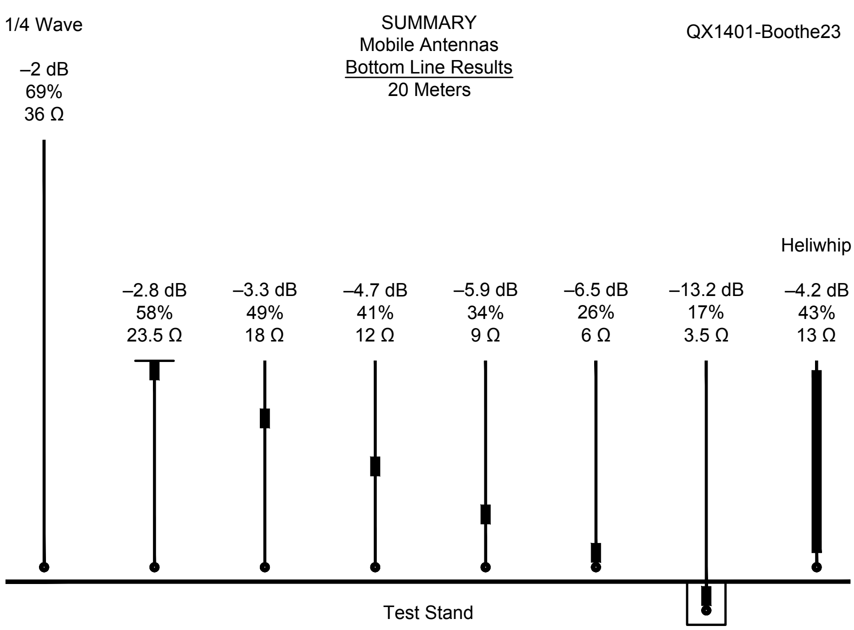

Figure 23 — This drawing summarizes the Series 1 test results for the various antenna configurations on 20 m.

Series 1 Conclusions

1) All other factors being the same, the coil loaded monopole with the coil closest to the top or end of the element will produce the greatest radiated signal. The lowest field strength by far will be seen from the one with a shielded coil at the base of the mast or whip. For a mobile antenna on 3.8 MHz, the difference is 24 dB! That's like going from 100 W down to 0.4 W! On 14.2 MHz, it's not so bad, like going from 100 W down to 10 W. No correlation was ever seen with the "optimum" positioning of the coil near the center of the mast.

Also, from the results shown, it's obvious that in the case of a *base loaded* antenna, a significant portion of the radiated field comes from *the coil itself*. Moving the coil into a shielded box reduces the field strength 6 dB on 14.2 MHz and 8 dB on 3.8 MHz!

2) For coil loaded monopole verticals, there's almost no measurable difference in field strength between high-*Q*, big wire, air wound coils, and low-*Q*, close-wound-on-a-form coils, no matter where in the mast they are located. As it turns out, this remains true whether the antenna is mounted over a poor ground plane like a vehicle or over a good ground plane like an extensive radial system. There is more about this in "High-*Q* and Low-*Q* Resonators Over Truck Versus Radial System" later in the article.

As mentioned earlier, other antenna variations were "thrown in" during Series 1 and Series 2 measurements. They included loaded monopoles with the lowest *Q* coils we tried, like the commercial "heliwhips" for 3.8 and 7.2 MHz. Results boiled down to the same generalities as stated above and below. Their field strength performance was low and related to the short length of "mast" below the start of the "lumped" inductance. Their bandwidth was high because of the two factors in point 3, below.

Personally, I think that big, air wound monster coils look like "Real Radio," but the data we collected show that they offer no advantage in radiated field strength. They might intimidate your competition, though.

3) Two things result in the greatest increase in bandwidth; Coils with higher length-to-diameter ratios and resonators with higher capacitance-to-inductance ratios. So, if you want more bandwidth, use long skinny close wound coils and use a design with as much capacitance (whip or hat) above or beyond the coil as possible. You won't be louder, but you'll be able to use a bigger part of the band without retuning. Also, things won't get "out of kilter" so easily when it rains or snows or, in the case of a mobile setup, you get close to trees or smack a bug with the coil.

Figure 23 shows the series of antenna configurations that we measured on 20 m. Figure 24 shows the configurations measured on 80 m.

Series 2

How the Length of the Base Mast Affects Efficiency

A resonator, consisting of a coil and an adjustable top whip was mounted on an 8 foot base mast on the test stand. After taking all the measurements, the mast length was reduced to 6 feet, then to 4 feet, then to 2 feet, and finally eliminated altogether. In effect, the last of these configurations resulted in a very short base loaded antenna. Figure 25 illustrates the various antenna configurations that we tested.

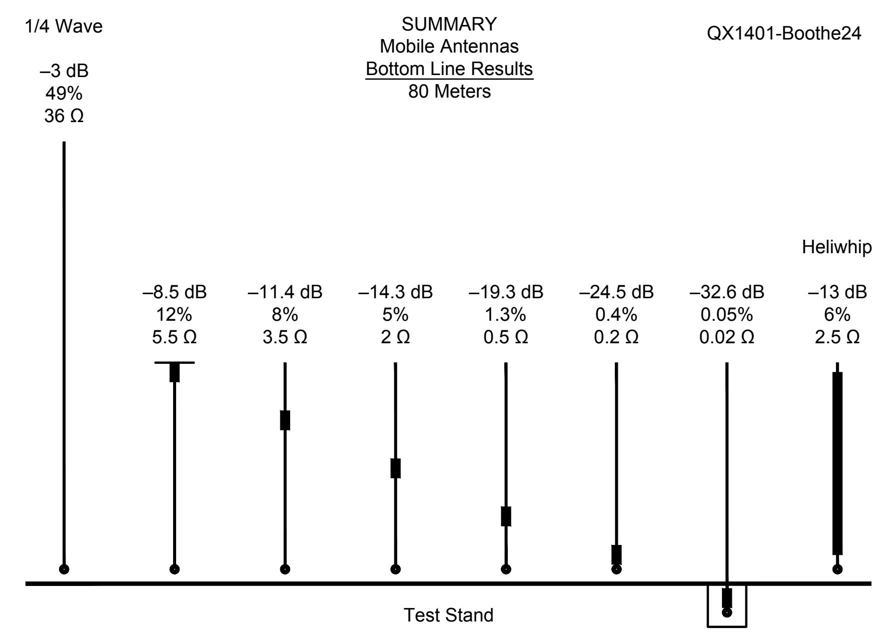

Figure 24 — Here are the Series 1 test results for the various 80 m antenna configurations.

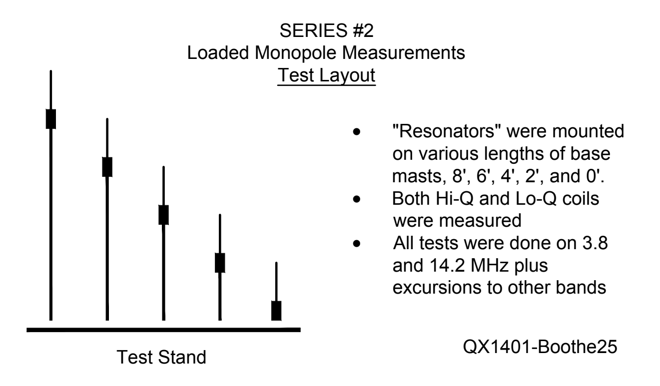

Figure 25 — This drawing illustrates the Series 2 test antenna configurations.

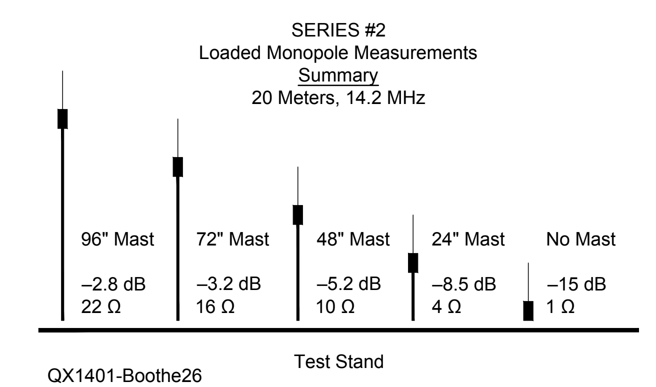

Figure 26 — Here is a summary of the Series 2 test results for 20 m.

Table 3
Series 2 Bottom Line Results
20 Meters, 14.2 MHz

Antenna Configuration	*Resistance at Resonance (Ω)*		*2:1 Bandwidth (kHz)*		*Field Strength (dB) Below Reference Antenna*	
	High-Q	*Low-Q*	*High-Q*	*Low-Q*	*High-Q*	*Low-Q*
¼ λ No Coil	53		690		–2	
96” Base Mast	40	40	375	456	–2.8	–2.8
72” Base Mast	34	34	215	342	–3.2	–3.2
48” Base Mast	29	28	120	195	–5.2	–5.2
24” Base Mast Unmatched	22	21	----	-----	–8.7	–8.8
24” Base Mast Matched	50	50	101	188	–8.3	–8.3
0” Base Mast Unmatched	19	18	----	----	–15.7	–15.8
0” Base Mast Matched	50	50	72	94	–14.8	–14.8

Table 4
Series 2 Bottom Line Results
80 Meters, 3.8 MHz

Antenna Configuration	*Resistance at Resonance (Ω)*		*2:1 Bandwidth (kHz)*		*Field Strength (dB) Below Reference Antenna*	
	High-Q	*Low-Q*	*High-Q*	*Low-Q*	*High-Q*	*Low-Q*
¼ λ No Coil	74		105		–3	
96” Base Mast	44	43.5	19	38	–8.8	–8.9
72” Base Mast	42	41.5	18	38	–11.4	–11.5
48” Base Mast	40	40	15	35	–15.2	–15.3
24” Base Mast	38.3	38.2	12	31	–22.2	–22.4
0” Base Mast	38	38	8	19	–28.6	–28.8

Table 5
Effective Ground Resistance (Ω)

Band	*Frequency*	*Big Vehicle (Truck Stand)*	*Small Vehicle (1993 Ford Escort)*	*On-Ground Radial System*
10 m	28.5 MHz	5	6	4
15 m	21.3 MHz	10	11	5
20 m	14.4 MHz	19	23	6
30 m	10.1 MHz	25	31	8
40 m	7.2 MHz	31	37	11
80 m	3.8 MHz	40	47	17
160 m	1.8 MHz	84	91	24

Of course, the resonator was readjusted for resonance as the base mast length was changed. As in series 1, all tests were done with high-*Q*, air wound, spaced, “square” coils as well as low-*Q*, close wound on long skinny form types on both 14.2 and 3.8 MHz, with occasional excursions to the other bands. Table 3 shows our results on 20 m and Table 4 shows the results on 80 m. Figure 26 shows a summary of our 20 m tests and Firuge 27 summarizes the results for 80 m.

Series 2 Conclusions

1) The length of the mast below the lumped inductance has the greatest effect on the field intensity of a coil loaded, “short” monopole, all other factors being the same. Combining the Series 1 and 2 numbers, I draw this conclusion: “In the case of shortened, loaded antennas, all other factors being the same, the one with the longest mast between the feed point and the start of the lumped inductance will win the field strength contest.”

For example, on 3.8 MHz, adding 2 feet to the base mast of a mobile antenna is like doubling your power. On 14.2 MHz, adding four feet to your mast is like doubling your power.

2) There is an almost *immeasurable difference* in field strength between low-Q and high-Q coils used to load shortened monopoles, no matter the length of mast below the coil. Note that in all cases, as the mast length is shortened, the bandwidth is reduced as well as the efficiency.

The rest of this report will present actual measured performance comparisons dealing with the following subjects:

• Ground resistance of large and small vehicles and a “typical” on-ground radial system.

• High-*Q* and low-*Q* coil loaded monopoles over a vehicle versus an on-ground radial system.

• Various mounting angles of resonator to mast on loaded antennas.

• Multiple resonators on single monopole masts.

• Use of “mag mounts” on mobile antenna installations.

• Capacity hat locations on loaded monopoles.

• Coil top loading versus capacity hat only top loading on shortened antennas.

• Various matching and tuning schemes for shortened, loaded antennas.

• Current in loading coils for shortened, loaded antennas.

• Alternate types of loading coils.

Ground Resistance of Large and Small Vehicles Versus a Radial System

Much has been said about this subject,

but little in the way of real numbers has been presented. These measurements were made at the Harlingen, Texas test site. We used helium filled balloons to support ¼ λ antennas fed against each subject ground plane. Although one would expect the actual numbers to be different for every vehicle, location, and climatological condition, the comparisons are interesting. See Part One for a description of the "Truck Stand" and the radial system. Table 5 shows our measurements across the HF bands for our three ground systems.

Conclusions:

1) The size of the vehicle has most to do with its ground resistance on any particular frequency and location. The smaller vehicle will have higher resistance and lower efficiency. Stamp collecting might be more rewarding than going mobile with a small motorcycle on 160 or 80 m, unless you can drag a counterpoise wire.

2) Ground resistance of a less than perfectly conducting plane is inversely proportional to the frequency of operation. So, if the vehicle is small, expect comparatively poor results on the lowest frequency bands. If you want really top results mobiling on 1.8 MHz, consider making your next vehicle one that can pull a flatbed, lowboy semi trailer, perhaps with a copper plated floor. Mount the antenna in the middle of the trailer. You still won't be king of the band, but you may be king of the road.

Even though the numbers indicate that a mobile antenna for 1.8 or 3.8 MHz may be in the 1% to 3% efficiency range, lots of great contacts, including DX, are made by people using that mode. In fact, my first DX contact from our new home was made from the mobile rig in the truck stand sitting in our driveway. The antenna was a 160 m resonator with a long 1 inch diameter close-wound coil of #20 enameled wire mounted on an eight foot mast. I called CQ on 1.824 MHz around sunrise, and was answered by Bob Briggs, VK3ZL. I should add that Bob has good ears.

High-Q and Low-Q Resonators Over Truck Versus Radial System

Some claim that the almost identical performance of high-*Q* and low-*Q* resonator coils is because of their use with poor ground resistance ground planes, like vehicles. This theory has been put forth in Internet discussions of our findings. These tests were done in Harlingen, Texas using a 6 foot mast below the resonators. They were repeated a number of times with the same results. The truck stand and the radial system are described in Part One. Table 6 compares our measurements using the truck stand with measurements made over an extensive on-ground radial system. That radial system, described in Part 1 of the article, consisted of 60 copper radials, with lengths from 40 to 60 feet, stretched out on the ground under the test antenna.

