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QEX

January/February 2019

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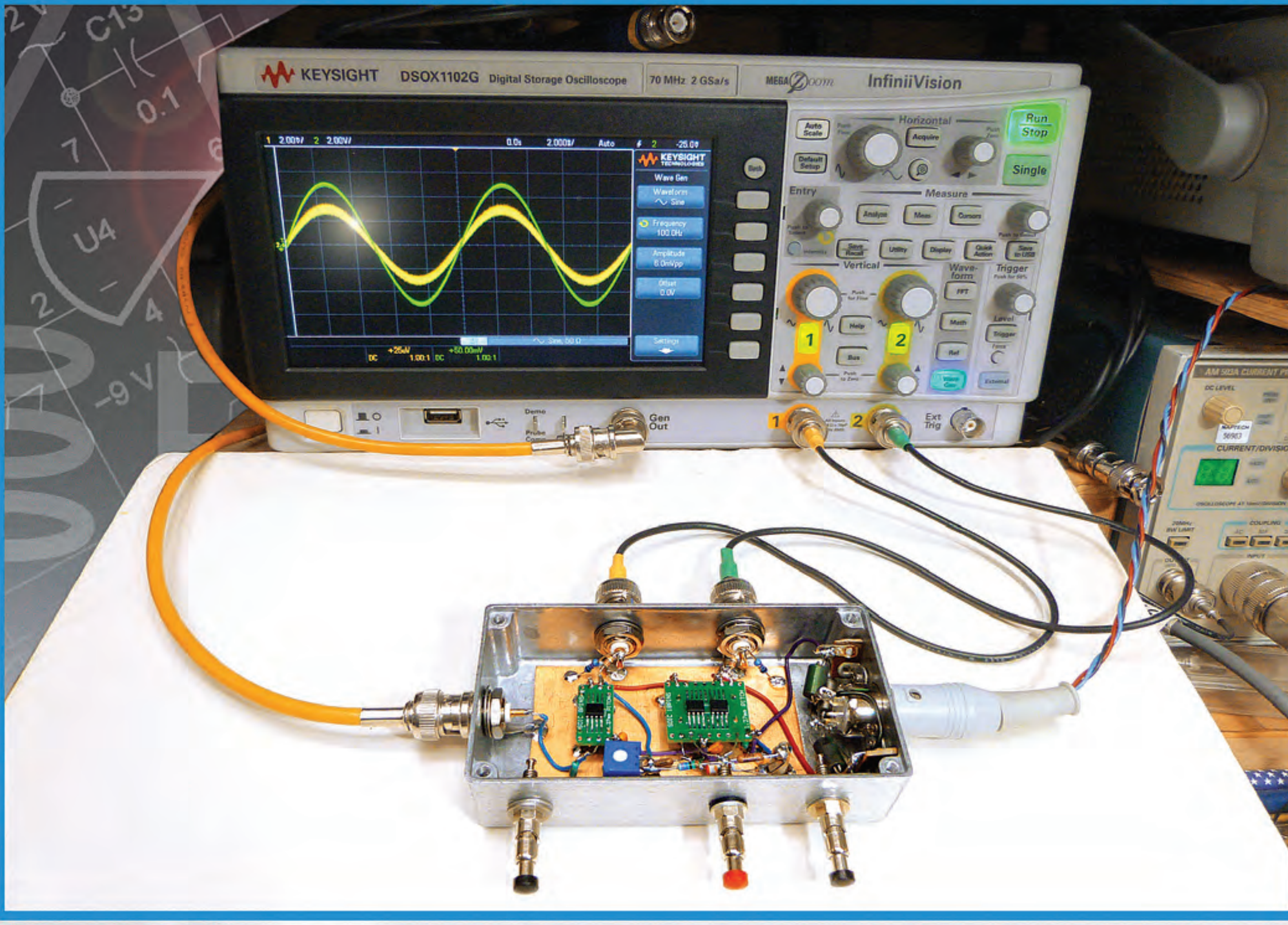
A Forum for Communications Experimenters

Issue No. 312



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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 312

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Subscription rate for 6 issues:

In the US: $29;

US by First Class Mail: $40;

International and Canada by Airmail: $35

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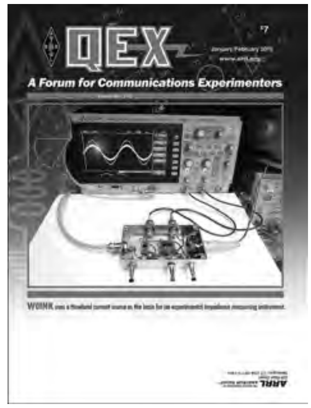
January/February 2019

About the Cover

Virgil Leenerts, WØINK, bases an experimental impedance measuring instrument on a Howland ac current source, a digital storage oscilloscope (DSO), Frequency Response Analysis (*FRA*) software and a waveform generator. The instrument needs no calibration and depends only on the measurement of ratios with a known value of a reference resistor. Using *FRA* software, Leenerts shows examples of impedance measurements of a capacitor, an inductor, a parallel RC network, telephone coupling audio transformer, and a tubular ferrite. In another example, he measures a capacitor without using the *FRA* software.



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**.

Kazimierz "Kai" Siwiak, KE4PT

Perspectives

Digital DXCC

Counting countries worked has been a passion among hams since the beginning of Amateur Radio. That passion spawned the ARRL DXCC Awards program, which is currently available in four mode flavors: CW, phone, digital and mixed modes. Digital-mode DXCC was for a long time practiced using RTTY, which itself evolved over the years to the currently popular ham version of two alternating carriers separated by 170 Hz, using 5-bit Baudot code plus one start and one-to-two stop bits, each bit 22 ms long. The DXCC award mode was called "RTTY" but the ARRL counts just about any digital mode for the RTTY award. The award is now known as the "DXCC Digital Award."

We've noted before in this column that the use of digital modes in Amateur Radio has grown dramatically in recent years. Indeed, according to the ARRL, "For newcomers, data emissions are far more popular than telegraphy" (Petition to FCC for Rule Making, February 2018). Today, a simple 'SDR System' comprising a recent transceiver plus a sound card — or an SDR radio — plus a computer running *WSJT-X* or *fldigi* or other software, enables a multitude of digital modes including FT8 and RTTY. FT8 is rapidly gaining popularity in the DX community.

Today some hams are asking for a separate DXCC FT8 award category. As this editorial is being written, the ARRL DX Advisory Committee is surveying the wishes of the ham radio community regarding the FT8 DXCC question. Insofar as *QEX* is concerned, we are a forum for communications experimenters, so tell us how you are evolving into this exciting digital world!

In This Issue

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

David M. Collins, AD7JT, designs an innovative CW keyer.

Bob Simmons, WB6EYV, introduces additional PIC programs for his VoIP board.

Thomas M. Alldread, VA7TA, builds a portable wide dynamic range RF field strength meter.

Paul Wade, W1GHZ, finds a range of measurement uses for an antenna analyzer.

Virgil Leenerts, WØINK, describes an experimental impedance measuring instrument based on the Howland current source.

Maynard Wright, W6PAP, revisits trimming the wire dipole antenna using remote measurements.

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

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Very best regards,

Kazimierz "Kai" Siwiak, KE4PT

David M. Collins, AD7JT

22111 N San Ramon Dr., Sun City West, AZ 85375; **ad7jt@cox.net**

The Ultimate Keyer

A number of unique design innovations were used to develop the CW Ultimate Keyer, which takes up less desk space than a 3 by 5 inch index card.

I have been interested in electronic keyers since I was first licensed as a teenager in the fifties. Of course, back then, all electronic keyers were vacuum tube keyers. One popular keyer of the day, the Hallicrafters HA-1 Keyer, used four dual triode 12AU7 vacuum tubes to perform a basic set of logic functions to generate self-completing dits and dahs as dictated by the key lever contact closures. With just the income from my paper route there was no way I could afford the $79.95 sticker price on the HA-1 nor did I have the wherewithal to design and build my own version.

During the seventies, the time demands of starting a career and family left little time for Amateur Radio and I let my license lapse. I retired in 2005 and my wife and I moved to our retirement home in Arizona.

Having never lost my interest in the hobby, I got relicensed in 2006 as AD7JT and started to familiarize myself with all the changes made to the hobby over the intervening years. I learned to use the keyers built into transceivers, and purchased a stand-alone keyer, but I was never completely satisfied with them. I had spent my thirty-year career working in the computer industry in both hardware and software development, and thought about applying some of that experience to creating the ultimate keyer.

I got my first opportunity to implement a keyer while working with George Heron, N2APB, of Midnight Design Solutions (**midnightdesignsolutions.com**) on the NUE-PSK digital modem. I had just completed adding a couple new features to the modem firmware when George asked me if I would be interested in adding CW to the modem's PSK and RTTY operating modes. I implemented the new mode using the keyboard for transmit and displaying the transmitted and received text on the modem's LCD display. Next, I added a connector for a paddle and implemented my first electronic keyer in the modem's firmware so text could be entered either with the keyboard or with an iambic paddle.

About that time I started to develop an interest in low power (QRP) rigs. I noticed that most QRPers had several QRP rigs and virtually all of them were CW rigs. A few rigs had built in keyers but they were usually a one chip, bare bones hardware design. Since the firmware in these keyers can only be controlled using the paddle and, possibly, a single COMMAND push button switch, setup usually involves a series of tedious text entry operations using special character sequences. Commercial keyers could be used on these rigs but there are some restrictions there, too, that keep that solution from being ultimate. I then started thinking seriously about what the ultimate keyer should be and if there would be enough interest to justify adding it to Midnight Design Solutions kit offerings.

The following describes the result of this project. While this design meets the functional goals for my ultimate keyer, I also tried to make the project small, low cost, and easy to assemble. Hence I describe several rather unique approaches to the choice of display and the packaging of the project that you may find of value. Further, this design incorporates some unusual hardware and firmware implementations that other designers may wish to study for potential use in their projects.

The Ultimate Keyer Requirement List

After building a few feasibility models of not-so-ultimate keyers, I finally settled on a list of basic requirements for my ultimate keyer. The keyer should:

- Be small size, very portable
- Have built-in visual display to aid setup and operation
- Have a simple, intuitive, and flexible user interface
- Be stand-alone (no PC required)
- Support multiple keyer modes (Iambic A, Iambic B, Ultimatic, Dot-preferred, Dash-preferred, Semi-automatic/bug, Manual/Straight key).
- Support macros (push-button activation, non-volatile storage, special accommodations for *My Call* and *Their Call*, macro chaining, looping, and nesting, keyer speed (WPM) control sequences, time delay insertion (for beacon mode), some editing support for paddle input
- Have serial interface to an ASCII terminal/terminal emulator (added convenience to setup, namely entering and editing macro text, display and/or log transmitted text, keyboard input of transmitted text)
- Include a side tone generator with built-in (mutable) speaker, phone jack with volume control, variable audio frequency
- Support practice modes.

Additionally, the following would be nice-to-have features:

- Support for high-current keying (up to 2 A)
- Support for classic radio keying (grid blocking and cathode)
- PCB mounted connectors eliminating discrete wiring
- Audio channel that mixes keyer side tone with received audio, independent side tone and received audio volume control, audio amplifier with gain control, flexible configuration
- CW audio band-pass filter.

The remainder of this article describes the final implementation of the basic ultimate keyer. I plan a future article to cover the extensions in the nice-to-have list.

Enclosure

I used an ABS plastic enclosure made by PacTec (**https://www.pactecenclosures.com**), model LH43-100. It is available in either black or bone color. With a name like Midnight Design Solutions, I chose black. The overall dimensions are 4.5" wide by 2.68" deep by 1.25" high. The front and rear panels are 4.173" by 1.062" and fit in grooves in the top and bottom halves of the enclosure. The grooves are about 0.062" wide, perfect for standard, 1.6 mm PCB stock. I envisioned the main keyer Printed Circuit Board Assembly (PCBA) replacing the original plastic front panel and thus having it serve as the outward facing front panel of the enclosure. Through-hole mounted switches, speaker, and a rotary control could mount on the front of the panel; all other components would be surface mounted devices (SMDs) mounted on the rear of the front panel. A second PCBA could hold the external connectors and replace the enclosure's standard rear panel. Figure 1 shows the front panel of the ultimate keyer, and Figure 2 show the rear panel. Additional images are on the **www.arrl.org/QEXfiles** web page.