Conclusions:

1) The lower ground resistance of an average on-ground radial system compared to that of a big vehicle will noticeably improve field intensity of a coil loaded monopole. This is certainly no surprise.

2) The relationship between high-Q and low-Q loading coils remains the same — that is there is no significant difference in performance between the two, whether used on antennas with high or low ground resistance.

Angle of Resonator to Mast

The question here was what effect changing the angle between the resonator and mast would have on performance. These tests were related mostly to coil loaded mobile antennas, but would apply to any shortened, loaded monopole. The tests were performed during both our Fletcher, North Carolina and Harlingen, Texas measurements. We used a 6 foot mast, with high-*Q* and low-*Q* coils and a top whip. See Table 7.

Conclusions:

1) The mounting angle of resonators to mast on inductively top loaded antennas has little to no effect on field strength, unless the angle is more than 90° from the mast.

2) Mounting resonators at different angles to either accommodate multiple resonators and/or to reduce vulnerability to damage will have no detrimental effect on signal strength.

3) Changing the angle of resonator to mast *will* affect the resonance, so retuning is usually in order.

Even when the resonator begins to parallel the mast, it does not result in a large cancellation of fields. On the other hand, if the top loading wires of *non-inductively loaded* verticals or inverted L antennas droop significantly, the losses can become quite significant.

Although the figures are not presented here, during any measurement sequence involving capacitive only top loading,

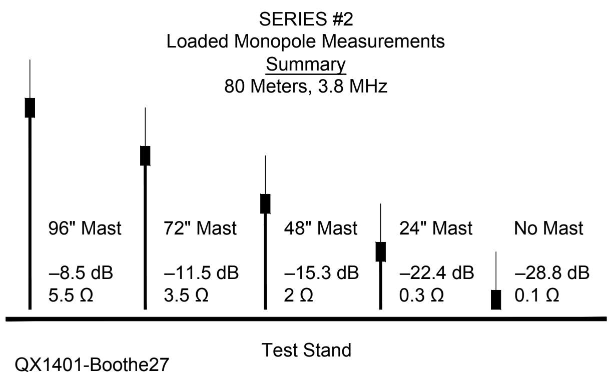

Figure 27 — This drawing summarizes the Series 2 test results for 80 m.

Table 6
Field Strength in dB Below the Reference Antenna

Band	*Frequency*	*Antenna Tested*	*Truck Stand*	*On-Ground Radial System*
20 m	14.2 MHz	¼ λ No Coil	–2 dB	–0.8 dB
20 m	14.2 MHz	With High-*Q* Coil	–3.2 dB	–1.5 dB
20 m	14.2 MHz	With Low-*Q* Coil	–3.2 dB	–1.5 dB
80 m	3.8 MHz	¼ λ No Coil	–3.1 dB	–1.2 dB
80 m	3.8 MHz	With High-*Q* Coil	–11.5 dB	–6.5 dB
80 m	3.8 MHz	With Low-*Q* Coil	–11.6 dB	–6.6 dB

Table 7
Field Strength in dB Below Reference Antenna for Different Resonator to Mast Angles

Antenna	*Vertical 0°*	*45°*	*Horizontal 90°*	*135°*
	Field Strength (dB) Below Reference Antenna			
14.2 MHz Low-*Q*	–3.3 dB	–3.3 dB	–3.3 dB	–3.5 dB
14.2 MHz High-*Q*	–3.3 dB	–3.3 dB	–3.3 dB	–3.5 dB
3.8 MHz Low-*Q*	–11.4 dB	–11.4 dB	–11.4 dB	–11.7 dB
3.8 MHz High-*Q*	–11.3 dB	–11.3 dB	–11.3 dB	–11.6 dB

significantly lower field strengths were observed as the big hat wires were allowed to droop down. The angle to the vertical element also greatly affected the tuning. This subject needs to be the basis of some future studies.

Multiple Resonators on a Single Mast

These tests were aimed at multi-band setups. They were done at Fletcher and in Harlingen on the test stand and the truck stand. A 6 foot mast was used below the resonator(s). The idea was to compare the signal strength performance to single resonator setups. As resonators were added to the mast, tuning was performed to readjust for resonance. As in the other Tables, Table 8 uses a perfect ¼ λ ground-plane antenna as the reference. The numbers for 7.2, 10.1, 18.15, and 21.3 MHz are based on only three test runs, but the pattern was the important point. Other mast lengths were tried with similar results as these. Resonators were mounted 90° from the mast. First, each resonator was measured alone. Then, resonators were added one at a time, retuned for resonance, and field intensity was measured. Results were the same for high-*Q* and low-*Q* resonators. Figure 28 shows how resonators were added, and also shows a two-tiered arrangement.

Conclusions:

1) Adding resonators to a mast for the purpose of operating on multiple bands/frequencies does not degrade the signal strength performance compared to a single resonator setup.

2) As resonators are added, retuning will be required.

Using Magnetic Mounts for Mobile Antennas

Putting a mobile antenna on a "mag mount" without low impedance grounding straps to the vehicle is the same as putting a capacitor in series with one half of that antenna. Depending on the size and number of magnets, plus the frequency of operation, this results in some amount of reactance. The reactance must be cancelled, or "tuned out."

Table 8
Field Intensity Readings for One to Six Resonators on a Mast Versus a ¼ λ Reference Antenna

Frequency	*One*	*Two*	*Three*	*Four*	*Five*	*Six*
3.8 MHz	–11.5 dB	–11.4 dB	–11.4 dB	–11.5 dB	–11.6 dB	–11.5 dB
7.2 MHz	–8.4 dB		–8.3 dB	–8.4 dB	–8.4 dB	–8.4 dB
10.1 MHz	–5.9 dB				–5.8 dB	–5.8 dB
14.2 MHz	–3.3 dB	–3.2 dB	–3.3 dB	–3.3 dB	–3.3 dB	–3.2 dB
18.15 MHz	–1.3 dB					–1.3 dB
21.3 MHz	–0.7 dB			–0.7 dB	–0.7 dB	–0.7 dB

Table 9
Magnetic Mount Characteristics

Mag Mount Type	*Surface Area (In^2)*	*Capacitance To Ground (pF)*
3 Each 3" Diameter Magnets	21	323
4 Each 3" Diameter Magnets	28	431
3 Each 4" Diameter Magnets	38	584
4 Each 4" Diameter Magnets	50	769
4 Each 5" Diameter Magnets	78	1200

Table 10
Mag Mount Reactance by Type and Band

Frequency	*3 Each 3" Diameter Magnets*	*4 Each 5" Diameter Magnets*
	Reactance (Ω)	*Reactance (Ω)*
28 MHz	17	5
21 MHz	25	7
14 MHz	35	10
7 MHz	70	20
3.8 MHz	140	40
1.8 MHz	280	80

(A) (B)

Figure 28 — Part A shows a multi-resonator setup and Part B shows a setup with resonators at two levels.

(A)

(B)

(C)

(D)

Figure 29 — Photo A shows a mag mount on a tool box plate in a pick-up truck bed. Photo B shows a mount on the roof of a car. The mount in Photo C is on a car trunk lid, with wide ground braids attached to the car body. Photo D shows the mag mount on another car roof.

We wanted to compare various designs of mag mounts, and to look at the performance compared to standard body mounts to see if there was a difference in field strength. All this information was derived from measurements made in Harlingen using a variety of mag mounts on various vehicles. The capacitance of any particular mag mount may vary from those we measured if a different thickness of protective covering is used on the bottom of the magnets. We used a Ballentine Labs Model 520 capacitance meter. On the field strength chart, figures for 3.8 MHz include both "matched" and "unmatched" numbers because at resonance, the SWR was more than 2:1 when using mag mounts. Figure 29 shows the various vehicles and mag mount styles tested. Table 9 gives the physical details of the various mag mounts we tested. Table 10 lists the reactance by band for two of the mag mounts, and Table 11 shows the field strength measurements. Figure 30 is a simple illustration of the problem with mag mounts.

Conclusions:

1) The use of a mag mount for a mobile antenna will result in a significant reduction of field strength. The loss will be worse for smaller mag mounts and for lower frequencies. Use of the smaller type on 14 MHz cuts the power radiated in half from that of a body mount. On 3.8 MHz, use of even the larger type results in a similar loss when matched.

2) The reactance added to a mobile antenna system by a mag mount is inversely proportional to the total surface area of the magnets. In other words, to least affect the original antenna design, use the mag mount with the most magnets of the greatest diameter available.

Better yet, if it's possible, add a *low impedance* connection to the vehicle skin. The difference, depending on mag mount and frequency, can be like multiplying your power by four, or even up to ten.

Capacity Hat Location

Many articles have stressed the importance of mounting capacity hats well above loading coils to avoid losses. Our object here was to quantify the difference in performance between hats adjacent to the top of the coil versus well above the coil. See Figure 31. These tests were done in Harlingen, Texas. Antennas for 1.8, 3.8, and 14.2 MHz were tested over both the truck stand as well as the ground radial system. See Table 12.

Conclusions:

1) Conventional wisdom is correct, but, once quantified it's not a very big deal. On 1.8 and 3.8 MHz you can get a couple tenths of a dB by moving the hat up away from the coil. You have to decide whether it's worth the work and risk for that kind of payback.

2) We also compared coils with and without metal end caps and found no difference in field strength performance, but a pronounced effect on tuning. This was especially true at lower frequencies, depending on coil size and proximity of windings to cap.

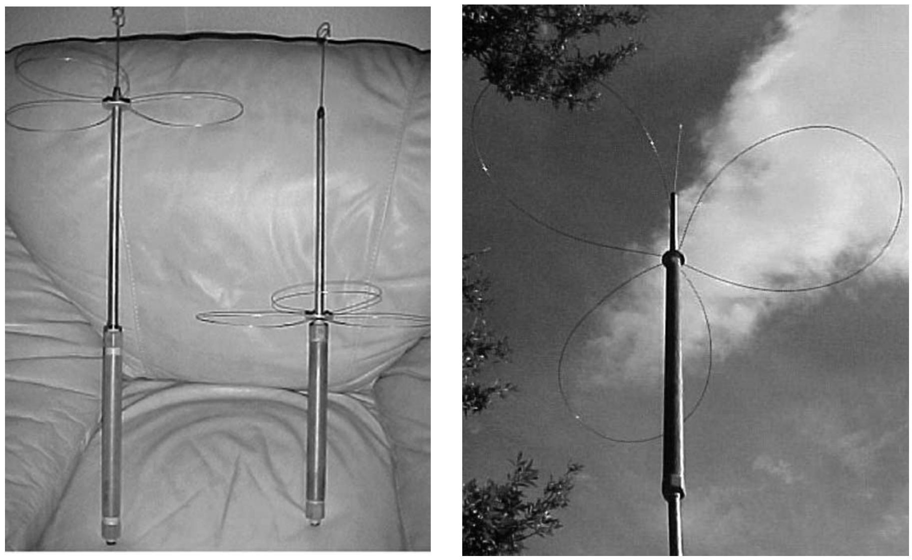

Figure 31 —Photo A shows a 3.8 MHz resonator with high and low capacity hats. Photo B shows a 1.8 MHz resonator with a low capacity hat.

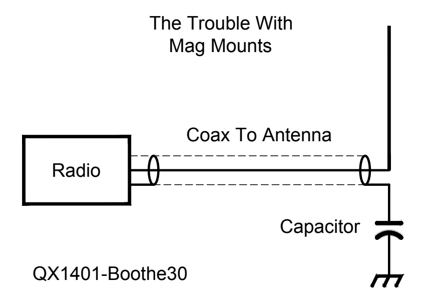

Figure 30 —This drawing illustrates the problem with using mag mounts. You are placing an unknown capacitor between the bottom of the antenna and the vehicle body/ ground plane.

Table11
Field Strength by Mount Type

Frequency	*Mount Type*	*Field Strength (dB)*	
		Unmatched	*Matched*
14.2 MHz	Direct Car Body	–3.2 dB	------
14.2 MHz	3×3" Mag Mount	–6.2 dB	------
14.2 MHz	4×5" Mag Mount	–5.2 dB	------
3.8 MHz	Direct Car Body	–11.3 dB	------
3.8 MHz	3×3" Mag Mount	–21.3 dB	–18.7 dB
3.8 MHz	4×5" Mag Mount	–15.3 dB	–14.7 dB

Table 12
Field Strength Compared to Reference Antenna

Frequency (MHz)	*Low Hat Truck Stand*	*High Hat Truck Stand*	*Low Hat Radials*	*High Hat Radials*
14.2	–3.3 dB	–3.3 dB	–1.4 dB	–1.4 dB
3.8	–11.6 dB	–11.4 dB	–6.6 dB	–6.4 dB
1.8	–19.4 dB	–19.1 dB	–10.5 dB	–10.2 dB

Coil Top Loading Versus Capacity Hat Only Loading

Many articles have indicated that capacity hats or wires should be used for top loading shortened monopoles rather than coils, for the sake of efficiency. We wanted to quantify the difference in performance. Sevick had offered valuable information on this subject in his work in 1973. We compared antennas over the radial system at the citrus grove test site in Harlingen, Texas. We used balanced capacity hats as opposed to "inverted L" configurations to avoid directional effects and any significant horizontal polarization. Table 13 shows our results. These antennas were erected on only three separate occasions, but the results were consistent.

Conclusions:

1) There is almost *no* signal strength advantage to using only top loading capacity hats or wires in lieu of top loading coils to resonate short monopoles, all other factors like vertical mast length being the same. This coincides with the fact that there is no significant difference in performance between high-Q and low-Q coils used for loading monopoles. Bandwidth was nearly identical on these examples.

2) During the tests on capacity-only loading it was noted that when the wires or hat, skirted or not, drooped down from the top of the mast, there was a significant drop in field strength. Although the numbers were recorded in our raw data, we have never matched the exact angle or number and size of wires to the particular field strength. We found that we had to keep the wires horizontal or higher in order to get top performance, which was a real task at the test site. We wanted to try this test on 1.8 MHz, but the logistics were beyond our practical capability at that location.

More work should be done in this area to better quantify the losses of drooping capacity hats. There are many "umbrella" and guy wire hat designs in articles and books that should be evaluated. An ultimate example, somewhat related to "umbrella" wire loading and linear loading is the "Meandered Line" antennas published in the IEEE Transactions, December, 1998. Its performance can be best likened to a large, unshielded dummy load, as experienced by Arch Doty, W7ACD when he built a big one for 160 m.

3) The various Inverted L designs may have an advantage over top loaded straight verticals (coil or capacitor) of the same size due to increased horizontally polarized radiation and bandwidth. This depends on the intended use and propagation variables, as well as the ratio of vertical to horizontal sizes and the angle of the top of the "L" to the vertical element.

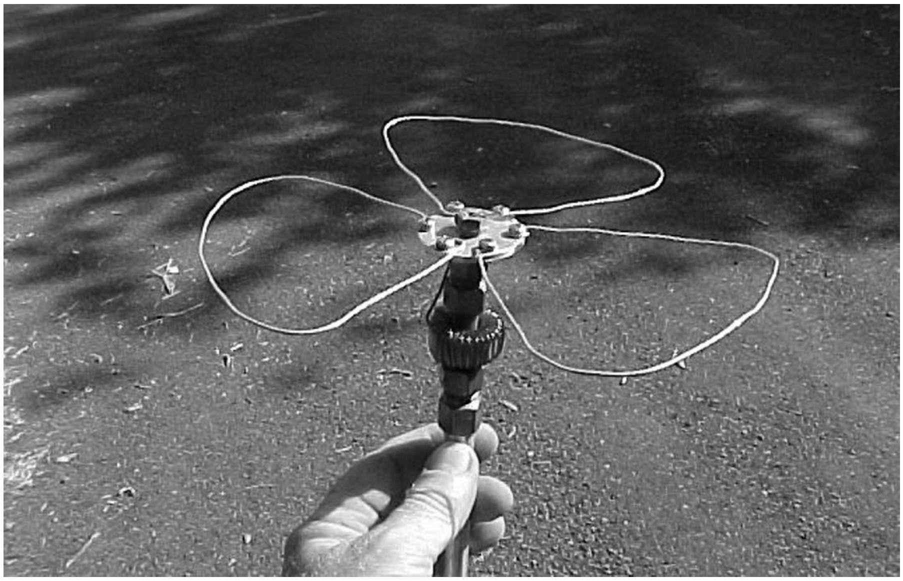

Figure 32 — You Can see the 14.2 MHz toroidal resonator below the loop wires.

Table 13
Field Intensity Compared to the ¼ λ Reference Antenna

Frequency (MHz)	*Mast Height (Ft)*	*Coil/Whip Resonator (dB)*	*Capacity Hat Only (dB)*
14.1	8	–1.1	–1.1
3.8	31	–3.1	–3.0

Table 14
Field Strength below Reference Antenna and Bandwidth for Less Than 2:1 SWR

Frequency	*Standard Coil*		*Toroid Coil*		*Pie-Wound Coil*	
	Field Strength	*Bandwidth*	*Field Strength*	*Bandwidth*	*Field Strength*	*Bandwidth*
14.2 MHz	–3.3 dB	478 kHz	–4.6 dB	590 kHz	–4.4 dB	490 kHz
3.8 MHz	–11.4 dB	30 kHz	–21.2 dB	122 kHz	–15.1 dB	52 kHz
1.8 MHz	–19.4 dB	5 kHz	N.A.	N.A.	–23.5 dB	27 kHz

Table 15
Matching at the Antenna Base versus Matching in the Vehicle Cabin
Field Strength in dB Below a Perfect Antenna

Antenna	*No Match*	*Matched at the Base*	*Matched in the Cabin*
3.8 MHz 6' Mast on Truck Stand	–11.5 dB	–11.1 dB	–12.5 dB
7.2 MHz 6' Mast on Ford Escort	–9.0 dB	–8.5 dB	–9.7 dB

Alternate Types of Loading Coils

The object here was to compare the performance of antennas with several types of loading coils. These tests were done in Fletcher, North Carolina as well as Harlingen, Texas. A lot more work needs to be done in this area. For instance, toroidal cores of the right "mix" and size must be found, especially for common power levels on the lower Amateur Radio bands. See Figure 32. The one used for the 3.8 MHz test overheated at 10 W. Nothing could be found for the 1.8 MHz toroid test. Also, a method for spacing the turns on pie-wound coils had to be developed. One way would involve printed circuit technology. That solution is an economic show stopper for the quantities needed for the Amateur Radio market. The turn-to-turn capacitance, especially on the lower frequency units caused significant losses. The pie-wound coils in these tests were our earliest prototypes.

These tests were run using a 72 inch base mast on the test stand and the truck stand. Table 14 summarizes our results.

Conclusions:

1) These alternatives show great promise if materials and processes can be further developed. They are particularly attractive considering their small size, weight and wind resistance combined with exceptional bandwidth.

WB9NUL and I ran the 14 MHz pie-wound resonator, shown in Figure 33, on a cross-country trip to the west coast. It was on an 8 foot mast. It was interesting that we didn't need the fishing line guy string that we normally used on a long-mast mobile antenna. At 50 MPH or faster, the antenna was frozen at about 20° back from vertical. Apparently at that angle the drag was equaled to the lift. The antenna had a nearly flat SWR across the whole 20 m band.

As an aside, I should add that we were so impressed with the possibilities of the pie-wound design, that we went to Washington D.C. and did a patent search. Once into the sub-sub-sub category of our interest, we had 15,000 patents to review! It took 3 days to go through them, and we found less than ten that were even vaguely related. Most were recent and held by large armed forces contractors. The earliest, and probably closest to our stated design purpose, was filed in 1925 by J. O. Mauborgne and Guy Hill. See Figure 34. We came away much enlightened but convinced that there was no need to pursue a patent. We learned a lot from the experience.

Figure 35 shows the various antenna arrangements we tested with alternative mobile antenna designs, along with a summary of our test results.

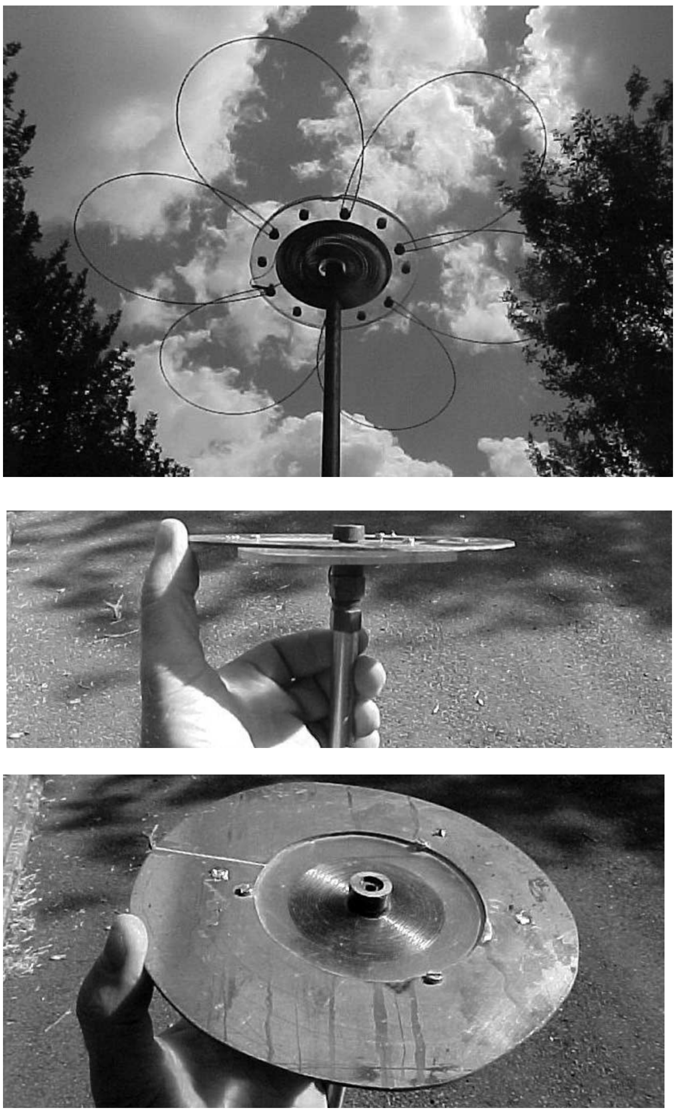

Figure 33 — Photo A shows a 3.8 MHz pie-wound resonator on a ½ inch mast. Photo B shows a side view of the 14.2 MHz pie-wound coil. Photo C shows a top view of the 14.2 MHz pie-wound coil.

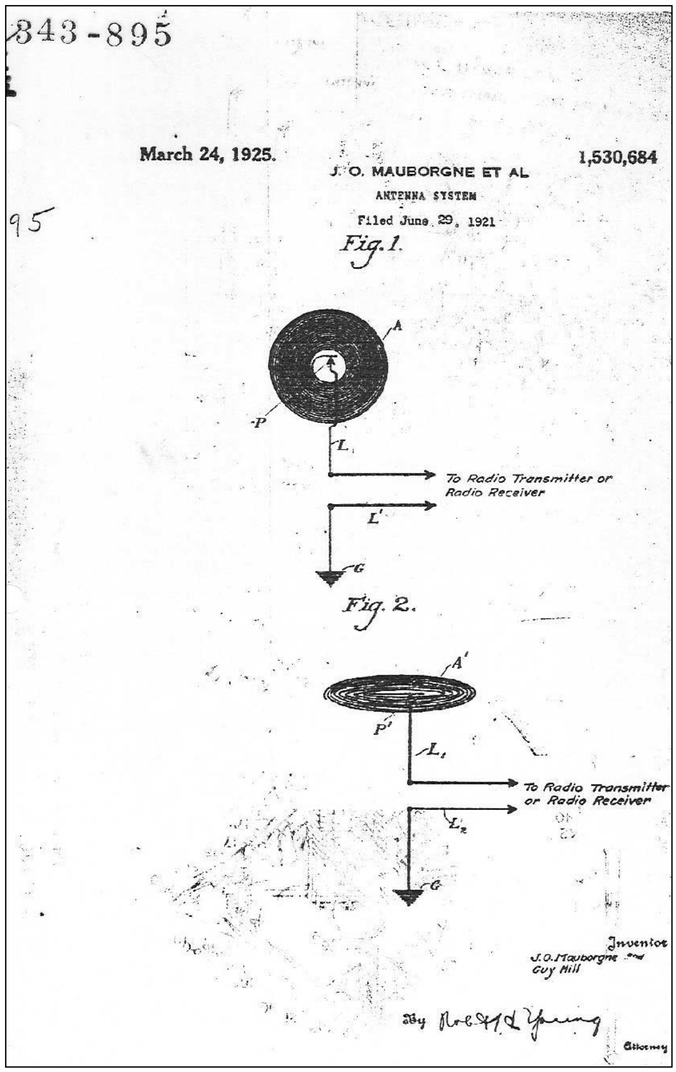

Figure 34 — This page is from a 1925 pie-wound antenna patent.

Matching and Tuning Schemes

When a mobile antenna under test in Series 1 and Series 2 had an SWR of 2:1or greater at resonance, readings were taken with both matched and unmatched conditions. The matching was done at the feed point of the antenna. See Figures 36 through 40 for various examples of matching arrangements.

Comparisons were made between matching at the antenna base versus in the vehicle cabin during the Harlingen, Texas tests. This can be likened to a tuner at the base of a short vertical in the backyard versus a tuner in the shack instead. This was an effort to simulate the use of autotuners and others at the transmitter end of the coax feed line. In order to get some examples, we used a 3.8 MHz antenna with a 6 foot mast on the truck stand, and a 7.2 MHz antenna on a Ford Escort at the Citrus Grove test site. Both antennas had under 2:1 SWR, but high enough in SWR that in both cases small solid state rigs would reduce their power levels when transmitting on them. For these measurements, the feed point matching device was either a shunt coil or shunt capacitor to ground. Of course, the antenna was retuned to resonance. The in-cabin matching device was a small commercial "mobile tuner" or a home brewed "T" or "L" network. As in all measurements to this point in this report, a precise 10 W was sent to the antenna system being tested. Table 15 summarizes our measurements.

Conclusions:

1) Matching at the base of a loaded monopole to achieve 1:1 SWR will usually result in some degree of improved field strength. The amount of improvement will depend on how far from 50 Ω you start with, and the frequency.

2) Matching a mismatched antenna with a tuner in the cabin or the shack, like an autotuner or "mobile" tuner will result in some small amount of loss of signal strength, *assuming the same power is delivered to the system.* This is likely due to losses in the tuner itself rather than in the short piece of coax used in a mobile installation. Of course, several other factors come into play here. This sort of setup is often employed so that the modern miniaturized solid state transceiver is "happy" and will deliver full power to the antenna but power is lost due to the efficiency of the tuner. The SWR on the coax will not be improved by the cabin or shack tuner, and so the concern becomes one of noise reception and energy radiated by the mismatched coax. In a base station, with perhaps 100 feet of coax, losses could be severe, especially on the higher frequency bands.

Also, at Harlingen, measurements were taken to quantify the loss when an antenna was tuned to the high end of the band and was being used on the low end of the band with a tuner in the cabin. This situation is common with operators using top loading resonators who want to quickly switch from phone to CW "on the run," as county hunters often do. The matching devices were the same as above. Table 16 summarizes these measurements

3) Using a cabin tuner to match a mobile antenna to a frequency far from its resonance will result in a significant reduction of signal strength. It *will* allow the transmitter to work into a matched load and that is certainly better than using no matching or retuning, but it is not the desirable way to operate on a long term basis.