Major Components

The next task was to come up with the detailed ultimate keyer circuit design. The goal, of course, is to meet the minimum or basic requirements with a minimum of devices and cost while fitting on the chosen enclosure's approximately 4" by 1" front panel PCBA. A secondary goal was to easily accommodate extending the basic keyer capabilities to include the items in the nice to have list. I won't go into all the rationale behind all the decisions that went into the circuit design. Instead, I'll just talk about the ultimate results.

Pushbutton Switches

Eight SPST push button switches that were to be located on the left portion of the front panel in two rows of four buttons (Figure 1). They must be small size, look attractive, and easy to actuate. These switches are to be mounted on the front of the panel so they will need to have through-hole pins positioned under the switch body to hide the through-hole pads. Such a switch is made by Gravitech (**www.gravitech.us/mipubusw2qt4.html**). The switch, or a very similar one, is readily available from multiple suppliers on eBay at very attractive prices. The switch body is about 1/4-inch square and about 1/8-inch high and has a very positive, tactile feel to it.

Speaker

The speaker size is critical and must mount on the front outward-facing side of the front panel, and it should be attractive. It is seen to the left of the control knob in Figure 1. Since I wanted to control the frequency of the side tone, a real speaker is required as opposed to a buzzer or Sonalert® device. I chose the transducer speaker model AT-1220-TT-2-R from Piu Audio (**www.puiaudio.com/**), and is available from Mouser (**https://www.mouser.com/**). This transducer has two pins located under the roughly half-inch diameter round body. The resonant frequency is about 2 kHz but it performs well over the range of 500 Hz to 9.5 kHz.

Display Module

The display is a four-character LED clock display, with each character a seven-segment LED with a colon between the second and third characters. The display comes mounted on a small (1.65" by 0.5") PCBA along with a controller (TM1637) and a 4-pin interface connector. Two pins are used for voltage (3.3 V or 5 V) and ground, and two pins are for a serial interface. The serial interface is similar to I2C but does not use an address, so a standard I2C interface cannot be used to control it. Instead, the firmware "bit-bangs" out the clock and 40-bit control data word. The module is made by Catalex and is readably available from several suppliers on eBay at very reasonable prices.

Rotary Control

My original attempts at designing the ultimate keyer used a potentiometer for speed control. I didn't like the lack of precision and the fact that it was a single-function control so I decided to upgrade to a rotary encoder with a built-in push button switch — RO1 in Figure 3. Combined use of the encoder output and the push button switch allows the firmware to navigate through an array of control states for entering a number of operating parameters with digital precision.

Micro Control Unit (MCU)

Since retiring in 2005, I have done a fair amount of embedded system design using

Figure 1 — Ultimate Keyer, front view.

Figure 2 — Ultimate Keyer, rear view.

Microchip eight and sixteen-bit PIC MCUs. Since I was pretty well set up with firmware and hardware design tools for the PIC lines, this was the obvious choice for the ultimate MCU family. I initially tried an eight-bit design programmed in assembly language but that turned out to be far more problematic than could be justified by the small price difference between the two MCU families. I wanted the MCU to be easily replaced so it had to be in a socket, which meant a 0.3” wide, dual inline package (DIP), not an SMD. I also wanted an MCU with a built-in non-volatile memory (NVRAM or EEPROM) for storing recorded messages (or macros) and a number of operating parameters such as keyer speed and iambic type. After reviewing the Microchip online product guide[1], I zeroed in on the PIC24FV32KA301-I/p, which has the ultimate keyer features shown in Table 1.

MCU IC socket

SMD IC sockets are available but they tend to be pretty pricy. As a result, I decided to modify a standard, through-hole IC socket to a gull-wing version for SMD mounting. The kit builder would just need to first bend the pins out at a 45° angle, then make a 90° bend up at the center of the pin. More details are available in the **www.arrl.org/QEXfiles** web page for this article. When the socket is soldered to the DIP SMD pads, the builder puts downward pressure on the socket to make sure all 20 pins contact all 20 SMD pads.

Interface Connector

A SMD male header mounted on the rear of the front panel connects to a female IDC connector wired to the interface connectors on the rear panel. The connector is a 2 by 5 header with gull-wing legs that solder to SMD pads on the rear of the front panel PCBA. The header comes in 40-position strips notched for easy separation into smaller header strips.

Miscellaneous Components

The remaining components are SMDs with very small footprints to enable them all to fit in the available space. Resistors and capacitors are case code 0805. The voltage regulator and transistors have SOT-23-3 footprints.

Front Panel Printed Circuit Board

Detailed images of the front panel are available in the **www.arrl.org/QEXfiles** web page. About the time I had started getting serious about the ultimate keyer, I was also investigating printed circuit board (PCB) design software. I found the freeware versions of most of the available programs have limitations too restrictive for this project. Then I found KiCad (**kicad-pcb.org/**), an open-source freeware total design package without limitations. The front panel PCB (see Figure 1) is laid out with the eight pushbutton switches in two rows on the left side of the front of the panel. The MPU and interface header are positioned on the rear side of the panel between the two switch rows. The rotary encoder is panel mounted about in the center of the PCB. The display module mounts on four standoffs behind a rectangular cutout on the right side of the panel. A semi-transparent red lens fits in the cutout. The only hardware showing are four black nylon, screws holding the display standoffs to the panel. The rear of the panel has SMD-like pads for connecting to the encoder solder pins and the display module connector.

Rear Panel Printed Circuit Board

Detailed images of the rear panel are available in the **www.arrl.org/QEXfiles** web page. The rear panel (Figure 2) mounts five connectors and a potentiometer. In most cases, the connector terminals are connected to a wiring harness made from the end of a 10-conductor ribbon cable with a 2 by 5 insulation displacement connection (IDC) connector on the other end. This connector plugs into the male header on the back of the front panel. SMD-type pads and etch connections are provided to connect the audio volume control potentiometer to the wiring harness and to the phone jack.

Schematic

The schematic showing how to interface all those components to U2, a 20-pin MCU, is shown in Figure 3.

Pushbutton Switches

The main physical layout challenge is the eight pushbutton switches. I adapted this technique from other a similar designs. This technique reduces from eight to two, the number of pins required for the MCU to read the switches. Each row of switches is sensed independently so it is possible for the firmware to simultaneously sense two switch closings, one from each switch row.

The sensing is done by first fully charging a capacitor (C2 or C3) to the logic high level (about 3.3 V) through an output pin in series with a 100 Ω resistor (R3 or R13). The pin is then changed to an input pin and the firmware then times how long it takes for the I/O pin to switch from a logic one (high) to a logic zero (low). As soon as the I/O pin switches from output to input, the capacitor starts to discharge through the chain of resistors (C3 will discharge through R6 through R11). When no pushbutton in the chain is pushed, the discharge path is the sum of the individual resistors or 15 kΩ. This is the longest possible discharge period. When one of the buttons is pressed, the discharge resistance is reduced from 15 kΩ to 1 kΩ, 2 kΩ, 3 kΩ, or 4 kΩ depending on which button is pressed. When the keyer is first powered up, the firmware determines the maximum discharge time for each row of switches and uses a pre-calculated percentage range of the maximum discharge time to determine which, if any, of the four pushbuttons is pressed. This calibrates the RC circuit and compensates for component value variations.

Once the pushbutton interface was reduced to two pins, connecting the remaining devices was easy.

Rotary Encoder

RO1 is a mechanical encoder with 24 detents per revolution. As the knob is turned, two output pins (ENC-A and ENC-B) are momentarily connected and unconnected to a third pin (ENC-C) which is grounded. The states of inputs A and B are shown in Figure 4. The grounding periods on the two pins (A and B) are offset so that when the control is turned counterclockwise, A is always grounded before B is grounded, and A is ungrounded before B is ungrounded. When the control is turned clockwise, B is always grounded before A, and B is ungrounded

Table 1.

The PIC24FV32KA301-I/P feature list.

Package	20-pin DIP
Usable I/O pins	16
DC power	2.0 V to 5.5 V
Program Memory:	11,264 24-bit instruction flash memory
Data RAM	2K bytes
EEPROM	512 bytes
System clock	Internal RC oscillator
Instruction execution rate	Up to 16 million instructions per second (MIPS)
Hardware timers/counters	Five, 16-bit
Serial interfaces	UART and I2C
Input pull-up resistors	Programmable
Field programmable	In-Circuit Serial Programming (ICSP)
Debug support	In-Circuit Debug (ICD) via two pins

before A. By monitoring A and B, the firmware can tell when the control is being turned, which direction it is being turned, and how far it is turned. The firmware activates pull-up resistors on the I/O pins used for A and B so an ungrounded pin will raise to a high logic level.

The most common (and least expensive) rotary encoders are mechanical. These encoders have a simple three-wire interface (A, B, and C; no power required) and are available with and without detents. I prefer detents so the ultimate keyer uses a mechanical rotary encoder with 24 detents. The number of detents determines the number of A and B pulses per rotation of the control knob. The detents also determine the resting points for the knob. Normally the resting points are located such that neither A or B is grounded (both high or logic 1).

There are a couple problems with

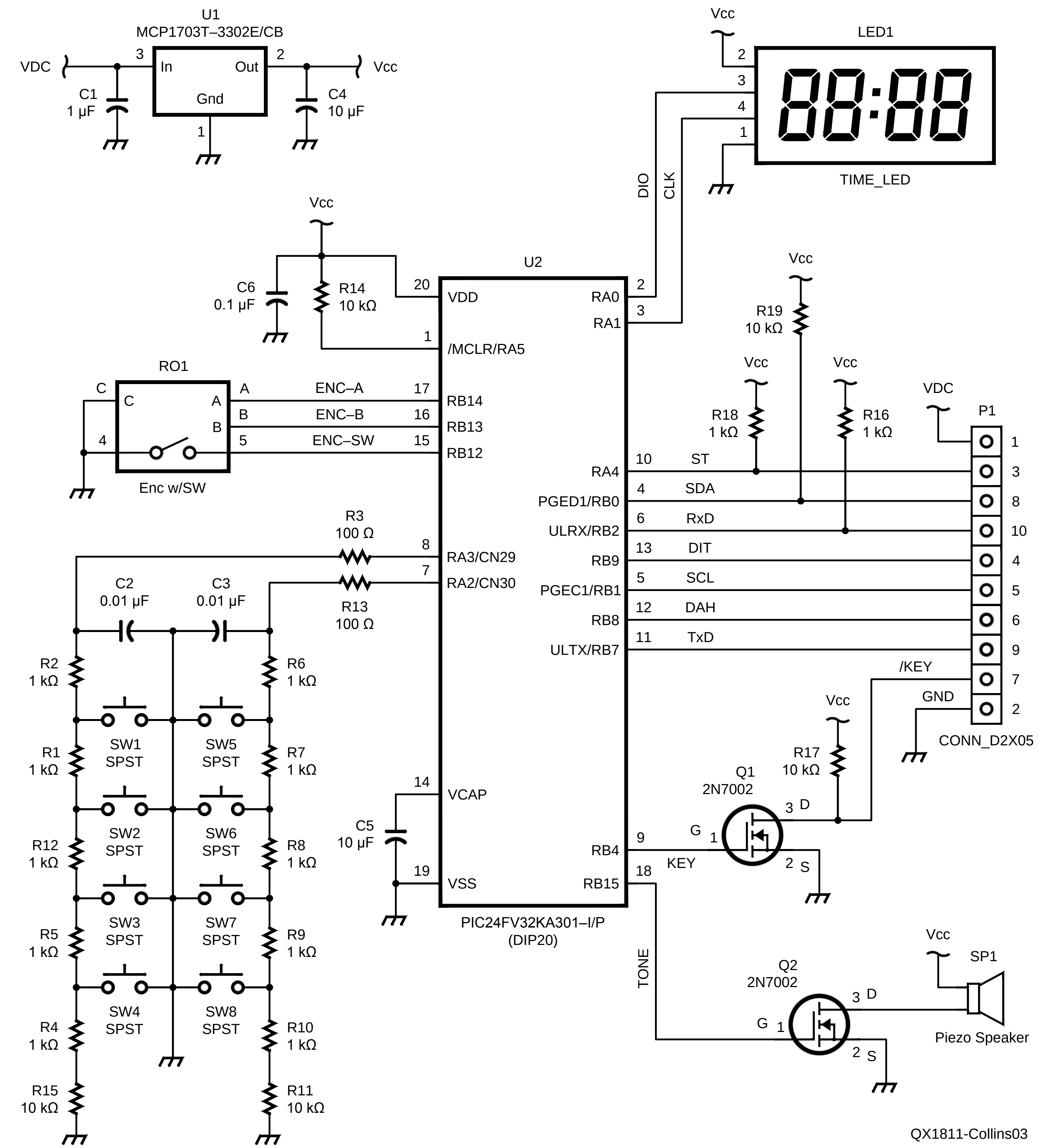

Figure 3 — Schematic diagram of the Ultimate Keyer.