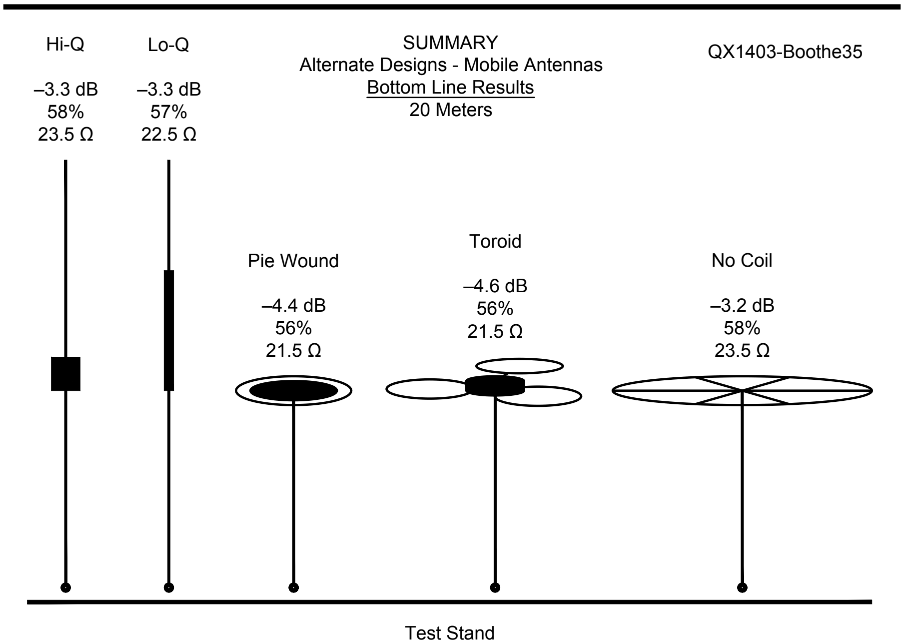

Figure 35 — This drawing illustrates the various alternative mobile antenna designs that we tested. The field strength, feed point impedance and efficiency of each antenna type is also shown.

Table 16
Antenna Tuned to Phone Band but Used on CW, With a Cabin Tuner
Field Strength in dB Below Perfect Antenna

Antenna Resonant on 80 m, 3815 kHz		
Measured at 3815 kHz (Resonant)	*Measured at 3525 kHz (CW)*	*Measured at 3525 kHz (CW)*
No Matching	*No Matching*	*Matched in Cabin*
–11.5 dB	*–28.7 dB*	*–27.7 dB*
Antenna Resonant on 40 m, 7240 kHz		
Measured at 7240 kHz (Resonant)	*Measured at 7040 kHz (CW)*	*Measured at 7040 kHz (CW)*
No Matching	*No Matching*	*Matched in Cabin*
–9.0 dB	*–15.5 dB*	*–14.5 dB*

Figure 36 — A small commercial "screwdriver" antenna.

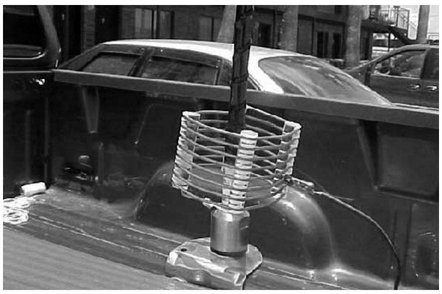

Figure 37 — Shunt matching coil at the base of an antenna.

Figure 38 — A large commercial "screwdriver" motorized antenna.

One of the ways operators get around this problem today is through the use of remotely tuned antennas, like the various "screwdriver" designs. To achieve the ever sacred 1:1 SWR without leaving the drivers seat, however, most designs sacrifice efficiency due to the short mast below the lumped inductance and the very lossy mounting structures many employ. As I said in the introduction to this report, "everything works," it's just a matter of what compromises we wish to make to satisfy our own priorities.

Current in Loading Coils

Our early efforts to determine whether the RF current dropped or remained the same from the bottom to the top of loading coils in monopoles were not too conclusive or very scientific. For instance, we applied excessive power to the antennas, shut down and quickly checked the temperature along the coils. They were warmer at the bottom. But, that certainly didn't satisfy us as a proof. We moved neon and fluorescent bulbs along the coils to indicate relative voltage while transmitting a carrier. Much higher voltage was indicated at the top of the coil and our logic told us that if the voltage went up, the current had to go down. But, that didn't prove anything either.

Our initial metered measurement of RF current in monopole loading coils was done in the yard at our home in Harlingen. See Figure 41. Various configurations of short loaded antennas were built and tested over an extensive radial system. We collected data for base, center and near top loaded antennas for 10.1 MHz and 7.2 MHz. We used both

Figure 39 — Note the parallel beam mounting structure.

Table 17
Current at Top of Coil With 100 mA of RF at the Bottom of the Coil

Antenna	*Base Loaded (mA)*	*Center Loaded (mA)*	*¾ Top Loaded (mA)*	*Very Top Loaded (mA)*
7.2 MHz 92” High Q Coil	66	45	37	NA
7.2 MHz 92” Low Q Coil	64	43	35	NA
10.1 MHz 92” High Q Coil	75	60	52	NA
10.1 MHz 92” Low Q Coil	74	60	50	NA
3.8 MHz 72” Mast and Resonator			79	NA
1.8 MHz 96” Mast and Resonator			65	NA
14.2 MHz 116” Toroid Coil			79	47

high-Q and low-Q coils. Eventually, we measured RF currents in many different loading coils on 1.8 and 3.8 MHz at the citrus grove site on both the truck stand and the big radial system. Table 17 is a sampling of current readings when the current at the base of the coil was 100 mA (RF). Figure 42 shows the RF ammeters installed at the top and bottom of a loading coil.

The test procedure and the reasons for the measurements are discussed in Part One.

Conclusions:

1) The current tapers from the bottom to the top of loading coils used to resonate shorter than quarter wave length monopoles. The Q of the coil has little to no effect on the drop.

The amount of taper *seems* related to that portion of the quarter wave that has been replaced by the coil, but that is an over-simplification. The reason the current tapers, other than a small amount of conductor resistance and radiation, is that in a standing wave antenna like a monopole over a ground plane, the net current at any point is the “vector” sum of currents at that point. At any point along the monopole, or a series inductor, there is a phase difference between the current coming from the source and the current reflected back from the open end or top/end of the monopole. The resultant net current is less as you move toward the open end of the monopole, where it is virtually zero, because at that end point, the forward and reflected currents are equal in magnitude and opposite in phase thus superposing to zero.

This information may answer the questions we had about the lack of impact of coil Q on field strength and the inability to confirm the published formulas to “optimally” locate coils in the mast. It may also explain why capacity only loading is no better than top coil loading, all else remaining the same.

Concluding Remarks

Some of the books, articles and modeling programs appear to have it wrong! Designers and builders of short, loaded antenna elements have often used this information, causing misguided decisions.

It would be prudent to question any design stemming from the assumption that the current in monopole loading coils is uniform. Furthermore, any modeling program that considers series loading coils in standing wave antennas to be a single point in the circuit are likely in error, and will lead the designer/evaluator astray. Similarly, statements about the effect of losses in loading coils, especially “low Q” coils, seem to be grossly exaggerated.

Our objective was to compare the effectiveness of different designs of shortened, loaded antenna elements. In the process, we came to some eye-opening conclusions. More work of this type should be done in order to help builders and buyers make good decisions.

I would like to reinforce a few things and offer some sources of important information.

Figure 40 —The coil used to resonate the antenna on 20 m.

Figure 41 — One of the coil current measurement setups.

Figure 42 — RF ammeters reading 100 mA on the bottom and 42 mA on the top of the loading coil.

First of all, as seen in the measurements presented in this article, the effectiveness of these kinds of antennas depends in part on the counterpoise against which they are working. We must remember that the loaded monopole is only half of the antenna and that there must be a second half so that an electromagnetic field is established between the two parts. That field is the source of radiated energy.

Certainly, mounting the loaded monopole in the center of a large conductive plate will provide the kind of radiating field you need, but unless you have a metal roofed building or such, you'll likely have to simulate that plate some other way.

There is plenty of information in Amateur Radio and broadcast literature about ground radial systems. Material has been published in the last decade on this subject by Robert Sommer, N4UU, Rudy Severns, N6LF, and Arch Doty, W7ACD.[20, 21, 22, 23, 24] I would suggest those works for your perusal. For some earlier classics on the subject, look up the articles by R. C. Hill, G3HRH, as well as G. H. Brown, and G. H. Brown, R. F. Lewis, and J. Epstein.[25, 26, 27]

Many people contributed to this project. Joyce Boothe, WB9NUL, my wife and best friend, has worked with me on all my endeavors for more than 30 years. It could not have been done without her. I particularly want to thank Arch Doty, W7ACD, who has been instrumental to the tasks at hand for a similar period of time. Other contributors of note include Cecil Moore, W5DXP, Mike Carver, KG5UZ, Cheryl Carver, KJ5PQ, Walter Schulz, K3OQF, George Ostrowski, K9PAW, Greg Chartrand, W7MY, Terry Dummler, WQ7A, and Barry Mitchell, NØKV. Of course, our old friends John Frey, W3ESU and Harry Mills, K4HU, both Silent Keys now, did a lot to help us in Fletcher, along with so many of their friends from the Hendersonville, North Carolina area plus a few locals in the Lower Rio Grande Valley. All of these friends made our quest for the answers possible.

Barry Boothe, W9UCW is an ARRL member and holds an Extra Class license. He has held his call since 1954 after holding WN9UCW for a couple months. He became interested in Amateur Radio at age 13, after experimenting with electricity and electronics during his junior high school years.

Barry was with Caterpillar for 31 years at facilities in the US and Brazil. He was a division manager when he took early retirement. He taught electricity and electronics classes at a community college for six years.

His primary ham radio interests have always been building, antenna research and low-band DXing. He has made 20 trips to Central and South American countries, always involving Amateur Radio to major degree. Barry won two cover plaque awards for QST *articles published in the 1970s. Another of his interests is woodworking.*

Barry and his wife Joyce, WB9NUL have lived in the Lower Rio Grande Valley for over 23 years. Joyce has held her call for 40 years. She is a county hunter and was president of MARAC, the mobile awards club for 7 years.

Notes

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Radiation Resistance, Feed Point Impedance and Mythology

An understanding of these topics is vital to antenna experimenters, yet many continue to misunderstand the definitions.

It is not surprising that there remains a great deal of confusion about antenna radiation resistance generally, and its relationship to feed point impedance in particular. Many authors state special cases as being general rules. To top it off, some actually state that there are multiple definitions of radiation resistance. We are all free to invent whatever definitions we wish about anything, but such made-up definitions do not relate to the larger body of knowledge and only serve to further confuse an already confusing issue. Many technical publications, including Amateur Radio publications, state false definitions of radiation resistance and erroneous values for a variety of examples.

Of all the world-class authors on the subject, John Kraus provides the most comprehensive discussions in a single text that I have found.[1] He defines radiation resistance from several different approaches and applications. The equations he published for the definition take several different forms in that they use differing variables. Some readers have mistakenly interpreted these multiple definitions as "different" definitions. In this paper I will show that the special case definition for radiation resistance in vertical antennas (of special interest to Amateur Radio operators) can be derived from the general case definition. Kraus does not provide these derivations, but the book is written as a teaching text. As such, one can imagine such derivations were assigned as graduate-level homework assignments, and perhaps even Masters' thesis topics.

[1]Notes appear on page 35.

By working through this derivation we can develop a deeper understanding of this very difficult yet critically important antenna parameter. Further, we can converge on an unambiguous definition of radiation resistance, using both a general equation and a verbal description.

For purposes of simplicity this paper will focus upon linear in-line antennas, i.e. single elements using a straight conductor that has a diameter very small compared to the wavelength. Multi-element arrays, "bent" arrays, planar structures and 3-d antennas can be very complex. NEC-based modeling tools can approximate multi-element antenna radiation resistance if used carefully. E&M modeling tools become indispensable for 2 and 3D structures. The methods and definitions in this paper, however, can for a basic understanding, necessary for calculation of Rr in more complicated antenna arrays.

Basic Concepts

"Radiation resistance," R_r, is the result of the antenna coupling (losing power by radiating) RF power into a medium, usually "free space." Whenever power is dissipated or "lost" a resistance is involved. In any real antenna there are also resistive losses that are not part of the radiation resistance. The power lost to ohmic resistance is dissipated as heat, not as "radiation" and is usually designated as R_L, or "loss resistance." It is relatively easy to visualize loss resistance in antenna elements (wires or tubing) or the ground. More difficult to conceptualize is "losing" power to space.

If "space" "accepts" RF energy by providing a medium for that power, it must have some type of impedance. An analogy is the transmission line. When the line is matched (SWR = 1:1) then the voltage and current are in phase and their ratio is equal to the characteristic impedance of the line. This is just Ohm's Law. If we have an infinite transmission line, it will accept RF power and appear as a pure resistance, yet there is no resistance (in a perfect line). We can think ofZ_0 (the characteristic impedance of free space) comparable to Z_0 (the characteristic impedance of a transmission line).

In any calculation involving impedance, Ohm's Law and the power law can be applied. $P = IE$, and/or $Z = E / I$. When electromagnetic energy is radiated into space there are magnetic and electric field components of the wave. These fields are measured in volts/meter and amperes/meter. So, the characteristic impedance of free space is: $Z_0 = V_m / I_m$. The "meter" terms cancel so we are left with a simple Ohm's Law calculation. This ratio of the electric and magnetic field values is a result of the permittivity,ε_0 and permeability, μ_0 of free space, thus Z_0 is also =

$$\sqrt{\frac{\mu_O}{\varepsilon_O}}$$

As an aside, the speed of light through any medium is:

$$c = \sqrt{\frac{1}{\mu\varepsilon}}\,.$$

Obviously the speed of light is intrinsically related to R_r. Furthermore, the speed of light and the impedance of *any* medium are also based on these two related equations. In free space and the far field of the antenna, the ratio of values of the electric and magnetic field are *always* constant. Again, the ratio is defined by $Z_0 = 377\ \Omega$, often abbreviated as $120\ \pi\ \Omega$.

The calculation of Z_0 of a transmission line using air (free space) as its dielectric is given by Equation 1.

$$Z_0\left(\text{coax}\right) = \frac{1}{2\pi}\sqrt{\frac{\mu_0}{\varepsilon_0}}\ ln\frac{D}{d} \qquad \text{[Eq 1]}$$

where D is the inside diameter of the coaxial shield and d is the outside diameter of the internal conductor. Thus, the characteristic impedance of an air core transmission line can be calculated the same way as the characteristic impedance of free space, with the difference being that the electric and magnetic fields are "trapped" between the two conductors. Inside a transmission line the waves are "guided," in free space they are "radiated."

Radiation Resistance Defined

We can approach deriving a general equation for R_r by using several different methods. In this paper I will attempt to show that the value of R_r given by different equations by Kraus, in effect, reflect the same definition.

Let's begin with a simple method to form an intuitive understanding; the power law. In principle, we can find any value of R by Ohm's Law as above or by the power law: $R_r = P / I^2$.

Imagine an isotropic transmitting antenna in free space. Surrounding the antenna of interest is a large imaginary sphere. The radius of the sphere is large compared to the wavelength of operation (in the far field of the antenna). The sum of all *power* propagating through this sphere is the same as the total power radiated by the antenna. The RF *current* at a *maximum* point along the antenna length represents the current value we use in the power law equation. Thus if we know the total power radiating away from the antenna in free space and the current at a maximum point along the antenna, we can directly calculate the radiation resistance using Equation 2.

$$R_r = P / I^2 \qquad \text{[Eq 2]}$$

This is the very simple and intuitive general form equation for radiation resistance for an isotropic antenna.

If the antenna is lossless, then these two terms will exactly define R_r. Since losses (usually series losses) exist, however, the current will be a bit higher for a given radiated power than R_r. The radiation resistance can never be equal to the feed point impedance because of losses (unless you use a super-conducting antenna). The same thing can be said about a pure reactance, however. That should not imply that we despair and forget about a proper definition for X_c, X_l, or R_r!

For non-isotropic antennas, we need a bit of refinement. As a general definition, Kraus defines R_r as given by Equation 3.

$$R_r = \frac{S(\theta,\phi)_{max}\, r^2 \Omega_A}{I^2} \qquad \text{[Eq 3]}$$

This is the general-form equation for radiation resistance. Two of the three key terms we have already mentioned: power and current, so this is really just a special expression of the power law ($R = P / I^2$, but with some important subtleties. As in the simple case given earlier, the power term is in the numerator. For non-isotropic cases, however, the term $S(\theta,\phi)_{max}$ is a necessary refinement of the simple isotropic solid angle of 4π, a field power density (called the Poynting vector) over the spherical coordinates (like longitude and latitude). In other words, the power term is now the sum of all the power propagating through *a portion* of the imaginary sphere instead of the entire sphere we discussed earlier. In this case, Kraus is using the point that is the maximum power point on the sphere's surface — more on this later. In the isotropic case, as above, all points on the sphere have equal power density, so Equation 3 simplifies to the simpler power equation.

Moving through the numerator, the sphere has a radius of r, and then there's a possibly confusing term, $\mathbf{\Omega}$. Usually $\mathbf{\Omega}$ designates "Ohms," but not in this case. Here, $\mathbf{\Omega}$ designates a solid angle, which in turn defines a portion of the sphere's surface (like the Pacific Ocean defines a portion of the surface of the spherical earth). Both ohms and solid angles are needed in this paper, so I'll follow a convention that uses bold font for the solid angle term. With solid angles, 4π defines the entire sphere, 2π a hemisphere, and so on. So, an isotropic antenna will radiate with a pattern of 4π, equal power in all directions. Then Ω defines the portion of the sphere that is the 3 dB beamwidth area surrounding the point of maximum power propagating through the sphere, or $S(\theta,\phi)_{max}$. The RF current squared term, I^2, appears on the antenna element at a *current maximum*.

So we see in this general equation that R_r is a function of three fundamental terms — total power radiated, antenna current, and the beamwidth of the antenna pattern. Most readers will recognize that beamwidth is also a function of antenna gain, where an isotropic antenna has an aperture (gathering area) of $\lambda^2 / 4\pi$. Gain is proportional to aperture, so an antenna with 3 dBi gain will have an aperture of $2\lambda^2 / 4\pi$, since 3 dB is a power difference of 2. An isotropic antenna has a solid angle, $\mathbf{\Omega}$, of 4π, therefore, $4\pi\ \mathbf{\Omega} = 4\pi / G$, where G is the power gain, in this case 1, assuming no loss. Thus, the higher the gain of an antenna, the greater the antenna's aperature and the smaller the antenna's solid angle defining the 3 dB beamwidth.

From an intuitive view, imagine that it is "more difficult" to radiate into a smaller portion of free space than the full sphere of free space, so the smaller the solid angle ($\mathbf{\Omega}$) the higher the gain, and the lower the R_r.

An example to illustrate this effect is to compare a "perfect" ground mounted ¼ λvertical antenna over a perfect conductive ground ($R_r = 36\ \Omega$) and a ½ λ dipole in free space ($R_r = 73\ \Omega$). The maximum broadside gain of the vertical is 5.14 dBi and the gain of the dipole is about 2.14 dBi, about 3 dB difference, or a power gain difference of exactly 2. The R_r difference is also 2, indicating a linear relationship between antenna power gain, antenna aperature, $\mathbf{\Omega}$ and R_r. Additionally, it is easy to imagine that the vertical antenna is only radiating into one hemisphere here, defined by the hemisphere above the ground plane, and the dipole is radiating into both hemispheres (no ground to divide free space), Thus the solid angle, $\mathbf{\Omega}$, is also half for the vertical compared to the dipole.

Equating Kraus' Radiation Resistance Equations

Of special importance for radio amateurs is the value of R_r for vertical antennas. In a previous paper, I provided a detailed explanation of R_r for vertical antennas.[3] For vertical antennas Kraus gives the following equations.

$$A_e = \frac{h_e^2 Z_o}{4R_r} \qquad \text{[Eq 4]}$$

Rearranging terms we get the radiation resistance for a vertical antenna:

$$R_r = \frac{h_e^2 Z_0}{4A_e} \qquad \text{[Eq 5]}$$

Where A_e is the antenna aperture, measured in m^2 (directly proportional to gain)h_e is the antenna height measured in meters, where:

$$h_e = \frac{I_{ave}}{I_0} h_p \qquad \text{[Eq 6]}$$

where I_{av} is the *average* current along the vertical antenna element, I_0 is the *maximum* current along the antenna, and h_p is the actual physical length of the vertical. The impedance of free space, Z_0, is measured in ohms, and R_r is the radiation resistance, also measured in ohms. Here, the terms containing linear dimensions cancel and we are again left with ohms.

Any definition of radiation resistance (defined by ohms) *must* yield only ohms, unless you want to redefine other terms as well to make your equation work! Valid equations defining radiation resistance simply substitute other terms that, in turn, must be valid. Let's see if the Kraus equation for the general case (Equation 3) can be shown to be the same as the particular case for a vertical (Equation 5).

We can answer this question by substituting terms for their equivalents. We will assume an isotropic case for both, thus Equation 3 becomes Equation 7.

$$R_r = \frac{S(\theta,\phi)_{max} 4\pi}{I^2} \quad \text{[Eq 7]}$$

where $4\pi = \mathbf{\Omega}$ for an isotropic antenna and Equation 5 becomes Equation 8.

$$R_r = \frac{4\pi h_e^2 Z_0}{4\lambda^2} \quad \text{[Eq 8]}$$

or

$$R_r = \frac{\pi h_e^2 Z_0}{\lambda^2} \quad \text{\{Eq 8A]}$$

where $\lambda^2 / 4\pi$ is the aperture of an isotropic antenna, or A_e. Thus we have assumed the isotropic case for both equations.

If Equations 7 and 8 can be shown as equal, then R_r has the same definition for the general case and the special case for the vertical antenna. Again, for the isotropic case, $S(\theta,\phi)_{max}$ represents the total power radiated by the antenna. Kraus also shows that the total radiated power can be defined as the value of the square of the radiated magnetic field: $H^2 = S(\theta,\phi)_{max}$, where

$$H = \frac{\sqrt{Z_0}\, 2\pi f I h_p}{4\pi c}$$

and where c is the speed of light. This is simply another form of the power law, where power is a function of the current squared, and a magnetic field strength is directly proportional to the current creating the field.

Therefore, radiated power is

$$H^2 = Z_0 \frac{4\pi^2 I^2 h_p^2}{16\pi^2 \lambda^2},$$

or

$$H^2 = \frac{Z_0 I^2 h_p^2}{4\lambda^2}$$

Now we can multiply by 4π ($\mathbf{\Omega}$) (for the isotropic case) and divide by$I^2_0$to derive the following equations.

$$R_{r=} \frac{Z_0 I^2 h_p^2}{4\lambda^2 I_0^2} 4\pi \quad \text{[Eq 9]}$$

or

$$R_{r=} \frac{Z_0 \pi I^2 h_p^2}{\lambda^2 I_0^2} \quad \text{[Eq 10]}$$

or

$$R_{r=} \frac{Z_0 \pi h_e^2}{\lambda^2} \quad \text{[Eq 11]}$$

Equation 11 is R_r derived from the general equation for an isotropic antenna. Now let's derive R_r from Kraus's special equation for vertical antennas (Equation 5). Again for the isotropic case, substituting $\lambda^2 / 4\pi$ for A_e, we have Equation 12.

$$R_r = \frac{4\pi h_e^2 Z_0}{4\lambda^2} \quad \text{[Eq 12]}$$

or

$$R_r = \frac{Z_0 \pi h_e^2}{\lambda^2} \quad \text{[Eq 13}$$

Equation 13 is identical to Equation 11 as long as the $\mathbf{\Omega}$ coincides with the value for A_e as explained above. If we change the gain from isotropic, both equations simply change value by identical coefficients as described earlier. (The isotropic case simplifies the derivation considerably, however).

Thus Kraus presents only one definition of R_r. Furthermore, we can derive R_r by working from an integration of the radiated output power together with *maximum* antenna current and the gain of the antenna (Equation 3), or we can derive R_r from the *distribution* of current on the antenna, the *maximum* current on the antenna, and the gain of the antenna (special case for the vertical antenna, Equation 5). Thus we can see that R_r is a function of all these terms. It all depends on using the proper terms to set up more convenient equations for specific applications. Like other terms used in Physics, R_r is a well-defined term that can be derived using standard equations, including the most fundamental equations of electromagnetic science: Maxwell's Equations.

Radiation Resistance and Feed Point Impedance

The impedance at an antenna's feed point depends upon the frequency of operation, physical characteristics of the antenna, the current distribution, its relationships to objects, the impedance of free space, and the *point* on the antenna where the power is applied. All these conditions result in a ratio of voltage and current (the real part of the feed point impedance) and the phase relationship between voltage and current (the reactive part of the feed point impedance).

Therefore, the same dependencies that determine the feed point impedance also affect the radiation resistance but the calculations to derive the two terms use different equations because they are not identical. I will attempt to offer a non-mathematical description of the necessary conditions for feed point impedance to equal radiation resistance (assuming no loss).

Thus far I have hinted at a basic relationship between radiation resistance and feed point impedance — at a current maximum along an antenna. This is an important first step, but we need some refinement.

If we measure the feed point impedance (resistance and reactance) as purely reactive (no real part of the impedance) then there is no power loss and thus no radiation. Of course there is *always* some loss in real antennas or circuits. If there is a resistive portion of the feed point impedance then power is being lost as either heat (conductor loss) and/or radiation. The feed point impedance and radiation resistance are *never* equal because there is *always* resistive loss (with the unlikely exception of using a superconductor antenna). In the case of an antenna, where the usual desired effect is to minimize loss and maximize power transfer to (or extraction from) free space we can express this often published relationship as antenna *efficiency*.

$$Eff = \frac{R_r}{R_r + R_l} \quad \text{[Eq 14]}$$

Eff is the antenna efficiency, R_r is the radiation resistance and R_l is the ohmic resistance resulting in power dissipated by heat.

Mythology

Myth #1: Radiation resistance is a "part" of the feed point impedance.

This is true only in specific cases and is a major source of confusion as a general definition. For example, if we center feed a ½ λ resonant dipole in free space, the feed point impedance is about 73 Ω of pure resistance. The radiation resistance is also about 73 Ω. The feed point impedance will probably be measured at a bit higher value than 73 Ω because of ohmic losses (heat) in the antenna. If the ohmic losses are 1 Ω, then the feed point impedance would be 74 Ω, and by Equation 14, the antenna efficiency would be about 98.6 %. In this special case, Myth #1 is true. In this special case, the transformation of source impedance (feed point) to load impedance (radiation resistance) is 1:1.

If we feed the dipole off center, however, let's say at ¼ of the distance from one end instead of half way, the feed point impedance is 138 Ω real but the radiation resistance is still 73 Ω and the ohmic resistance remains very small. If Myth #1 were true, then we have 65 Ω of unaccounted resistance. The ohmic losses are still only due to about 1 Ω, so subtracting 73 from 138 is meaningless. The feed point impedance has simply been "transformed" by moving it off the point of current maximum along the dipole. Since the values of current and voltage change along the length of an antenna, antenna elements can become impedance transformers for feed points. In the case of a ½ λ dipole, however,

R_r remains constant no matter where the feed point is placed along the antenna, but the feed point impedance changes dramatically with changing the feed point location.

Myth #2: Radiation resistance is equal to feed point impedance plus losses in a center fed antenna.

Again, as above, this is true only in a special case. Consider a center-fed folded ½ λ dipole. The feed point impedance is about 300 Ω, but the radiation resistance remains the same as a single-wire dipole, about 73 Ω. Assuming that a folded antenna has 4× the radiation resistance of a single conductor antenna is a common error. Antenna elements (as well as transmission lines) can also behave as transformers as in the cases of folded antennas, while terms defining radiation resistance remain constant. This is also another obvious case where Myth#1 is false. (As an aside, folded dipoles exhibit lower *Q* than a single-wire counterpart, making them more broad-banded).

The separation of the two conductors in a folded dipole is assumed to be a very small fraction of a wavelength. The currents flowing on adjacent points of the two conductors simply add when forming the radiation wave. If the currents are in phase and equal (another assumption of the folded dipole), the effective current is doubled (as far as radiation is concerned), but the feed point is connected to only one conductor. This result is the feed point current is ½ the total effective current at that point, resulting in a 4× increase in feed point impedance, but, again, R_r remains constant at 73 Ω.

Another example is a two-element collinear antenna, which is actually a full-wavelength dipole fed in the center. The radiation resistance increases to near 100 Ω, but the feed point impedance is over 1000 Ω. Again, there is no direct relationship between feed point impedance and radiation resistance.

Myth #3: The feed point impedance of a base-fed vertical is radiation resistance plus the antenna losses.

This is only true for single conductor verticals that are electrically ¼ λ or shorter. A perfect ¼ λ vertical over a perfect ground will have a radiation resistance of about 36 Ω, the same as the feed point impedance.