mechanical encoders with detents that need to be addressed in the firmware. The first problem is switch bounce. The second, a little more subtle, is "detent bounce". When you turn the control knob and release it at an arbitrary position, it may be between detents and will move to the nearest detent, which may be in the direction you were turning the knob or in the opposite direction. Moving in opposite direction can cause a problem if not handled properly in the firmware. To resolve both of these problems, I implemented a simple state machine (Figure 5) in firmware that essentially "knows" the direction the knob was turned and can recognize the bounce-back action and ignore it.

Front Panel Assembly

The front panel contains the user interface for the keyer.

LED Display Module

The 4-digit LED display module and the red lens are assembled on the ultimate keyer front panel. The red lens starts out as a piece of red acrylic plastic approximately 1/8 inch thick. The plastic comes in one foot square sheets, which I cut to 3/4" strips with my table saw. I then use my router table to mill a 1/16 inch rabbet along the long ends. I then cut each 12" strip into 1-3/8" pieces. I use a home made jig to hold the lenses while I mill rabbets along the lens ends. The resulting raised portion of the lens fits into the rectangular cutout on the right side of the front panel PCB.

The display cutout in the front panel is made with a router during the PCB fabrication process so the cutout will generally have rounded inside corners that may cause the lens to not lay flat all around the cutout. This can be fixed by either using a file to square the rounded corners in the cutout, or to round the square corners on the raised part of the lens. The lens must also be modified by cutting 45° bevels at the two corners that will be on the right when the lens is mounted. This is to clear two of the standoffs holding the LED display in place. It is not necessary to glue the lens in place. The 9 mm standoffs leave less than 1/16" clearance between the display face and the lens. The lens is held in place by the display module itself.

Once the display module is mounted, the four pins on the display connector are bent down to touch corresponding four pads on the back of the front panel. They are then soldered to the pads.

Rotary Encoder

The rotary encoder mounts to the front panel with one nut and washer. The three pins for the encoder and two pins for the built-in push button must be bent down towards the SMD pads on the PCB directly

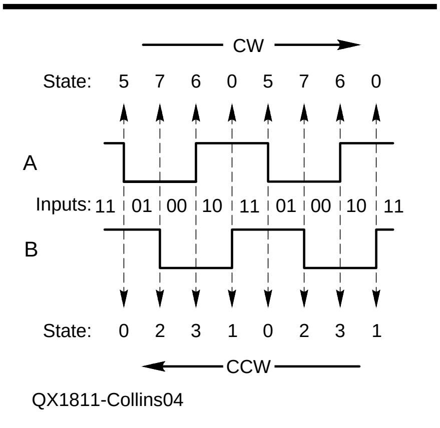

Figure 4 — Rotary encoder outputs.

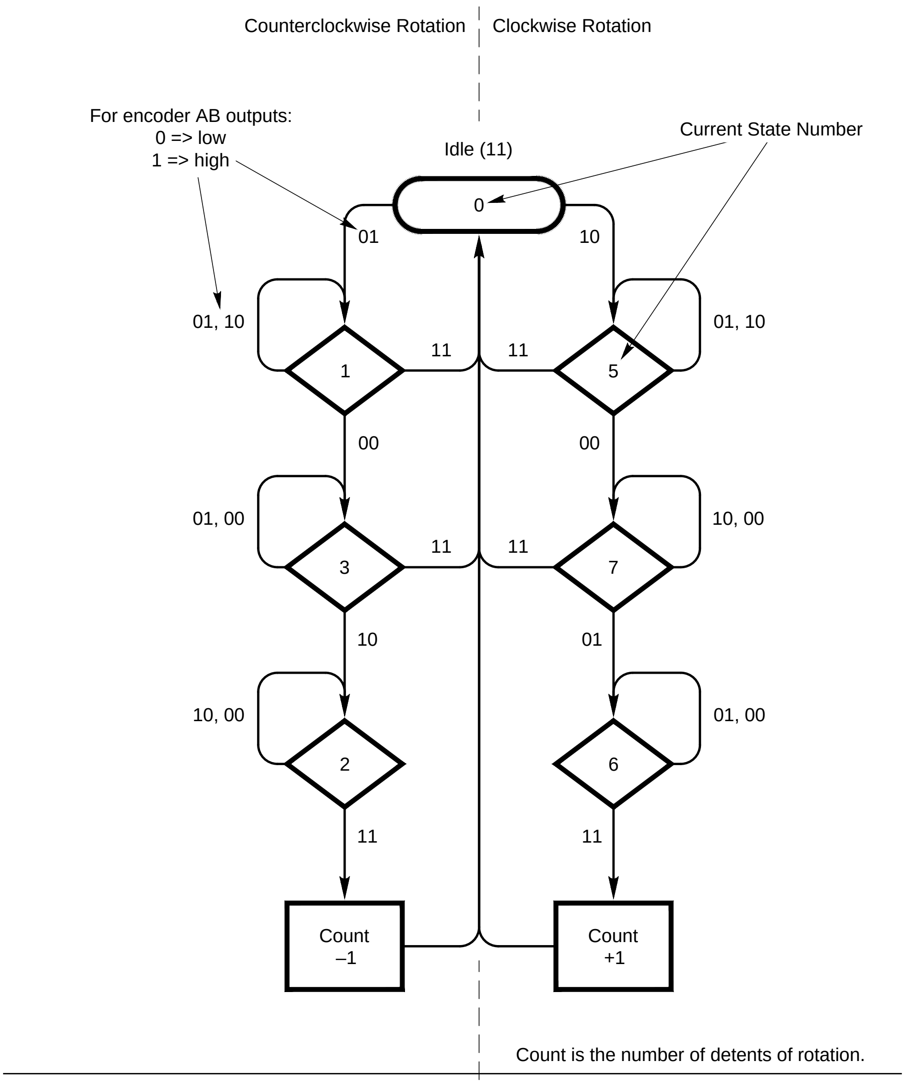

Figure 5 — Encoder state diagram.

Figure 6 — Seven-segment character font set.

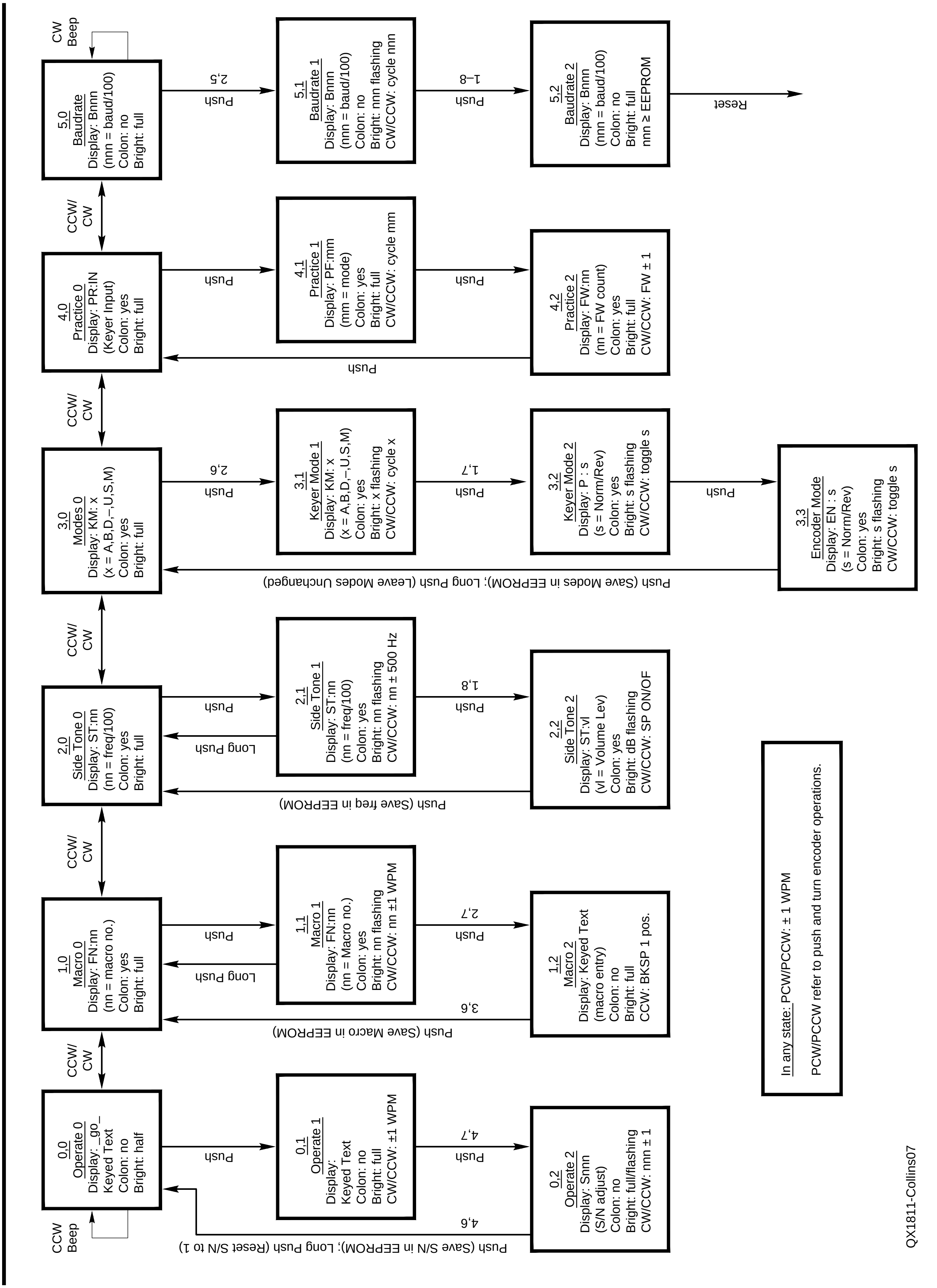

Figure 7 — Ultimate keyer control state diagram.

below the pins. The pins are then soldered to the pads. If the pins are not long enough to reach the pads, small pieces of solid wire are used to bridge the gap. See the images in the **arrl.org/QEXfiles** web page.

Component Assembly Views

Additional images on the **arrl.org/QEXfiles** web page show more detail of the through-hole and SMD component mountings.

Display Fonts

Before getting into describing the ultimate keyer operation, a few words about the LED display are in order. This display is normally used in clocks to show time with its four seven-segment characters. Seven segment displays were designed to display only decimal digits '0' through '9'. Later the fonts were expanded to also display the additional six characters needed to display hexadecimal numbers (A, b, C, d, E and F). I've gone further and defined seven-segment fonts for the remaining letters and several punctuation marks. Figure 6 illustrates all the decimal digits, all the letters, and a few punctuation marks. The symbol '/' is used to represent WPM (W/M) when adjusting keyer speed.