Now let's place a capacitance hat on the ¼ λ vertical, which also has an equivalent electrical length of ½ λ. Instead of the current maximum appearing at the base, the current maximum is now at the top of the vertical. The radiation resistance remains the same at 36 Ω, but the feed point impedance is over 1000 Ω. So much for using the base fed vertical myth.

As the vertical is made longer than ¼ λ, the feed point impedance is no longer the same as the radiation resistance. This is most dramatically shown for a ½ λ (actual height) vertical whose feed point impedance is over 1000 Ω of real impedance value, yet the radiation resistance is about 100 Ω.

Let's look at another example: a folded ¼ λ vertical. In this case we have a two ¼ λ wires closely spaced and shorted at the top. One wire is fed against the ground, the other is connected to ground, thus appearing to be a folded dipole with the ground acting as counterpoise. The radiation resistance again remains the same 36 Ω, but the feed point impedance is now 144 Ω. Again we see an impedance transformation, but no effect on radiation resistance.

Myth #4: The feed point impedance is equal to the radiation resistance plus losses only at a current maximum on the antenna.

This is getting closer to a correct correlation, but the examples of both the horizontal and vertical folded antennas prove this general statement to be untrue. We can now define a set of practical conditions (especially for most amateur work), however, where the feed point impedance actually equals the radiation resistance.

Relationship Between Radiation Resistance and Feed Point Impedance: The real portion of the feed point impedance equals the radiation resistance plus losses of the antenna only for single-conductor antennas fed at a current maximum along the antenna.

The feed point will coincide with a current maximum at the center of a balanced antenna that is less than or equal to an electrical ½ λ long. It will also coincide with the center of a horizontal antenna that is an odd number times ½ λ.

For base-fed vertical antennas, the feed point will be at a current maximum when the electrical length of the vertical is less than or equal to ¼ λ or odd multiples of an electrical ¼ λ.

For other situations, intuition easily breaks down and an analytical tool becomes invaluable. Current maximums are conveniently illustrated in many antenna simulation software tools. For example, *EZNEC* shows current values along conductors in all of its simulations. Therefore, if your feed point is located at a current maximum, the real portion of the feed point impedance will be the simulated radiation resistance plus losses. Any statement equating radiation resistance plus losses and feed point impedance should not appear as "general rule" statements but rather include a brief description of "why" for some set of special cases.

Another complexity: In antennas longer than ½ λ the current distribution (and thus the radiation resistance) can be changed by changing the feed point location. So, when calculating and/or measuring the location(s) of current maximum(s) along an antenna element, be careful that key terms that define R_r are often changed by changing the feed point position.

In amateur applications, radiation resistance is most often important in vertical antenna installations, especially when the vertical is shorter than ¼ λ, and especially critical in HF mobile installations. In these cases the above definition does indeed apply. The common mistake, however, is to apply the definition to a more general case, which usually leads to mistakes. In almost every amateur vertical antenna installation, loses will be series ground losses. In mobile low-band antennas, however, the conductor losses of the antenna proper may also play a part as the radiation resistance may be *milliohms.*

For a much deeper understanding of the terms used and derivations presented in this paper the reader is invited to read the three references given. The Kraus text develops the terms and formal proofs using advanced mathematics, especially integral and vector calculus. In two earlier *QEX* articles, I attempted to simplify the complexity needed to quantify antenna theory, in this case the Kraus text.[2, 3] This paper, in turn, focuses specifically on a deeper treatment of radiation resistance and the often-confused relationship between radiation resistance and feed point impedance deriving fundamental theory and derivations from the three references.

Bob was first licensed in 1966 at age 15. He is an ARRL Life Member and Technical Advisor. He is currently a senior RF engineer for Trimble Navigation, working in Research and Development for advanced GPS and RFID systems, including antenna design. He has a BS in Physics from the University of Oregon and has published more than 60 papers and articles in professional and amateur publications on RF engineering topics and has been awarded 6 patents.

Bob's current Amateur Radio interests include low-band and 6 meter DXing, with 9BDXCC (160-10 meters), DXCC Honor Roll, accomplished using only tree-supported wire antennas, CWDXCC Honor Roll, 5BWAZ. Beside an Elecraft K3 transceiver and a few accessories Bob's entire station is homebrew, including the first tower he ever owned. He published the first conceptual diagram of an SDR in 1988, published the first paper describing the use of DDS in FM broadcast exciters, dramatically improving the linearity of analog FM stereo and now the industry standard, designed the first MOSFET-ring RF mixer (Si-8901), as well as numerous other contributions.

Notes

[1] John D. Kraus, W8JK, *Antennas*, 1988, McGraw-Hill.

[2] Robert J. Zavrel Jr, W7SX, "How Antenna Aperture Relates to Gain and Directivity," *QEX*, May/Jun 2004, pp 35 – 38.

[3] Robert J. Zavrel Jr, W7SX, "Maximizing Radiation Resistance in Vertical Antennas," *QEX*, Jul/Aug 2009, pp 28 – 33.

Tom Williams, WA1MBA

Microwave Update 2013

78 GHz LNA Wrap-Up

1 Project Review

A group of amateurs got together in 2005, purchased some MMIC's, and decided we would try to make a useful LNA for the 4 mm band. I slowly developed bias and RF networks. The first three years produced oscillators. Please read reports in every Microwave Update from 2006 until the present. Eventually I learned and built a first fully operational and practical unit, and integrated into a radio in 2011. Results were reported at MUD 2012, assured reproducible and then I took orders to fill the amateur need in late 2012. Construction was carried out throughout the winter and spring of 2012/2013 with delivery in late spring. This report serves as the wrap-up of the entire project.

2 Build

I orchestrated the purchase and construction, and personally performed pre-testing, testing, troubleshooting, measurement, orders and delivery. I relied heavily on my detail assembler Tom Sawyer who did the part attachment (silver epoxy gluing), ribbon bonding, and performed re-work. I also hired an assembler for the bias boards, and borrowed a sweeping NF system from a friend KB1IPR. Altogether, there were 45 units built – 36 of one style and 9 of another.

2.1 Blocks

My scrawled block designs from first concept to final design were converted into Solid-works models and AutoCAD Drawings by a friend, Don Verrastro to whom I remain indebted. N0IO, Mark Lewis machined all of the prototype blocks throughout the development, as well as the blocks for this build. Machining millimeter wave blocks with integrated waveguide requires considerable precision, not just to keep tolerances, but in describing tool paths (and other tricks) which result in minimal burrs and tool marks. Often there are small errors in designs which a good machinist like Mark will find and consult to fix. All this attention to detail is what makes professional millimeter wave devices work, and was just as necessary for these amateur LNA blocks as for professional grade. A sufficiently large number of blocks were fabricated because we needed to get a head-start on this very time-consuming step when we did not know how many orders there would be. Fifty of the "Through Style" blocks and twenty-five of the "One Sided Style" blocks were fabricated. Once received from the platers, I had to match up top and bottom pieces (Mark serialized the tops and bottoms of each block half together). Then they were inspected to assure that any remaining burrs were removed, and that if there were any cosmetic problems, such blocks would be set aside. See figure 1 where inspection and first assembly was performed. Notice the set of color-coded fine tools in the upper portion of the photo. Those are spatula/chisel tools with flats measuring 5.0 to 2.0 thousandths of an inch across.

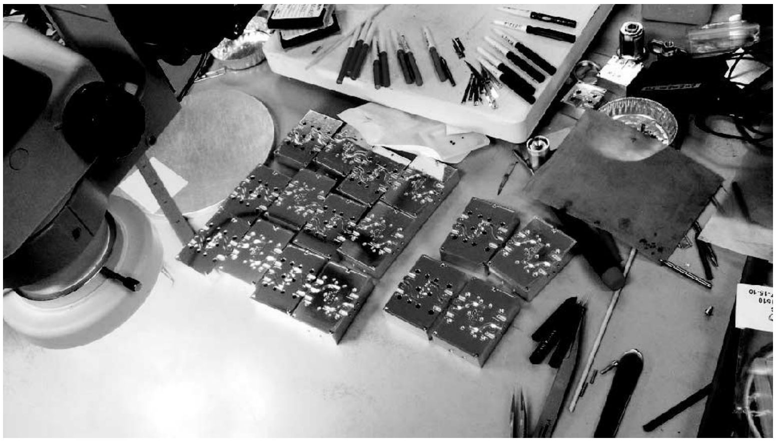


Figure 1. Here we see some of the split blocks being inspected prior to insertion of components.

2.2 Parts

Although the number of electronic parts in the LNA is not so large, they include a bias board which must create a negative bias and sequence it properly with the positive bias. There are very precise feed-through pins (which must not short against the walls), microwave single-layer chip capacitors, of course the MMIC amplifier chips, and a set of RF boards made of millimeter-wave capable material, 0.005 inch thick, with printed microstrip patterns kept to +/- 0.001 inch. Assuring sufficient parts was a task unto itself, especially considering the need to purchase some parts, such as microwave chip caps in fairly large volumes (400). The price paid would just cover all costs if I got a reasonable price break from volume purchases and achieved an acceptable level of re-work. For instance, the MMIC was $53 in small quantities, but $38.50 each in the 100 quantity needed for this build.

2.3 Batches

Builds were performed in batches, as is often done in professional operations. This is more efficient, and if there are problems, minimizes the rework. We had very few parts which required re-work, and there were no errors made that propagated across an entire batch. Figures 2 A,B,C show the growing number of completed units as the batches were constructed.

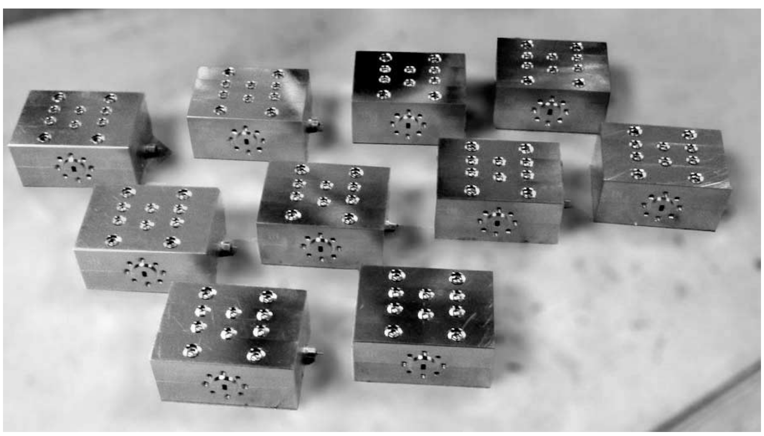


Figure 2 A. Here is the first batch of units after assembly. The first batch proved that we could get good noise figure and what care was needed regarding dressing of the RF ribbon bonds. All these are “Through Style” units.

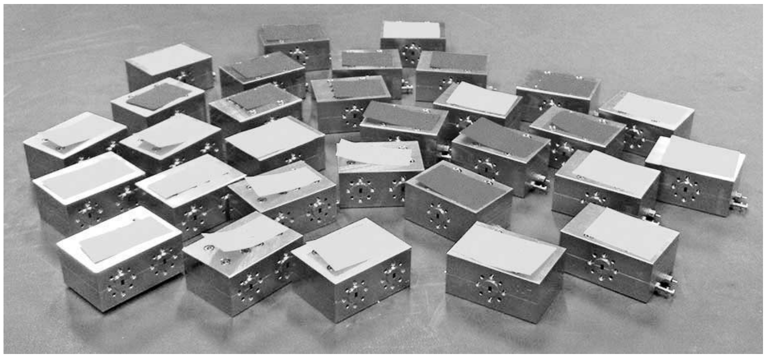


Figure 2 B. Here are the first and second batch together, totaling 30 units. The Post-It-Note labels described measurements (18 yellow) or problems (12 pink). The One Sided style units are on the left, as you can see both waveguide ports. The Through Style units are on the right.

Figure 2 C. Here we see three batches together. At this point only three of these 34 built units still needed re-work (there were 11 more in the final batch being built when this picture was taken). A total of 45 were built.

2.4 Debug

Although I was able to assemble all the bias boards during the prototyping phases, the size of this build was beyond my energy level and patience. I found an assembler who had equipment at home and was eager to perform the task of assembling nearly 50 bias boards. Her work was excellent and quite affordable. As I received them, I tested them by soldering a load network to some pads put on the board for that purpose. The network had both the plus and minus supplies connected to scope probes and the input DC went through a milliamp meter. I adjusted the power source to be current limited at just a few milliamps above the expected current draw, and immediately could tell if there was a problem. Then I was able to test the turn-on and turn-off speed, and make sure that in all circumstances the negative was on first and off last (to prevent damage to the MMIC's). The circuit and board have been described over the past few years at Microwave Update.

Once entire units were assembled, I used the same power supply setup to assure that there were no problems. If a unit drew too much current there was a short somewhere. There were some kinds of assembly errors that burned up MMIC's despite my procedure, and others which stressed them causing high noise figures. In some commercial assembly operations using similar high performance millimeter wave MMIC's, one expects to loose between 10 and 15 percent of parts. I purchased 100 LNA MMIC's, enough for 50 units if all worked. In the end we built 45 working units and had no spare MMIC's, so 10 chips were lost in re-work, or 10%. This was about what I expected.

3 Results

As mentioned in other MUD reports, and is evident by many papers on the topic, accurate measurement of noise figure at lower frequencies is difficult, and at millimeter wave frequencies very difficult. I had purchased a WR12 noise source, calibrated at 78.2 GHz, and although found it to be dead-on in one test, later discovered that its ENR was drifting day-to-day enough to cause a NF error of +/- 0.7 dB. This error was unacceptable.

3.1 Measure accurately

I had to quantify noise figure and small signal gain at a number of frequencies to meet world-wide need. My first goal was to measure all of them at 78.192 GHz, and then figure out how to measure at the other frequencies as well. I obtained a fundamental mixer, and was using a GUNN oscillator for the LO, and getting varying results with my noise source. I found that the noise source drifted from day to day, but not much during an hour or two. Also, the GUNN drifted over the course of measurement. So I borrowed a synthesized MMW LO source, where later I could dial in other LO's for measurements at other frequencies, and rid myself of the GUNN system for NF purposes.

Using a liquid nitrogen load (equipment warmed up for over one hour, and frequent cleaning of moisture from the outside of the load), results were very repeatable (see figure 3). This technique is very accurate if precautions are taken. Accuracy is derived from the fact that liquid nitrogen always boils at 77.2 Kelvin, and the room temperature can also be measured accurately. A well matched horn is pointed at a room temperature load and a liquid nitrogen load with those temperatures input to the noise figure meter.