The center colon is used in some displays following an abbreviation for a setting and preceding the current value for that setting. An underscore '_' is occasionally used to represent a space. Keyed text is shown as it is being transmitted by horizontally scrolling each character onto the display from right to left.

Operation

The rotary encoder and its integral push button switch are used to setup and direct keyer operation. At this time the keyer may be in one of 19 control states. Think of an array of six columns each with three rows, except one has four rows, see Figure 7. Each point in the array represents one of the keyer's 19 control states. States are identified by name and by position in the array. The position is defined by the column number (0 through 5) and the row number (0 through 2 or 3) separated by a comma. For example, the upper left corner position is labeled '0,0', the lower right corner position is labeled '5,2'.

Referring to the control state diagram, the keyer firmware always starts in control state 0,0 (first column, first row). In the first row (0), rotating the control knob will move between columns. When in this row, the display is dimmed slightly. Pressing and releasing the control knob moves down the columns, one row for each press and release. In any row except row zero, turning the control knob will generally change a parameter such as side-tone frequency or keyer mode. Turning the control knob will change a parameter. The current value of the parameter is displayed, flashing, on the LED display. Pressing and releasing the control knob will confirm and capture a new setting. When the last (bottom) row in a column is reached, pressing and releasing the control knob will return to the top row (0).

At any time, in any control state, pressing the control knob and turning it at the same time will adjust the keyer speed (WPM). The control state row is not changed when the control knob is released if the control knob has been turned to change the keyer speed.

Table 2 briefly describes operations available in various control states. In the display settings, underlined characters will be flashing. The displays for row zero are shown in parentheses after the column name.

Conclusion

Currently my ultimate keyer meets the basic set of requirements listed at the start. It took me very little time to get used to the unorthodox font displayed on the seven-

Table 2.
Operations available in various control states.

COLUMN 0: OPERATE (_go_)	
0,1 Display: keyed text	Rotation: adjusts keyer speed
0,2 Display: Snnn	Rotation: adjust QSO serial number (nnn)
COLUMN 1: MACRO ENTRY (FN:)	
1,1 Display: FN:nn	Rotation: select macro as follows: 1 - 7, MC (My Call), or TC (Their Call)
1,2 Display: Keyed macro text	Rotation: backspace
COLUMN 2: SIDE TONE SETTINGS (ST:nn)	
2,1 Display: ST:nn	Rotation: adjust side tone frequency (nn x 100 Hz)
2,2 Display: SP:aa	Rotation: turn keyer speaker on (aa = ON) and off (aa = OF)
COLUMN 3: MODES (KM:tr)	
3,1 Display KM:tr	Rotation: change keyer mode (t) as follows:
A	Iambic A
B	Iambic B
D	Dot preferred
-	Dash preferred
U	Ultimatic
S	Semi-automatic (bug mode)
M	Manual (straight key)
3,2 Display KM:tr	Rotation: change paddle mode (r) as follows: Normal = n, Reversed = r
3,3 Display EN: a	Rotation: change encoder mode (Normal = n, A and B Reversed = r)
COLUMN 4: PRACTICE MODE (PR:IN)	
4,0 Display: keyed text ('_' = space)	Rotation: Change state column
4,1 Display: PR:an	Rotation: select text mode
4,2 Display: FW:nn	Rotation: adjust keyer Farnsworth count In this state, the keyer will generate Morse code in five-letter groups at the current keyer rate. Character selection depends on the text mode selected.
COLUMN 5: BAUDRATE (Bnnn)	
5,1 Display: (Bnnn)	Rotation: change serial port baud rate (nnn = baud rate ÷ 100 BPS)
5,2	New baud rate is captured and saved in EEPROM and firmware restarts.

segment displays. The single control works well and I find its use to be very intuitive. The final configuration is small (3” by 4.5” by 1.25”) and weighs less than 5 ounces. The small size makes it easy to press the push buttons using only one hand by pressing on the back of the enclosure with a couple fingers while pressing the button with the thumb. The tactile feel of the push button switches gives a sold feel and feedback when a switch closes and opens. When idle, the ultimate keyer draws about 25 mA. With key down, it draws about 40 mA, plus about 10 mA if the speaker is on.

I find that using something like a heavy key-paddle is inconvenient for field use. Such a key- paddle generally weighs more than the keyer and low-power rig combined. Therefore, I have added a touch key to the nice-to-have list, and have been working on a simple design to mount it on the ultimate keyer enclosure.

The Midnight Ultimate Keyer is available in kit form[2] along with detailed user documentation and assembly instructions. The kit including all components, PCB front and rear panels, and the enclosure are available for $34 plus shipping.

I am now working on the nice-to-have list. In anticipation of required firmware changes, I have included a boot loader that can be used to update the firmware from an ASCII terminal or a PC running a terminal emulator. When I complete the nice-to-have list and, if the interest is there, I will make it the topic of a future, follow on article.

David M Collins, AD7JT, was born and raised in St. Paul Minnesota. He was first licensed as WNØLSH in 1952 at the age of 13, and upgraded to General class as WØLSH a few months later. David graduated in 1961 from the University of Minnesota with a Bachelor of Physics degree. He worked for nearly 45 years in the computer industry as a hardware and software design and engineering manager. Dave retired in 2005 and moved with his wife to Arizona where he was relicensed as Amateur Extra class, AD7JT in 2006. Dave earned the WAS and DXCC awards despite living in a HOA community that forbids Amateur Radio antennas. Since retiring, Dave has worked on several Amateur Radio projects based on microcontrollers including: digital modem enhancements such as a real-time clock-calendar, CW transmit mode keyboard entry, CW receive and decode mode (presented at the 2011 Digital Communications Conference), CW keyer (paddle entry), operation without a keyboard using only paddles for input. He also worked on QVGA graphic display drivers and interface functions, and scalar network analyzer with attachments such as a return loss bridge for antenna testing and tuning, crystal test fixture for grading quartz crystals for crystal filters, a frequency counter, reactance meter, a measurement receiver, and audio frequency network analysis. Further projects include a GPS satellite display terminal, GPS-based precision clock and frequency standard, CW touch keys and various stand alone keyers.

Notes

[1]Microchip Technology Inc., **www.microchip.com/ParamChartSearch/chart.aspx?branchID=8181**.

[2]The Ultimate Keyer is available in kit form from Midnight Design Solutions, **midnightdesignsolutions.com/muk/index.html**.

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VoIP Board Redux

Here are additional PIC programs for the VoIP board described in March/April 2012.

Internet linked Amateur Radios are abundant these days, but fundamental knowledge of their underlying internet technology is not so well known. This article provides some very basic examples of that technology for folks who aspire to understand it and perhaps create their own. The March/April 2012 *QEX* issue[1] included a basic internet VoIP board, with a Lantronix XPORT module and a PIC chip, and two of those boards provide a real-time full duplex, internet audio link.

This article uses one the same boards, with different PIC software. Four supplemental PIC programs are described here, to remotely operate that board using an ordinary web browser as a remote control panel. The intention here is to provide bare bones examples of how this can be done, with the greatest simplicity and clarity.

The reader is referred to the referenced *QEX* article, included here in **www.arrl.org/qexfiles** web page for details about the construction of the board itself, and background information about IP addresses and port forwarding, to connect the board to the internet. This article addresses only the supplemental PIC programs, to show how that same board can be employed for other internet remote control purposes.

Description

Three of the PIC programs described here can directly drive an ordinary web browser, (like Internet Explorer) and two of them also provide browser display and/or control of I/O pins on the PIC chip. The last program can directly drive a Google Earth display to generate a direction finding (DF) bearing line, painted onto a satellite image of Earth. It can therefore provide the foundation for a remote DF display.

The Lantronix XPORT module handles all the complicated internet stuff. It must be configured to use TCP messages — a requirement of HTTP — to use these new programs. The PIC assembly programs described here include the files shown in Table 1.

For brevity, just enough information is provided here to explain the operation of the programs, but references to further reading — suitable for beginners — are provided at the end of the article and in Table 2. A narrative description of the programs is provided for anyone who wishes to accomplish the same thing with alternative technologies such as Arduino, Raspberry Pi, etc.

The BareBones.asm Program

The BareBones.asm server program enables the VoIP board to behave like an extremely simple HTTP file server. It enables the VoIP board to generate a regular website page, which can be directly viewed with an ordinary web browser such as Internet Explorer. This simple webpage consists of just a single static line of text, see Figure 1.

The IP address and port number of the XPORT module must be entered into the browser address window, to access this website. For example, if the XPORT module address is set to 192.168.1.65 and its port is set to 12345, then the entry in the browser address window would be:

http://192.168.1.65:12345

The communication begins when the browser opens a TCP connection with the server. Once a TCP connection is established,

Table 1

PIC assembly programs.

BareBones.asm	“bare bones” HTTP file server program
BBS_2.asm	“bare bones” HTTP file server, version 2
BBS_3.asm	“bare bones” HTTP file server, version 3
BBDS.asm	“bare bones” HTTP file server for a Google Earth DF display

Table 2

Sources of information.

HTTP:	**www.jmarshall.com/easy/http/**
HTML:	**www.jmarshall.com/easy/html/**
KML:	**https://developers.google.com/kml/documentation/kml_tut**

Figure 1 — Text sent by the BareBones WebServer.

all the subsequent browser/server messages are ordinary text messages — sent over the TCP link — compliant with the HTTP specs. The browser then sends an HTTP GET request with, among other things, a filename to identify the particular file the browser wants to get. The server program then fetches the requested file, and sends it back to the browser, with some HTTP reply messages prefixed to the actual file contents.

The HTTP server in this PIC program is extremely simple. A proper server program would parse and examine the entire HTTP text message sent by the browser to identify the type of response expected by the browser.

The BareBones server described here does not bother to parse or interpret the browser request message. It assumes that any message that arrives at that particular IP address and port must be an HTTP GET request, for the one and only file that is available from this particular server. Therefore, the BareBones server program needs only to look for the specific string of characters that signals the end of the browser GET request message. That string can be either a double CR/LF character sequence, (such as CR / LF / CR / LF) or a double LF character sequence (such as LF / LF).

Once either of these character strings sent by the browser is detected, the BareBones server sends a static HTTP reply back to the browser, containing the very simple 84-byte long HTML file, which is hard-coded directly into the PIC chip's program memory. The browser receives this reply, interprets it and displays it.

The BBS_2.asm Program

The BBS_2.asm program generates a dynamic web page, in which some of the HTML file content changes depending on some sort of circumstances that exist at the server. For the BBS_2.asm program, a second line of text is displayed, which indicates with a 2 digit readout how many times the web page has been "served" by the PIC since dc power was turned on. Hitting the REFRESH button on the browser control panel, or F5 on the keyboard, will cause the counter value to increment each time the page is refreshed, see Figure 2. The counter "rolls over" to zero when it reaches 100.

Because the HTML file content changes each time it is served to a browser, it cannot be stored in the read-only PIC program memory. The original file, which serves as a template, is permanently stored in program memory. When dc power is first applied, the PIC software makes a copy of this file into the RAM memory, so that some of its content — the two characters for the refresh counter — can be changed dynamically, as the program runs.

The intention here is to show that the browser display can be used to indicate some sort of dynamic conditions that exist at the server, for example, things like input pin states or ADC readings, by using those conditions to dynamically change the content of the HTML reply sent back to the browser.

Please note the browser display will not automatically update whenever a change occurs at the server, because this server, and all other HTTP servers, can only respond to requests sent by the browser. To achieve a near real-time display that automatically refreshes, a suitable workaround solution would be to include an HTML REFRESH command in the reply file contents:

<meta http-equiv="refresh" content="5">

This command must be located in the HTML file between the <head> and </head> tags. The example shown here will cause the browser to automatically REFRESH once every five seconds. Adding this line will change the size of the HTML file, so the value specified in the CONTENT-LENGTH line of the HTTP header must be adjusted to reflect the increased file size. It will also change the RAM addresses of the two characters that express the refresh counter value.

The BBS_3.asm Program

The BBS_3.asm program shows how a browser can be used to send commands to the VoIP module. In the BBS_3.asm program, this consists of commands to turn on/off two utility LEDs on the VoIP module. An indication of the current state of these two LEDs is also provided in the browser display (see Figure 3), and the browser display is refreshed/updated automatically whenever a command is sent.

The commands are sent using HTML HyperLinks in the browser display and HTML FORM SUBMIT methods, in which the command is appended — as a short suffix — to the filename that is GET requested from the server whenever a HyperLink is clicked in the browser window. The suffix

Figure 2 — Refresh counter reset.

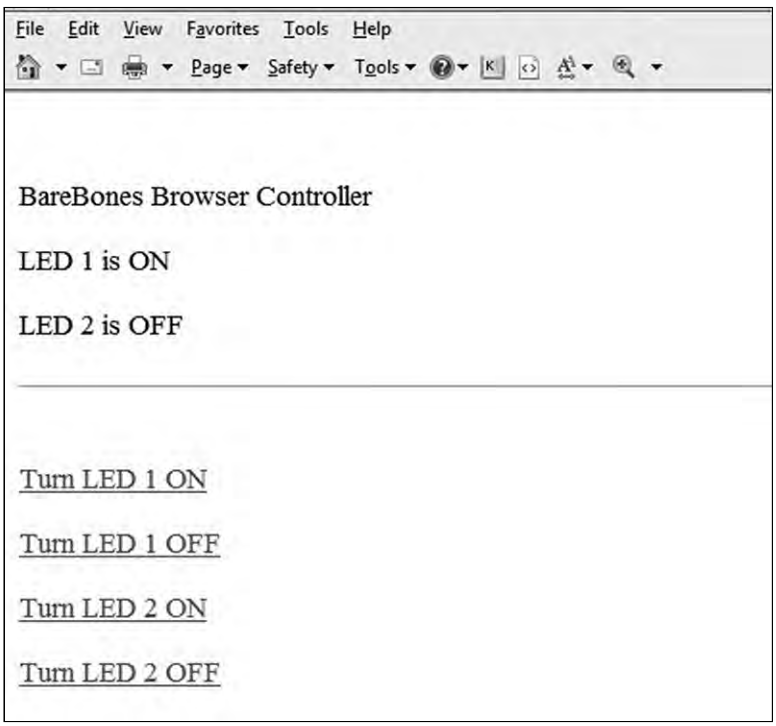

Figure 3 — Turning LED 1 and LED 2 on and off.

consists of a question mark, followed by the command string. The question mark is an HTML-compatible delimiter character that separates the requested filename from the command string.

This new feature requires a more elaborate server routine in the PIC, since the browser request must now be parsed and interpreted by the PIC chip to identify the browser command. Even so, it is still assumed in the server routine that any message arriving at that IP address and port is an HTTP GET request, so an HTTP reply will be sent to the browser, after the browser command is identified and executed.

The BBDS.asm Program

The BBDS.asm program replaces the HTML/text file with a KML file, which can directly drive a Google Earth display with a simulated output from a remote direction finding (DF) base station. It will generate a station icon on the Google Earth display, as well as a 10-mile long DF bearing line extending away from the icon. Figure 4 shows the true bearing of 315° to my station from the station location of W1AW.

Figure 4 — The BBDS.asm program can directly drive a Google Earth display, including a simulated output from a remote direction finding base station. Here the station located is W1AW. [Google Earth image.]

This is a static display that never changes, and is provided to show how a Google Earth display can be remotely driven by the VoIP module. The background satellite image, and the station icon image, are fetched directly from the Google Earth servers. The main information provided by VoIP board is the location information required to paint the station icon and the DF bearing line onto the Google Earth display.

The Google Earth display must be configured to fetch the information from the VoIP module, by creating a NETWORK LINK in the ADD menu of Google Earth. Network links are similar to bookmarks in browsers. The network link must specify the IP address and port of the VoIP module, using the same syntax as the other examples provided in this article. The NETWORK LINK can also be set in Google Earth to automatically refresh at regular time intervals, allowing the creation of a DF display that provides near real-time readings of the DF bearing.

More About HTML and KML Files

If you want to learn about HTML and KML, the only tools required are a text editor like NotePad or WordPad to create and edit the files, and a web browser to display HTML files and/or a copy of Google Earth or Google Maps, to display KML files.

The files can be created, stored and tested on the same computer that runs the browser and Google Earth programs. It isn't necessary to fetch the files through an internet connection. Once the files have been created and tested locally, they can be hard-coded into the PIC chip for installation on the VoIP board. Remember to count all the characters in the files, including <CR> and <LF> characters, because that number must be expressed in the HTTP Content-Length statement.

When creating the files, they must be saved as plain unformatted text files, and the filename extensions must then be changed to a suitable suffix to trigger the proper display program when the files icons are double-clicked. For example, an HTML file created with a text editor and saved with filename MyFile.txt will trigger NotePad or WordPad whenever its icon is double-clicked. But if the filename extension is changed to "htm" such as MyFile.htm, then double-clicking the same icon will open a browser to display the file contents. Creating a KML text file and changing the filename extension to "kml" such as MyFile.kml, will trigger Google Earth or Google Maps to display the file, whenever its icon is double-clicked.

Filename extensions can be changed by right-clicking the file icon and selecting RENAME from the pop-up list that appears. Move the cursor over to the filename extension to the right of the period, and change it. To edit HTML and KML files with a text editor, right-click on the file icon and select "OPEN WITH" from the pop-up list that appears. Another window will appear, with a list of other programs that can be used to open the file, including a text editor. You can add a text editor to this list, if one is not shown.

The HTML and KML files are included in the **www.arrl.org/qexfile** web page package. Two copies of each file are provided, one as an ordinary 'txt' file and another as an 'htm' or 'kml' file, but the actual file contents are identical. These files do not contain the HTTP prefix messages that are also sent. The prefix messages can be found in the PIC '.asm' code.

Closing Remarks

The examples provided here are very basic, but more elaborate browser control panels can be achieved by adding JavaScript to the webpage served by the VoIP board. The PIC chip has several unused pins that could be employed to drive external circuits. This would involve significant additional PIC code, beyond the scope and intent of this article.

The VoIP board provides a good learning tool for internet technology, since all the complicated UDP / TCP "stuff" is handled in the XPORT module, which simplifies things a great deal. For those seeking a deeper understanding, further reading about UDP and TCP messages is worthwhile, since they comprise more than 95% of the messages actually traversing the internet.

A free software such as WireShark (**https://www.wireshark.org/**) is also recommended. This is a network protocol analyzer. It is a utility program that allows you to observe all the message traffic on a computer network, down to the byte level. Essentially, it is an oscilloscope for observing computer network messages. Additional beginner's information on HTTP, HTML and KML can be found in Table 2.

An excellent book on the topic of KML is *The KML Handbook: Geographic Visualization for the Web*, by Josie Wernecke, one of the principle engineers at Google who helped document the KML language.

Bob Simmons, WB6EYV, was first licensed as a Novice in 1964 at age 13, and remained licensed (more or less) constantly ever since. He also earned an FCC Commercial Operator license in 1967. Bob served 6 years of Naval Reserve duty as a radar technician (ETR2) with about 6 months of total sea time. He spent several years of civilian work in nautical and marine electronics in Los Angeles harbor, as well as doing some land mobile radio work, followed by 5 years in flight line avionics, working on business jets. He moved to Santa Barbara, CA in 1992 and to Santa Maria, CA in 2015. Bob is retired, but works part-time for a defense subcontractor. Bob maintains a website ***www.picodopp.com*** *dealing with radio direction finding equipment, including technologies to enable internet-linked remote DF stations. His primary interest is applying new technologies to old problems, and encouraging homebrew activities.*

Notes

[1]R. Simmons, "A Simple Internet VoIP Board", *QEX*, March/April, 2012, pp. 20-24. See **www.arrl.org/qexfiles**.

Letters

Amplifier Overshoot-Drive Protection (Sept./Oct. 2018)

Dear Editor,

There are a couple of potential risks with the Amplifier Overshoot-Drive Protection described by Phil Salas, AD5X. Gas Discharge Tubes (GDTs) were designed for protecting power and telephone lines against intermittent transients — particularly lightning or switching transients. They later found use protecting video amplifiers in CRT displays against occasional internal CRT arcs. These types of transients occur "every once in a while," maybe a dozen times a year.

The first risk is the lifetime of the GDT. Each arc slightly degrades the GDT electrode coatings, which slowly raises the arc-over voltage. Littelfuse GDTs are typically rated for a surge life of 10 to 300 arcs under standardized test conditions — typically with a cooling off period between each arc. While I cannot predict how long the GDT will last as a limiter in Amateur Radio applications — I'd assume far more than 300 arcs — they will eventually stop working. They are just not meant for this sort of continuous use.

The second risk is damage to the exciter. A GDT is pretty much a dead short circuit when it arcs. This means the transmitter is suddenly seeing a very low output impedance equal to GDT series resistor in parallel with the amplifier input impedance. Let's say the amplifier is a good 50 Ω, then the IC706 momentarily sees a load impedance of about 12 Ω, or an SWR of 4:1. Even though it occurs only for a short time, this is added stress to the Icom IC706MKII RF. — *73, John Perlick, KØUM*, **jperlick@ariacorp.com**

The Author Replies,

Good comments. The GDT I used has a lifetime spec of 400 shots at 500 A. Assuming the radio does put out its full 100 W into the 16 Ω load when the GDT fires, you're looking at about 2.5 to 3 A. So the question is how does this relate to improved GDT life? I found one reference when I was first experimenting with this that implied that the lifetime varied close to square-law with current. That is, at 1000 A you could only get 100 shots. Assuming that this is true, the 400 shots should be more like 50,000 shots at 3 A. However, I have not been able to find lifetime info at the sub-5 A currents you'd see here. So I'm just making the assumption that the square-law lifetime determination follows at low current as well as high current.

One reference, **nextek.com/wp-content/uploads/2015/01/Technical_Note_-_NexTek_Arrestors_-_Gas_Discharge_Tube_Maintenance_v3.pdf**, shows that going from 50 kA to 1 kA (50 to 1) increases the number of shots by a factor of 1000. But again, information at 3 A is not available. Hopefully some of our *QEX* readers may have more information on this that they can share with us.

Keep in mind that this is not a continuous operation state. The radio's ALC does control the power after the first 2 ms of the first 'dit' of a transmission and it holds the power for 5 to 6 seconds after the last 'dit' of the transmission. So normally you would see this overshoot once per transmission during a QSO. Once you pause more than 5 or 6 seconds, there will be another overshoot — at least with the IC-706MKII.

I believe that there is minimal stress on the exciter's power amplifier. Higher current causes thermal stress in solid-state devices, which can result in failure. While the SWR due to the low 16 Ω impedance results in higher current in the finals, this lasts for less than 2 ms. So there is virtually no thermal stress that occurs in the devices, just like there is little heating in the 16 Ω resistor during the 2 ms period of the overshoot. I would be more concerned if the SWR was due to high voltage since high voltage breakdown can be a problem even with very short time durations. — *Phil Salas, AD5X*.

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

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Wide Dynamic Range Field Strength Meter

This portable 90-dB dynamic range RF field strength meter is simple to build.

While reviewing the specification sheet for the venerable AD8307 logarithmic RF detector IC, I noticed that the output circuit type shown in a block diagram is a current source. It occurred to me that this feature could be easily utilized to drive a conventional analog meter directly. Although I enjoy building gear that utilizes modern day microcontrollers and wireless devices my thoughts became focused on how simple it would be to build a wide dynamic range field strength meter (FSM) by mixing recent and legacy technology. This article describes a portable 90-dB dynamic range RF FSM that is simple to build. In comparison to the traditional basic diode detector type FSM, this minimal component count instrument can be used to measure field strengths over a much wider dynamic range and with much better accuracy.