3.2 Method(s) of measurement

My setup for 78.192 was SSB, i.e. I have an image reject filter. At other frequencies measurements were DSB. If the bandwidth is flat, (gain and NF are the same at the upper and lower sidebands), then the DSB and SSB measurements should be the same.

Figure 3. Here is a photo of the SSB liquid nitrogen NF and Gain measurement setup. The liquid nitrogen is in the Styrofoam container at the right, which also contains a block of RAM which acts as a load at its immersed temperature.

However, because of variation of both gain and NF over frequency, DSB and SSB measurements are not the same for these LNA's. Therefore, I had to warn people that if their frequency is not 78.192, and they have an image reject filter, they might get a slightly different reading of NF and gain. Conversely, those who use 78.192 and have no filter will also have a different NF and gain than reported by me.

I took liquid nitrogen measurements several times with two units, which I labeled as "golden standards" and did so at every frequency of interest. I used serial number T24 (see figure 4). It had rather good performance of NF around 4.85 dB. Then, when I wanted to test some units at a particular frequency, I would set up the synthesizer for the right LO, *connect my noise source* and unit T24 to the setup, and measure NF. I would then adjust the ENR value I gave to the NF Meter until the NF meter read the same as T24 did when calibrated with liquid nitrogen. Usually this ENR value would be close to the design value for the noise source. At this point, at this frequency, the setup was reading the correct NF and gain (same as a liquid nitrogen standard measurement).

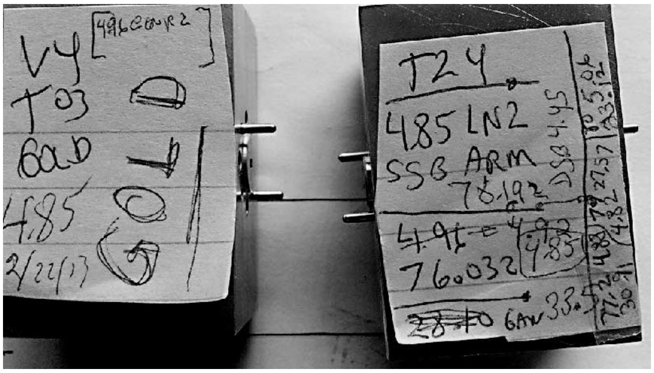


Figure 4. The "golden units" calibrated in order to measure other units. The unit on the left was used for prototypes. Unit T24 on right was used for these measurements.

If I had to measure a number of units, after every four or five I would measure T24 to see if the noise source or system had drifted. It never drifted more than 0.05 dB NF from the original measurements. I linearly interpolated between T24 measurements to further correct the recorded NF of the measured units for system drift. During most measurement sessions the drift was so low (0.01) as to be irrelevant. To check the process, one time when there was drift of over 0.03 dB, I subsequently measured one unit directly with liquid nitrogen and found the two methods to agree to better than 0.02 dB. I reported the NF of each unit to two decimal places (e.g. "4.88 dB") and believe the reported value to be accurate to better than +/- 0.05 dB at the frequency measured.

Another setup was used to sweep the band, gathering NF and gain (see figure 5). This setup used a noise source which was calibrated using established techniques. I found it to correlate with my liquid nitrogen based measurements with a maximum difference of about +/- 0.4 dB. I swept each unit and provided the graph with each unit except a few which were shipped before I had access to this setup. For those few, only gain was measured over frequency.

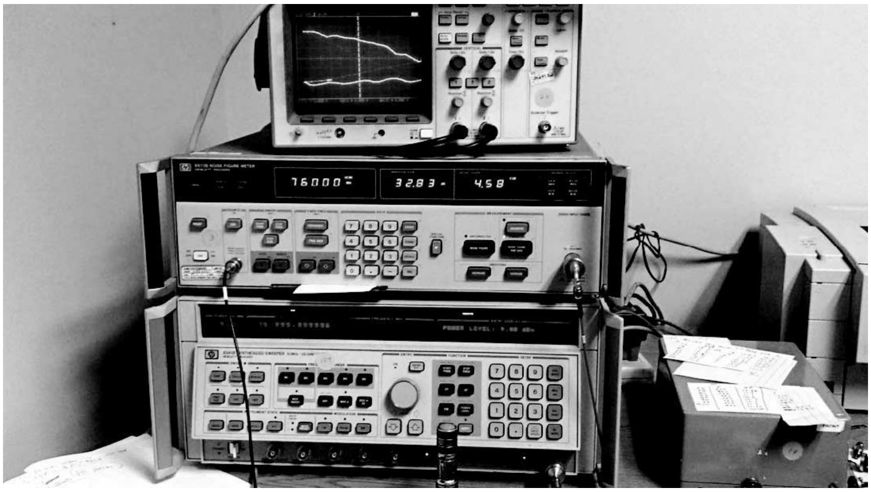


Figure 5. Here is the equipment used for swept NF and gain (thanks to KB1IPR, Rich for the use of this equipment). On the bottom is the LO synthesizer, next the NF meter, and on top a storage scope. Just off the photo to the right is a printer, and the grey box in the foreground between the printer and the equipment contains the mixer and control circuits.

3.3 Data

As far as I know, there is no official calling frequency in the US. The requested frequencies for the orders were:

Region/ Use	Frequency GHz	Number of Requests
Australia - special	80.000	2
US - special	79.000	2
One US Calling Freq.	78.192	25
Japan Calling Freq.	77.750	3
EME	77.184	4
Europe Calling Freq.	76.032	10

All devices were first tested at 78.192 because that was the only frequency where I had a reliable SSB measurement. Then, units with high NF at 78.192 were measured at other frequencies to see if they happened to have low NF there. There were two styles, 36 “Through Style” and 9 “One Sided Style”. This division actually helped as it further constrained the possibilities into two groups where the process of evaluating NF at various frequencies was performed independently. Eventually, a reasonably fair overall set of units was chosen for each frequency group (and type). Then the units were sorted and assigned first to the amateurs who were the founders of the project (except for the author) and then in the order which requests were received, with increasing NF for later requests. As can be seen from the distribution of noise figures, this method both reduced the overall noise figures of the parts delivered while also distributing them fairly so that the first orders received the best. Furthermore, no *delivered* NF was above 5.4 dB, a pretty respectable LNA in this band.

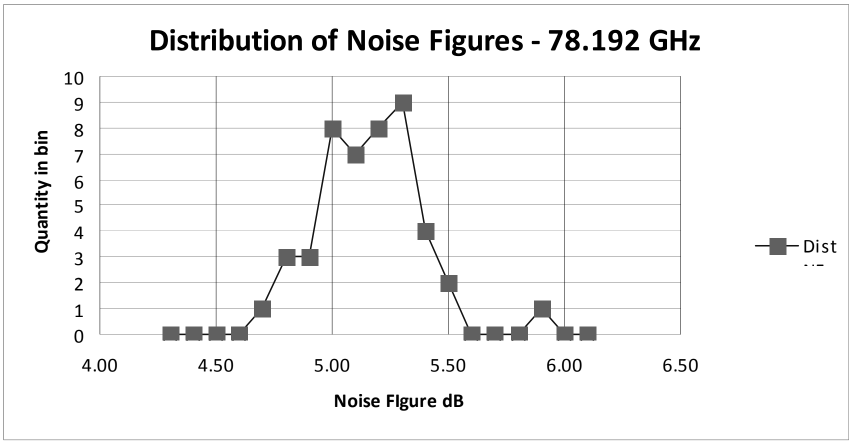


Figure 6. This chart shows the distribution of noise figures *measured at 78.192 GHz*, the most common calling frequency in the US.

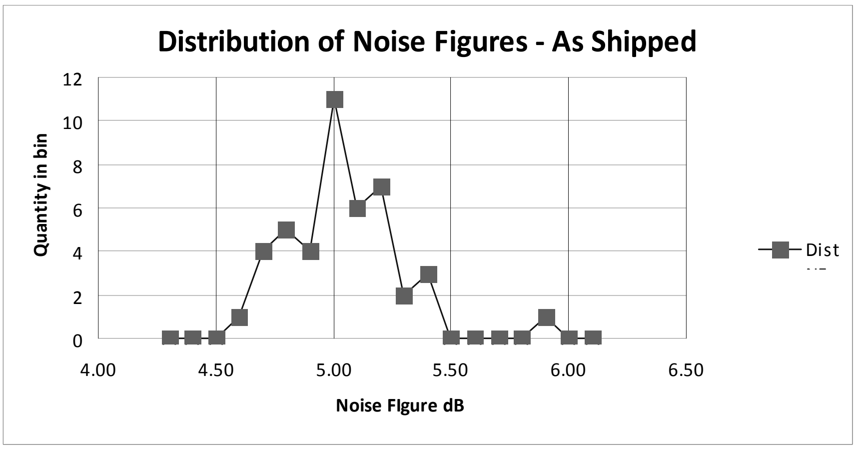


Figure 7. This is the distribution of noise figures *measured at the frequency of use* for each unit shipped. All noise figures fell between 4.6 dB and 5.3 dB. The one outlier at 5.8 dB and the three in the 5.4 bin were kept by WA1MBA for future analysis.

As shown in figure 8, the performance of T24 was quite good. This swept chart, although not highly accurate, gives a good general representation of NF and gain across the band of interest. Each horizontal division is 1/2 GHz and each vertical division is 1 dB NF and 3 dB gain. The NF chart (lower curve) runs from 3 to 13 dB, and the gain chart (upper curve) runs from 5 to 35 dB. The chart specific to each unit was shipped with it, as well as calibrated gain and NF values for the frequency of interest. As mentioned above, chart accuracy is believed to be better than +/-0.4 dB for NF and +/- 2 dB for gain.

Figure 9 shows a composite of a large number of the noise figure and gain charts.

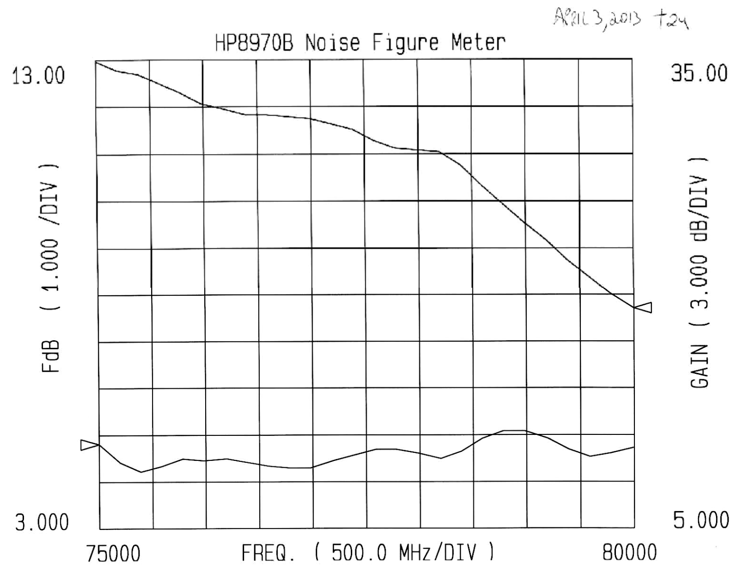


Figure 8. Above is the performance chart for T24, one of the "golden" test units. This unit was used as a standard to measure others. As can be seen, the NF was between about 4 and 5 dB. T24 was also measured using liquid nitrogen at all specific measurement frequencies in the frequency table. Except for the first few units shipped, a chart like this was included with each unit.

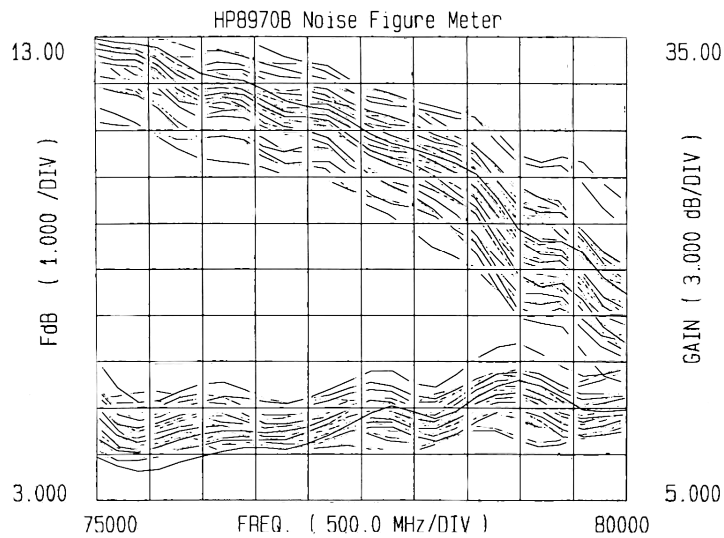


Figure 9. The chart above shows an overlay of the NF and Gain swept measurements. This gives us a feeling of the variation in completed units. The variation is caused by a combination of: MMIC parameters, placement, and bond differences. Four of the worst ones (highest NF and lowest gain on this chart) were held for future analysis.

4 Comments and Summary

After completing this task, my first inclination was to say "never again". In the end, I came very close to breaking even, and own a few early prototype units that work, one of which is in my radio. I fully appreciate why commercial suppliers of amateur LNA's are charging nearly twice as much for their units as was paid for these. I have some spare blocks and a few parts for repairs (and available if someone else wants to build).

As I get older I have less patience for this tedious work, and seem to receive more demands on my time. I know there are always going to be new challenges in amateur millimeter wave bands, and am active on them because of those challenges. Those include a moderate power amplifier for this band (there are100 and 300 milliwatt amplifier MMIC's available now) and LNAs for the 122 and 134 GHz bands (at least one chip available). I think that before any of that gets my attention I will spend some time on 78 improving my dx and increasing my QSO's, especially now that there are a few of these units in the field.

Upcoming Conferences

2014 Annual Conference, Society of Amateur Radio Astronomers

June 29 – July 4, 2014
National Radio Astronomy Observatory
Green Bank, WV

The Society of Amateur Radio Astronomers (SARA) will hold its 2014 Annual Meeting and Technical Conference Sunday June 29 through Friday July 4, 2014, at the National Radio Astronomy Observatory (NRAO), Green Bank WV. This is a joint conference with the Radio JOVE Project. Papers will be presented on radio astronomy hardware, software, education, research strategies, and philosophy. Dr Joe Taylor, K1JT, will be our keynote speaker at the Green Bank SARA Conference

For more information go to **www.radio-astronomy.org**.