The Analog Devices AD8307 logarithmic detector used in this design has been available for a few decades and has appeared in many Amateur Radio projects in the past. This well-proven detector has been typically used in microcontroller-based RF power measurement applications where the output level of the detector is digitized and the resulting measurement displayed on an LCD screen. The simplicity of the design that is described here stems from the use of a legacy d'Arsonval panel meter that can be directly driven by the AD8307 without the need for any microcontroller or any additional active component circuitry.

The use of a legacy analog panel meter might seem somewhat archaic in our modern digital era. However, in my opinion, the analog meter greatly simplifies the design, is sufficiently accurate for a FSM, and perhaps most importantly, uses circuitry free of RF EMI. This factor is significant as during normal use this broadband sensitive instrument is very often in close proximity to the sense antenna. Thus an EMI-noisy digital design could result in sensitivity-limiting residual readings. Additionally, analog meters can be read at a glance and are easy to view in bright sunlight — a typical field measurement environment. Finally, while adjusting equipment to obtain maximum or minimum signal strength, analog meters provide a much easier to use peak or null signal strength indicator than a digital display.

Design Concept

The block diagram provided within the AD8307 spec sheet indicates the output circuit consists of a current source mirror providing an output current that varies at a rate of 2 μA per dB of input signal level change. This output current is usually passed through an internal 12.5 kΩ load resistor. The 2 μA current change within this load resistor results in an output voltage variation of 25 mV/dB. In more typical designs this voltage level is digitized via an ADC analog input of a microcontroller to permit digital processing. The different approach used here feeds the 2 μA/dB current source output directly to a low resistance 200 μA panel meter. The output voltage of the detector is shunted to near ground by the low resistance meter, which in this circuit design is connected in parallel with the internal load resistor. The detector 2 μA/dB output drive that is fed through the 200 μA meter (which has a 0-to-10 scale) results in a scale calibration of 10 dB per major division. Since I already had a good quality Weston 50 μA meter on hand that conveniently had both 0-to-10 and ± 5 zero-center scales, I adopted it for use here. I reduced the meter sensitivity to 200 μA by adding a suitable shunt resistance. The AD8307 and 200 μA panel meter along with a few surrounding passive components are all that are needed to build a reasonably accurate, wide dynamic range FSM.

Expanding the Scale

After experimenting with the prototype proof-of-concept basic circuit and confirming that it worked as expected, I decided that a 10:1 scale expander capability would be beneficial. The 90-dB wide dynamic range scale is very desirable for viewing the large variations expected from antenna pattern front-to-side lobe, front-to-back ratios, and for quick checks of transmitter ERP levels. However such a wide dynamic range scale doesn't lend itself well for viewing small level changes.

When using the 100 dB analog scale a 1 dB change corresponds to a shift of only about a needle width. In some cases 1 dB is not significant, for example if doing an antenna front to back test where the ratio might be around 25 dB then a dB or two, one way or the other, just doesn't matter much. But if tweaking an antenna for maximum gain or an RF amplifier for maximum output, +1 dB does matter as it represents about a 25% power change. To remedy this deficiency I added a simple op-amp add-on circuit with a gain of 10 that uses a potentiometer for setting a zero center reference. This provides a 10:1 scale expansion for performing relative power tests. When activated this circuit expands the scale to 1 dB per major division, which provides a good fine tuning indicator. The measurement needle width resolution becomes roughly a tenth of a dB — fine enough for most field strength measurement needs.

Instrument Power

This instrument can be powered from either four internal AA cells or from an external 5 to 6 V source. Power supply design for the basic FSM circuit is simplified because the AD8307 output current calibration is not, within fairly broad limits, sensitive to supply voltage variations. A bench test confirmed that the supply voltage can be varied from 4 to 6 V without a noticeable change in detector output current for a given RF input level. There is no need for a voltage regulator. Coincidentally, the normal discharge curve for a 4-cell alkaline battery from fresh to end-of-life exhibits a voltage decline from 6 down to 4 V. A battery saving momentary push button power switch, intended for making short duration measurements, ensures the internal batteries are not left under load for extended periods. A switch and multiplier resistor are provided to also permit the use of the panel meter as a voltmeter with a 10 V scale for checking the power source voltage level.

For longer term RF level measurements external power can be supplied from any clean 5 to 6 V dc power source. For portability a suitable adapting USB cable would allow the use of a rechargeable lithium 5 V battery bank similar to those intended for cell phone power backup. The load is less than 20 mA, so a typical lithium pack should provide power for several days of continuous operation on a single charge. Not all USB battery banks are suitable for this application. Some designs emit switching-regulator EMI and/or detect this very light load as a disconnection resulting in auto shut-off after a short delay.

Circuit Description

Figure 1 is the schematic for the basic version without the scale expansion add on. For applications where scale expansion is not important this basic circuit is all that you need. When equipped with a 1:16 impedance ratio input transformer it will provide a reasonably accurate RF signal power measurement ranging from -80 to +10 dBm. Each major division on the 1-to-10 division scale represents a 10 dB change. Figure 2 illustrates the conversion chart of a voltage range from roughly 70 µV to 700 mV rms within a typical 50 Ω impedance circuit to dBm.

Most popular AD8307 power meter circuits use a simple unbalanced input design rather than harnessing the differential balanced input capability of this device. Commonly the simpler unbalanced design is obtained by shunting one input to RF ground with a bypass capacitor and by connecting the other input via a dc-isolation coupling capacitor to a 51 Ω input termination resistor.

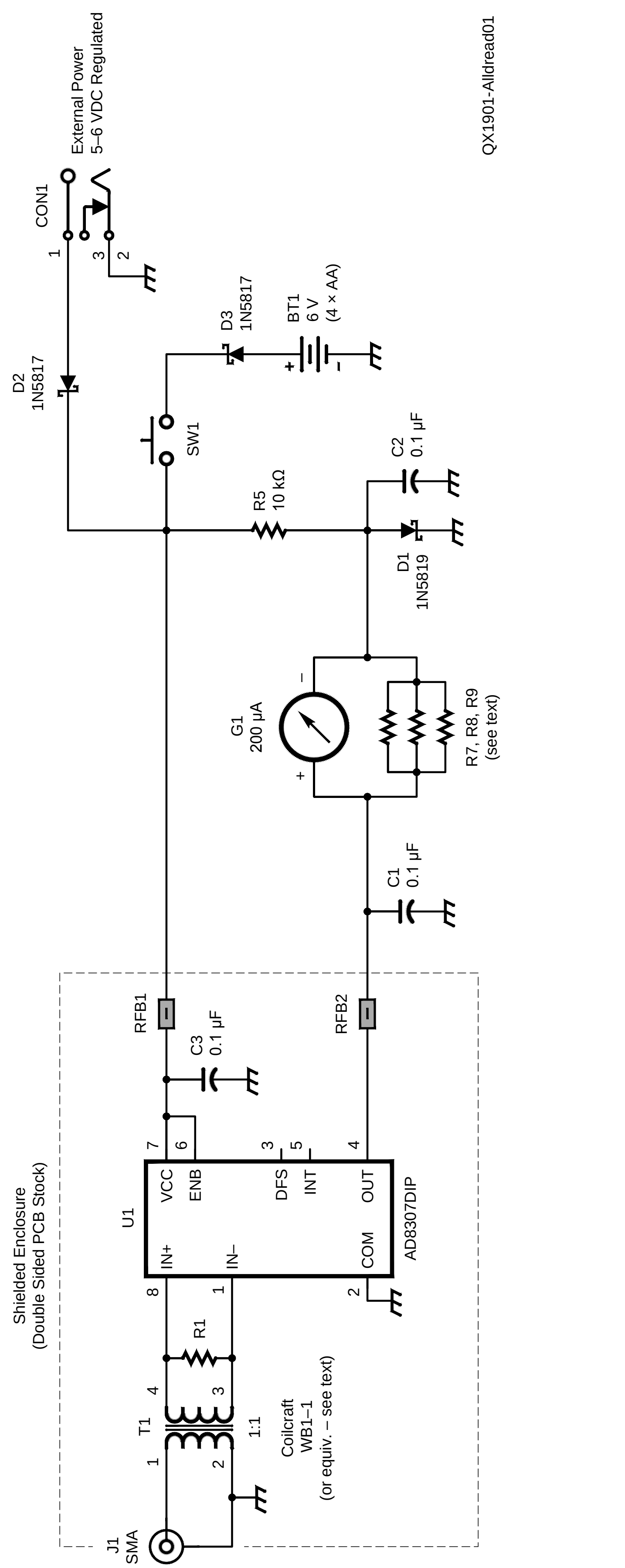

Figure 1 — Basic wide dynamic range field strength meter schematic.

This provides a simple broadband 50 Ω unbalanced input. For this application a transformer-coupled differential input offers some advantages. Transformer coupling that uses the balanced input to the IC offers improved common mode noise rejection, low frequency power line related noise isolation, input circuit protection from extraneous voltages and the opportunity to improve the sensitivity without the need for additional active components. The sensitivity improvement is obtained by more closely matching the detector differential input impedance of 1.1 kΩ to 50 Ω, by using a broadband step-up RF input transformer.

The instrument sensitivity is directly proportional to the impedance ratio of the wide band input transformer (T1 in Figure 1). The design trade-off for using a high step-up impedance ratio transformer is loss of high frequency response. This design uses a 1:16 impedance step up ratio. The frequency response (Figure 3) is flat within 1 dB across the LF and HF spectrum segments then rolls off 2 dB at 50 MHz and about 6 dB at 150 MHz. Although the roll-off is significant across the lower half of the VHF spectrum the instrument is still quite usable up to 150 MHz. Of course a calibration factor must be added if performing VHF absolute power measurements.

The 1:16 transformer steps up the 50 Ω input source impedance to 800 Ω. Compared to a transformer-less input design, this improves the instrument voltage sensitivity by a factor of 4, or 12 dB. The value of resistor R1 in Figure 1 depends on the transformer ratio. For example: with 1:1 use 51 Ω; with 1:4 use 240 Ω; with 1:6 use 430 Ω, with 1:8 use 620 Ω; and with 1:16 use 3 kΩ. In this case resistor R1 is 3 kΩ, which is connected across the 1.1 kΩ AD8307 differential input impedance and matches the 800 Ω transformer secondary impedance. This input circuit conveniently provides a close 0-10 scale meter calibration alignment that can be directly related to -90 to +10 dBm absolute power. Note that use of the -90 to -80 dBm bottom 10% region of the meter scale (0-to-1 area) is not recommended for obtaining accurate measurements, since that low level region is influenced by the AD8307 detector noise floor.

Figure 4 shows a frequency sweep of the input return loss. The input match is good with return loss greater than 14 dB (VSWR less than 1.5:1) up to 30 MHz. Then above 30 MHz the VSWR climbs to 2:1 at 50 MHz and 3.5:1 at 144 MHz. If performing a mismatch sensitive measurement above 30 MHz a 6 dB coaxial attenuator connected directly to the input should be used to improve the VSWR to 1.3:1 or better.

The logarithmic detector circuitry of

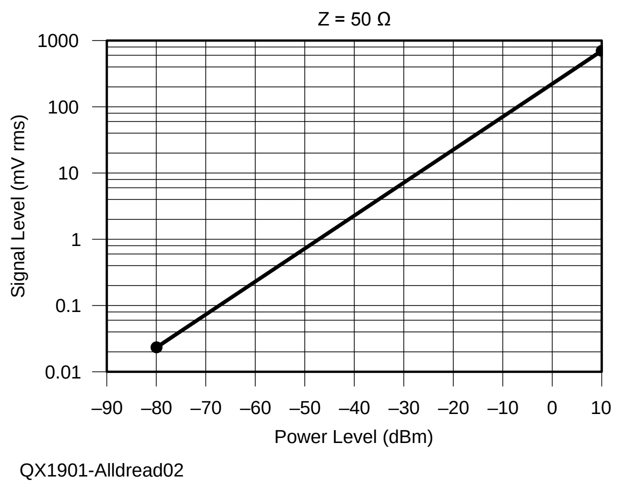

Figure 2 — Meter scale units conversion chart in a 50 Ω environment.

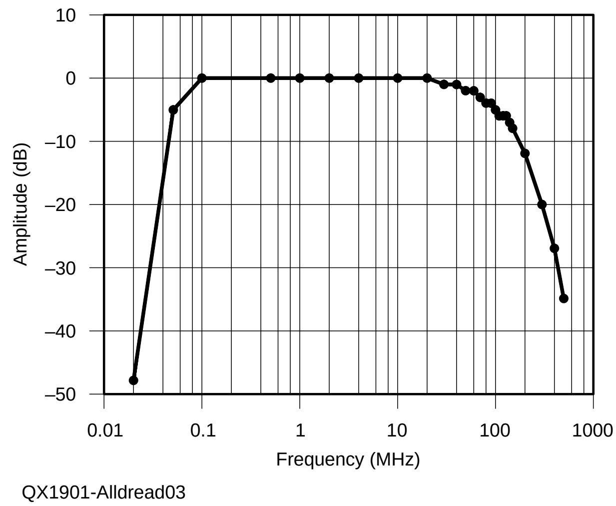

Figure 3 — Field strength meter frequency response.

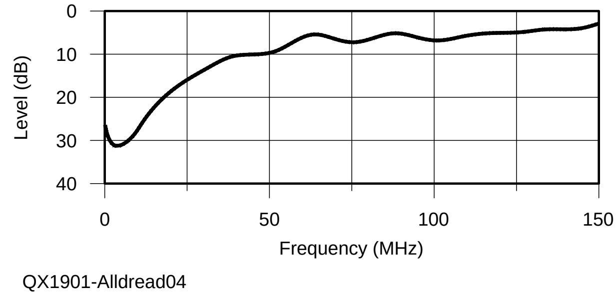

Figure 4 — Field strength meter input return loss frequency response.

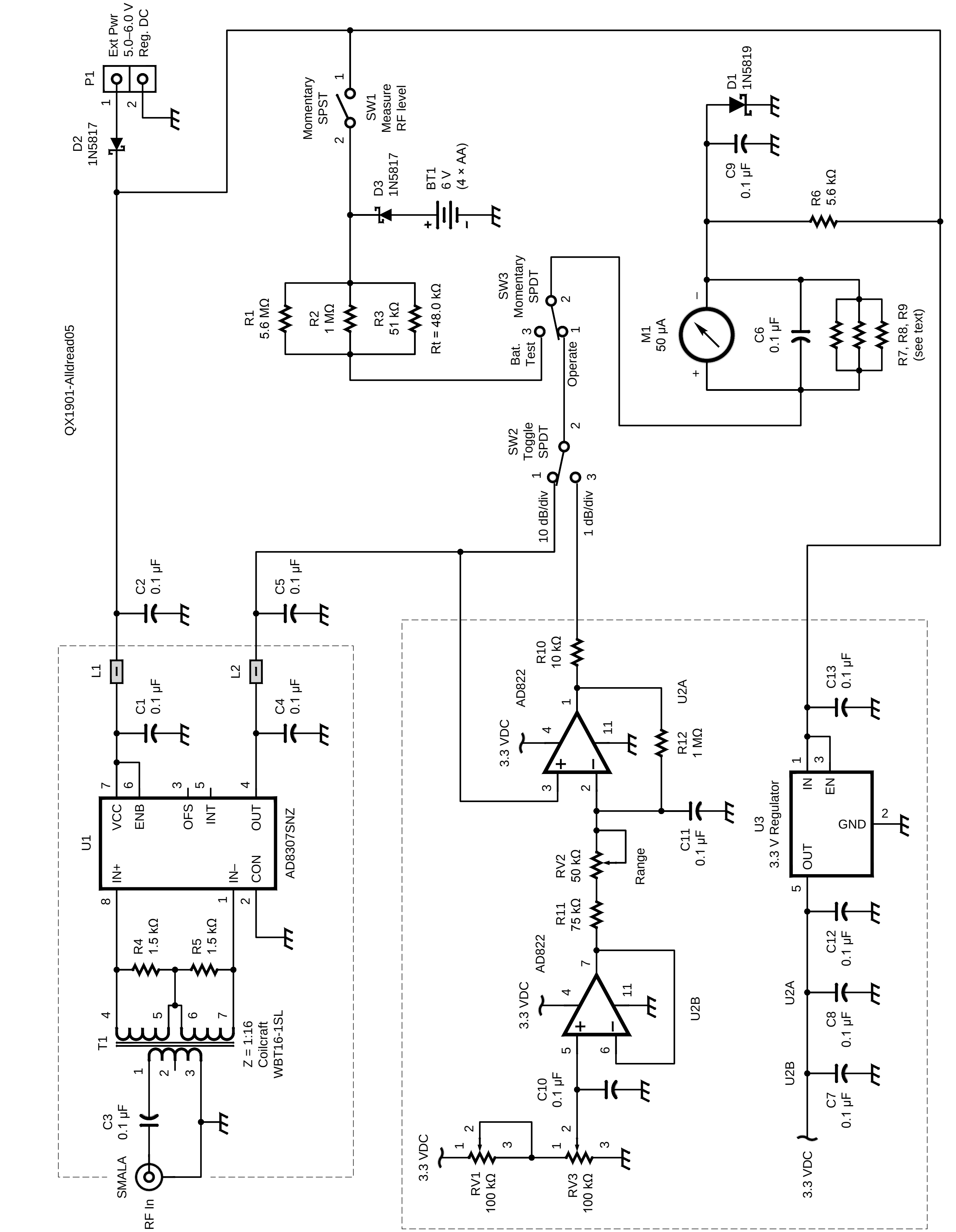

Figure 5 — Field strength meter schematic including the 10:1 scale expansion circuit.

Figure 1 is mounted in a separate RF-tight enclosure. For detailed information on the AD8307 logarithmic detector please refer to the exceptionally good data sheet[1] provided by Analog Devices. Feed-through ferrite beads and bypass capacitors are provided for both dc power source injection and the detector output to prevent any possible unwanted RF EMI ingress into the detector enclosure.

During idle conditions without any RF input there is a residual output voltage of about 0.3 V from the AD8307 detector. This output is generated by the noise floor of the high-gain logarithmic amplifier chain inside the IC. R6, C9 and Schottky diode D1 (Figure 5) within the meter ground return path have been provided here to prevent the residual noise generated output from occupying precious space on the meter scale. The forward voltage drop across D1 provides an offset bias of about 0.4 V. This prevents meter deflection during the idle state, and provides the desired zero meter reading with no RF input. Without this bias the idle noise related output would deflect the meter up to about 1 on the scale of 0 to 10 and the usable dynamic range of the meter scale would be reduced by about 10 dB.

Shunt resistors R7, R8 and R9 reduce the 50 μA meter full scale sensitivity to 200 μA. C6 prevents RF from entering the meter movement coils. SW3 is the momentary battery test switch and the parallel combination of R1, R2 and R3 provide a precise multiplier resistance of 48.0 kΩ, which with this particular meter provides an accurate 10 V scale for checking the source voltage and is a convenient means for checking the battery condition. Diodes D2 and D3 isolate the internal battery and external power sources.

Figure 5 shows the FSM with the scale expansion add-on option. Voltage follower U2B provides impedance isolation for center zero potentiometer RV3. Trimpot RV1 limits the adjustment range of RV3. This circuit minimizes battery drain by permitting the use of high resistance potentiometers for the center zero setting. U2A provides the 10:1 voltage gain needed for scale expansion. U3 provides a regulated 3.3 V power source for this circuit. Unlike the AD8307 this circuit is source voltage sensitive. Switch SW2 provides switching between the 10 dB/div and the 1 dB/div relative power measurement modes.

Parts Procurement

All the parts are standard and available from most major distributors with the exception of the panel meter. If you wish a simplified construction approach that minimizes the need for soldering

Figure 6 — Field strength meter housed in a 4-inch cube junction box.

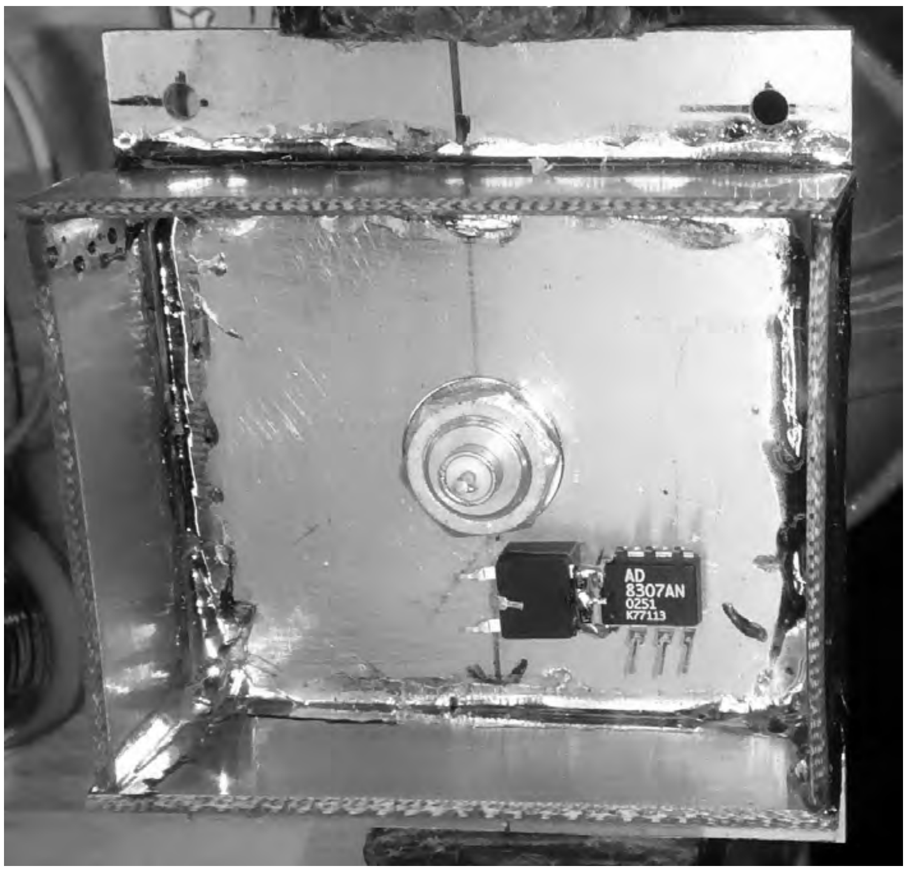

Figure 7 — "Dead bug" mounting style of the AD8307 RF detector, and the 1:16 wideband step-up transformer.

components, and you are willing to trade off some sensitivity for a bandwidth extension benefit to 500 MHz, there are inexpensive assembled AD8307 PCB modules available from online Asian sources.

These modules are complete with an SMA RF input connector and some even have shielded detector circuitry. The cost of the modules is about the same as an AD8307 IC alone from North American suppliers[2]. Using this approach about all that you would need to make a basic wide dynamic range FSM is the AD8307 module, a case, a battery-holder, switches and suitable panel meter along with a few leaded components.

Some distributors have stopped selling mechanical panel meters in favor of modern day digital display alternatives. Others still sell legacy panel meters but at sticker-shock prices, in some cases exceeding $100. However at the time of this writing meters manufactured in Asia are still sold online for less than $10. While browsing[3] eBay I noticed a 4", 1.5% class, 100 μA meter with a usable 0-100 scale that has major scale marks every 10 μA. It could be used here with a suitable shunt to lower the sensitivity to 200 μA — the needed shunt resistance should equal the meter resistance. Surplus equipment commonly used good quality panel meters that could be used for this application.

I used a 4" by 4" by 4" electrical junction box (Figure 6) for the case. These relatively inexpensive high volume production, sturdy, sealed boxes were available from Home Depot stores for about $10 at the time of this writing. This particular box initially had rear wall mount tabs that were easy to snap off. The carrying handle is a $0.50 plastic drawer handle from Home Depot.