Central States VHF Society

July 25 – 27, 2014
Austin Marriott South
Austin, Texas
4415 So. IH35
Austin, TX 78744
Hotel Reservation Phone:
888-253-1628

The Central States VHF Society, Inc. is soliciting papers, presentations, and poster displays for the 48th Annual CSVHFS Conference on July 25–27, 2014. Papers, presentations, and posters on all aspects of weak-signal VHF and above Amateur Radio are requested. You do not need to attend the conference, nor present your paper, to have it published in the Proceedings. Posters will be displayed during the two days of the Conference.

The papers will be published in the *Conference Proceedings*, which will be available at the conference. You do not have to attend the conference nor present the paper to have it published in the *Proceedings*. Posters describing your project can be displayed during the 2-day conference.

Presentations and Posters at the conference may be technical or non-technical but will cover the full breadth of amateur weak signal VHF/UHF activities. The presentations generally vary from 15 to 45 minutes, covering the highlights with details in the *Proceedings* paper. Topics of Interest include:

- VHF/UHF Antennas, including modeling/design, arrays and control
- Construction of Equipment – such as transmitters, receivers and transverters
- RF power amplifiers – including single and multi-band, vacuum tube and solid state
- Preamplifiers (low noise)
- Regulatory topics
- Software defined radio (SDR)
- Test equipment – including homebrew, using and making measurements
- Operating — including Contesting, Roving and DXpeditions
- Propagation – including ducting, sporadic E, tropospheric and meteor scatter
- Digital Modes – WSJT, JT65 and others
- EME (Moon Bounce).
- Digital Signal Processing (DSP)

Non weak signal topics such as FM, repeaters and packet radio are generally not considered, although there are exceptions. If you have any questions about the suitability of a topic, contact K5TRA.

If you would like to contribute a paper, presentation, or poster, please contact Tom Apel, K5TRA, *as soon as possible* with the title and a short description. Author Guidelines and other details are available at the Society website: **www.csvhfs.org/2014conference/2014callforpapers.html**

Tom Apel, K5TRA, 7221 Covered Bridge Dr, Austin, TX 78736-3344; **csvhfs2014@gmail.com**

Submissions Deadlines:
Proceedings – April 23, 2014
Presentations – June 27, 2014
Posters – June 27, 2014
Banquet Speaker
The Saturday evening Banquet Speaker will be Jimmy Treybig, W6JKV.

The 33rd Annual ARRL and TAPR Digital Communications Conference

Austin, Texas
September 5-7, 2014
Austin Marriott South
4415 S. IH-35
Austin, Texas 78704
Hotel Reservation Phone:
512-441-7900

Now is the time to start making plans to attend the premier technical conference of the year, the 33rd Annual ARRL and TAPR Digital Communications Conference. This year's DCC will be held September 5 – 7, 2014 in Austin Texas, at the Austin Marriott South. This is the same hotel as the Central States VHF Society Conference. Regular attendees will note that the conference is a couple of weeks earlier than normal this year. It is the weekend after Labor Day.

The ARRL and TAPR Digital Communications Conference is an international forum for radio amateurs to meet, publish their work, and present new ideas and techniques. Presenters and attendees will have the opportunity to exchange ideas and learn about recent hardware and software advances, theories, experimental results, and practical applications.

Topics include, but are not limited to: Software defined radio (SDR), digital voice (D-Star, P25, WinDRM, FDMDV, G4GUO), digital satellite communications, Global Position System (GPS), precision timing, Automatic Packet Reporting System® (APRS), short messaging (a mode of APRS), Digital Signal Processing (DSP), HF digital modes, Internet interoperability with Amateur Radio networks, spread spectrum, IEEE 802.11 and other Part 15 license-exempt systems adaptable for Amateur Radio, using TCP/IP networking over Amateur Radio, mesh and peer to peer wireless networking, emergency and Homeland Defense backup digital communications, using Linux in Amateur Radio, updates on AX.25 and other wireless networking protocols and any topics that advance the Amateur Radio art.

This is a three-Day Conference (Friday, Saturday, and Sunday). Technical sessions will be presented all day Friday and Saturday. In addition there will be introductory sessions on various topics on Saturday.

Join others at the conference for a Friday evening social get together. A Saturday evening banquet features an invited speaker and concludes with award presentations and prize drawings.

The ever-popular Sunday Seminar has not be finalized yet, but is sure to be an excellent program. This is an in-depth four-hour presentation, where attendees learn from the experts. Check the TAPR website for more information: **www.tapr.org**.

Call for Papers

Technical papers are solicited for presentation and publication in the *Digital Communications Conference Proceedings*. Annual conference proceedings are published by the ARRL. Presentation at the conference is not required for publication. Submission of papers are due by 31 July 2014 and should be submitted to: Maty Weinberg, ARRL, 225 Main Street, Newington, CT 06111, or via the Internet to **maty@arrl.org**. There are full details and specifications about how to format and submit your paper for publication on the TAPR website.

Even if you are not presenting a paper at the conference, plan to bring a project or two to display and talk about in the popular Demonstration Room, or "Play Room" as it is commonly known.

Letters to the Editor

An Extremely Wideband QRP SWR Meter (Jan/Feb 2014)

Hi Larry,

With regard to my article in the Jan/Feb 2014 issue of *QEX*, Glenn Pederson, WB9QIQ, brought an obvious error to my attention. In the section titled "The Math," I incorrectly stated:

We square the power reflection coefficient to get the voltage reflection coefficient, because we ultimately want the Voltage Standing Wave Ratio. That's the "V" in VSWR. Since we're still dealing with the logarithms, we obtain the square by multiplying the logarithm of the power reflection coefficient by 2.

Glenn caught the obvious error. Actually the voltage reflection coefficient is the square root of the power reflection coefficient.

I overlooked this fact because the equation was obviously correct, as it gave exactly the correct answer in all test cases. It is my simplistic explanation that is in error. I herewith correct that error with way too much math. Sorry, this is as simple as I can make it. Other approaches require manipulation of exponents. The original equation was correct simply because the reciprocal of the slope of the log detector response is exactly 40. Talk about serendipity.

Here is a corrected version of "The Math" section of my article. My thanks to Glenn for bringing this error to my attention, and subsequently checking my math.

The Math

The outputs of the coupler board are two voltages that indicate the logarithms of the forward and reflected powers from the two couplers.

The linearized logarithmic detector response from Figure 4 is given by Equation 1.

$$\text{V} = (0.025\text{ V / dB} \times 10\log \text{Power in dBm}) + 2.15\text{ V} \qquad \text{[Eq 1]}$$

The reflection coefficient, traditionally represented as the Greek character rho (ρ) is a ratio of reflected signal level compared to forward signal level. In terms of reflected and forward power, and remembering that dividing numbers is the same as subtracting their logarithms, we obtain Equation 2.

$$\rho_{\text{power}} = \frac{P_{\text{r}}}{P_{\text{f}}} \qquad \text{[Eq 2]}$$

or

$$10\log\rho_{\text{power}} = 10\log P_{\text{r}} - 10\log P_{\text{f}} \qquad \text{[Eq 2A]}$$

VSWR meters measure voltages corresponding to the reflected and forward powers to calculate reflection coefficient and then VSWR. The simplest way I can explain this is to start with the difference between these voltages, which is a ratio of the reflected power to the forward power, and which is in turn, the power reflection coefficient, ρ_{power}. So, we can write Equation 3.

$$\left(V_{\text{r}} - V_{\text{f}}\right) = \left[0.025\times\left(10\log P_{\text{r}}\right) + 2.15 - 0.025\times\left(10\log P_{\text{f}}\right) - 2.15\right] \qquad \text{[Eq 3]}$$

This conveniently simplifies to Equation 4.

$$\left(V_{\text{r}} - V_{\text{f}}\right) = \frac{\left(10\log P_{\text{r}} - 10\log P_{\text{f}}\right)}{40} \qquad \text{[Eq 4]}$$

Equation 4 can be further simplified to Equation 5.

$$\left(V_{\text{r}} - V_{\text{f}}\right) = \frac{\left(\log\dfrac{P_{\text{r}}}{P_{\text{f}}}\right)}{4} = \frac{\left(\log\rho_{\text{power}}\right)}{4} \qquad \text{[Eq 5]}$$

or

$$\left(\log\rho_{\text{power}}\right) = 4\times\left(V_{\text{r}} - V_{\text{f}}\right) \qquad \text{[Eq 6]}$$

Take the square root of the power reflection coefficient to get the voltage reflection coefficient, ρ_{voltage}, because we ultimately want the Voltage Standing Wave Ratio. That's the "V" in VSWR. Since we're still dealing with logarithms, we obtain the square root by dividing the logarithm of the power reflection coefficient by 2.

$$\log\left(\rho_{\text{voltage}}\right) = \frac{\left(\log\rho_{\text{power}}\right)}{2} = 2\times\left(V_{\text{r}} - V_{\text{f}}\right) \qquad \text{[Eq 7]}$$

So far, we've taken the two voltages, subtracted them, and multiplied by 2. Then we obtain ρ by raising 10 to the power log ρ_{voltage}. Finally we obtain the VSWR from Equation 8.

$$VSWR = \frac{1+\rho_{\text{voltage}}}{1-\rho_{\text{voltage}}} \qquad \text{[Eq 8]}$$

All this is just a few lines of code in a high level language, as Figure 6 in the original article shows.

— 73, Dr Sam Green, WØPCE, 10951 Pem Rd, Saint Louis, MO 63146; **w0pce@arrl.net**

Code Listing

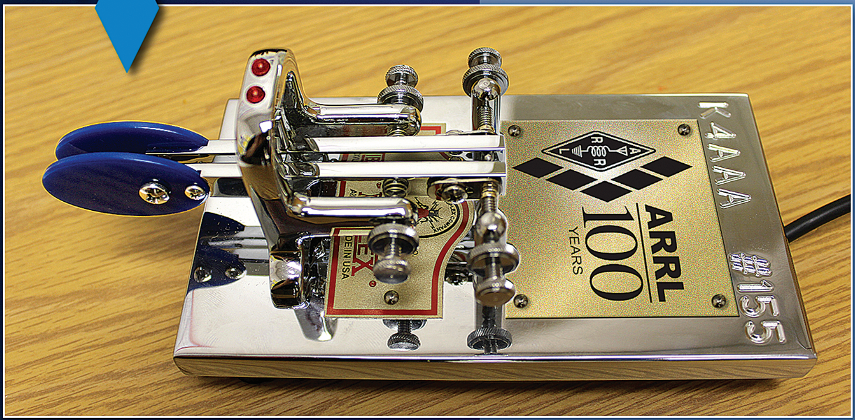
```
#include <LiquidCrystal.h>
// initialize the library with the numbers of the interface pins
LiquidCrystal lcd(12, 11, 5, 4, 3, 2);

int sensorPin0 = A0, sensorPin1 = A1;
float a,b,c,d,e,f;
float sensorValue0 = 0, sensorValue1 = 0;  // store values from sensors
float fudge = 2.40/1024 ; // for External 2.4 V reference

void setup() {
lcd.begin(16, 2);    // set up the LCD's number of columns and rows:
analogReference(EXTERNAL);
}

void loop() {
sensorValue0 = analogRead(sensorPin0);
sensorValue1 = analogRead(sensorPin1);

lcd.setCursor(0, 0) ;
//  lcd.print("W0PCE  QRP Meter  "); // normal operation
a=sensorValue0*fudge ;
b=sensorValue1*fudge ;
lcd.print("Vf=");    lcd.print(a); // diagnostic operation
lcd.print("  Vr=");  lcd.print(b);

lcd.setCursor(0, 1) ;
c=2*(b-a) ;
d=pow(10, c);
e=(1+d)/(1-d) ;
lcd.print("  SWR = ");  lcd.print(e);
delay(500);
}
```

Figure 6 — Arduino code performs WØPCE QRP SWR Meter functions.

ARRL 100 YEARS

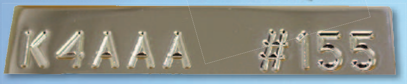
ARRL2014

ARRL Centennial Key



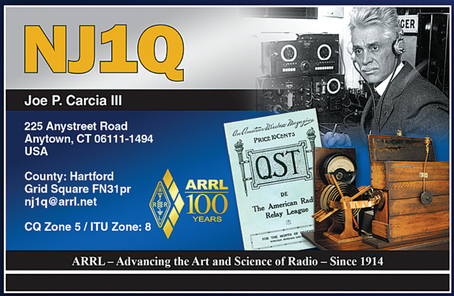
Vibroplex has partnered with ARRL to produce this commemorative key to help mark the occasion of ARRL's Centennial Year. This fully-function key will be a treasured part of your station that you'll enjoy for years to come!

- Deluxe model Vibroplex lambic paddle
- Jeweled key movements; blue paddles
- Heavy base; chrome mirror finish
- ARRL 100 Years logo plate, riveted to base
- **Etched with your CALL SIGN and serial number!**



ARRL Centennial Key
Only $249.95
plus shipping & handling
Orders placed by March 31 will ship by May 31, 2014.

* Shipping and handling charges apply. Sales Tax is required for all orders shipped to CT, VA, and Canada. Prices and proct availability are subject to change without notice.



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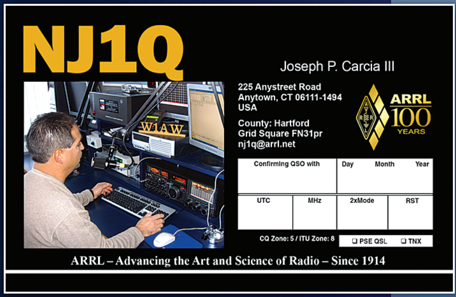

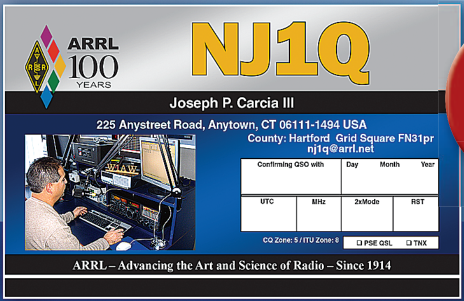

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