Figure 7 shows the dead-bug style mounting of the AD8307 RF detector along with the 1:16 wideband step-up transformer made by CoilCraft. Note the two 0805 size surface mount type 1.5 kΩ resistors (R4, R5 in Figure 5) are soldered in series across the transformer secondary pins. These RF components are mounted within an RF-tight enclosure made from double-sided F10 PCB stock. The transformer is mounted close to the centered SMA connector to minimize the connection length to the primary pins.

I built the 10:1 scale multiplier circuit PCB (not shown) on tenth-inch grid prototype PCB stock. I used 8-pin sockets for the regulator and amplifier ICs, and #30 AWG wire-wrap wire to interconnect the components.

Measurement Procedure

The use of this meter is very intuitive. Simply connect the RF source or sense antenna to the input. When using the wide dynamic range scale, just press the momentary power button switch when the input signal to be measured is present and note the meter reading. Each major division is 10 dB, and 0 on the scale represents -90 dBm. For example, if the meter reading is 3.2 then the signal is 32 dB above -90 dBm, or -90 + 32 = -58 dBm.

If you need to know the approximate level in micro-volts, use the graph in Figure 2, which shows that -58 dBm is about 280 μV rms across 50 Ω impedance.

For reasonably accurate measurements at VHF a 6 dB attenuator should be connected to the meter input to improve the match. Then 6 dB must be added to the measured signal level along with the frequency dependent calibration factor shown in Figure 3.

As previously mentioned, readings that fall within the bottom 10% of the scale (from 0 to 1) are influenced by the noise floor of the AD8307 detector thus the accuracy of this region of the scale is compromised.

More precise relative readings with the 10× scale multiplier installed can be activated with the scale toggle switch. In this mode the meter must first be center-zeroed with the level of the reference signal by adjusting the zero potentiometer for a mid-scale reading. Then the new expanded scale sensitivity becomes 1 dB/div with a measurement range of about ±5 dB. This is very useful for observing small level changes.

Conclusion

This FSM is quite useful for making quick checks of RF field strengths. Although it is not a highly accurate instrument that measures all signal parameters, it is simple and easy to use for quickly checking that ERP levels are normal. With the AD8307 logarithmic detector, the FSM is relatively stable and offers a much wider dynamic range than a simple diode detector FSM. This instrument could possibly be useful for communications teams involved in public service events. It could provide a means just piror to activity commencement for quickly confirming that handheld radios and mobiles are ready to operate at normal ERP levels.

*Tom Alldread, VA7TA, became interested in electronics when still in grade school. In his teens, he repaired radio and television sets, obtained his Amateur Radio license at age 19. He obtained his Commercial Radio Operator certification upon graduating from technical college a year later. Tom subsequently graduated from the Capitol Radio Engineering Institute Engineering Technology program. He worked as a microwave, multiplex and VHF radio equipment maintenance technician, an instructor, an engineering standards and design specialist, and in the Middle East as an adviser for long distance network operations management. Tom is now retired and lives on Vancouver Island with his wife Sylvia, VA7SA. Tom, a member of RAC, enjoys operating CW, designing equipment, and supporting emergency communications. For the past decade he has been the net manager for the SSB/CW 20 m Trans-Canada Net (**www.transcanadanet.com**). He is interested in microcontroller development projects associated with RF technology and Amateur Radio. Tom received second place award in the Luminary 2006 Design Stellaris contest, and first place in the 2011 Renesas RX contest. He is also interested in computing, RVing, hobby farming and bicycling. Tom maintains the **www3.telus.net/ta/** web page.*

Notes

[1] **www.analog.com/media/en/technical-documentation/data-sheets/AD8307.pdf**.

[2] Search "500MHz-RF-Signal-Power-Detector-AD8307-Module-Field-Detection" at **www.ebay.com/**.

[3] Search "DC-0-to-100uA-Class-1-5-Accuracy-Panel-Analog-Ammeter- Ampere-Meter" at **www.ebay.com/**.

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Antenna Analyzer Pet Tricks

Discover a range of measurements you can make with your antenna analyzer.

Antenna analyzers have become a very popular accessory, and many hams have acquired one. Since most of us don't change antennas weekly, the antenna analyzer may not see frequent use. But these instruments can do many other useful things that might have you reaching for one more often. Teach your antenna analyzer a few new tricks and it might become your favorite pet.

An antenna analyzer is really a one-port network analyzer, a powerful RF measurement tool. When professionals need to characterize an RF component, whether it is an antenna, transmission line, or filter, the "go to" piece of test gear is a network analyzer. A VNA (Vector Network Analyzer) is expensive, complex, heavy, and delicate. While not as versatile as an expensive network analyzer, the inexpensive antenna analyzers that have recently appeared on the market can be very useful in characterizing transmission lines, filters, and RF components as well as antennas. Let's look at some measurements that can be made with an antenna analyzer that a microwave engineer would do with an expensive network analyzer.

My antenna analyzer is an inexpensive one from China, shown on the left in Figure 1. I chose it because it covers a wide frequency range, 137.5 to 2,700 MHz, including six VHF, UHF, and microwave bands. The user interface can be charitably described as execrable, but can be decoded by referring to the manual. The major advantage is that it is cheap — cheap enough to drop off a tower without crying. The measurements I am about to describe may be made with any antenna analyzer.

I borrowed another antenna analyzer from my neighbor, Chip Taylor, W1AIM. This one, from the Ukraine, is on the right in Figure 1. The frequency range is from 1 to 1,400 MHz, covering all bands from 160 meters to 1,296 MHz; important since Chip also operates HF. The user interface is better, since the package is larger with more buttons, but the price is much higher – dropping this one would be painful.

When operated at a single frequency, as in Figure 1, both analyzers display several different quantities: VSWR or SWR, Z, R, X, RL or |S11|, and C. All of them are variations on impedance measurements, and most are useful in different contexts, as we shall see. But the single frequency quantities don't provide a lot of insight, and if we are just looking at VSWR at a single frequency, an inexpensive meter will often suffice.

Figure 1 — Handheld antenna analyzers.

VSWR or SWR

The obvious use for an antenna analyzer is to measure antenna VSWR or SWR ([Voltage] Standing Wave Ratio), something hams obsess over. Both analyzers can display VSWR across an entire band, as in Figure 2, rather than at a single frequency like the classic VSWR meter. A quick picture with a cell-phone camera can record the plot for later reference, to be sure the antenna hasn't changed. In New England, rain, ice, and snow can detune an antenna, so we can check the effect. I neglected to check one winter and blew up a solid-state kilowatt amplifier.

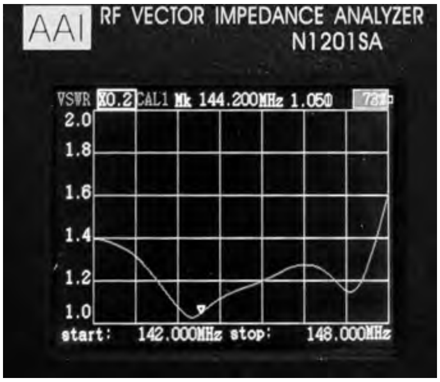
Return Loss (RL) and VSWR are both measures of relative reflected power. Return Loss is the difference between Forward and Reflected power in dB. A dead short (or an open) will reflect all of the forward power — the return loss is 0 dB, or VSWR is infinite. A good antenna might reflect only 1% of the power, so the Return Loss is 20 dB. VSWR and return loss are related. A low return loss (lots of reflected power) indicates a high VSWR. The conversion may be done by calculation, but most antenna analyzers can display both Return Loss and VSWR. My analyzer displays |S11|, the magnitude of S-parameter S_{11}, which in decibels is the negative of Return Loss.

Antenna analyzers are sold as "antenna analyzers" but they do lots more. All the things they do are built on their ability to measure VSWR. But first, that VSWR trick is very useful. Not only can you detect degradation of an antenna in place, but you can even pre-tune a beam antenna on the ground, before putting it up, without the ground affecting our measurement. Test the antenna pointing straight up, with the reflector a few feet above the ground. This might involve a wooden ladder and some ropes for temporary support, and is obviously easier with a VHF or UHF antenna than a 40-meter beam.

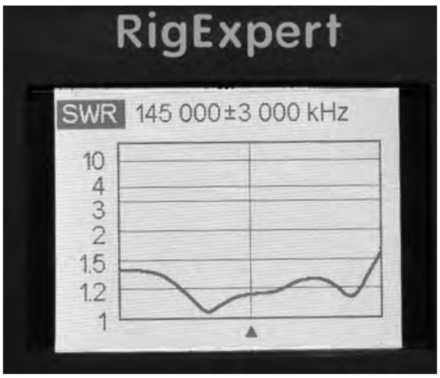
Rotor Loop Test

Once an antenna is up in the air, its cables start to degrade. For a rotatable antenna the rotor loop — the flexible part of the feed line — is often the first component to cause trouble. Any significant variation in swept frequency VSWR display in Figure 2 while the antenna is being rotated could indicate a broken or loose connection or water in the coax.

I found this trick some years ago, before convenient portable antenna analyzers. I managed to connect a large network analyzer, too heavy to lift, to a misbehaving antenna. On the swept VSWR or Return Loss display, the hills and valleys moved around as the antenna rotated. I had used coax with air cavities for the rotor loop and water drained out after I removed a coax connector.



(A)



(B)

Figure 2 — VSWR of 2-meter beam swept across entire 2-meter band on (A) the AAI N1201SA, and (B) RigExpert analyzers.

Feed-line Loss

All feed lines have some loss, and bad feed lines exhibit more loss than others. We can measure feed-line loss between the shack and an antenna without climbing the tower. Most antennas will reflect nearly all of the power at some out-of-band frequencies. If all the power is reflected at the antenna, then the difference between forward and reflected power — the Return Loss measured in the shack — is due to the power lost in the feed line. Since the reflected signal has traveled through the feed line twice, the cable loss is half the Return Loss in decibels.

We find the cable loss by widening the frequency range until we find frequencies with very high VSWR, or low Return Loss. Then we can estimate the cable loss as half the Return Loss. In Figure 3, the Return Loss plot of my 144 MHz Yagi is swept over a wider frequency range than in Figure 2. The antenna is well matched (high Return Loss) in the band, but the Return Loss outside the band is small, about 4.9 dB. The feed-line loss is half of that, or about 2.5 dB.

Cable Length or Distance to Fault

Some cables are worse than lossy: they are broken, shorted, kinked, or disconnected. A large discontinuity in a cable, like an open or short, causes a large reflection. Transmission lines are also impedance transformers. A quarter-wavelength of transmission line inverts the impedance Z, so that a short circuit ($Z = 0$) at one end appears as an open circuit (Z = infinite) at the other end. A second quarter-wavelength inverts the impedance again, back to the original impedance. A third quarter-wavelength inverts again, so ¾ λ is the same as ¼ λ. Additional quarter-wavelengths repeat the pattern, so that every half-wavelength produces an identical impedance. Alternatively, we can look at frequency. A half-wavelength at one frequency is two half-wavelengths at twice the frequency, three half-wavelengths at three times the frequency, and so on.

We use this property to find cable length with an open circuit or short on the far end, or the distance to a fault on a transmission line. Displaying impedance Z on the analyzer over a wide frequency range will show a series of high impedance points and low impedance points (Figures 4 and 5). The high impedance peaks are much sharper, so it is easier to read their frequencies. The difference in frequency ($\Delta Freq$) between any two peaks is the frequency where the cable is an electrical half-wavelength long. We can then calculate the length:

$$\text{Electrical Length} = \frac{c}{\Delta Freq \cdot 2}$$

where c is the speed of light, ~3×10^8 m/s.

In Figure 4A, we adjust the Marker (Mk) frequency at the top of the screen to find the frequency of the first peak, 251.5 MHz. Similarly, in Figure 4B, the Marker frequency is at the second peak, 378.5 MHz. The frequency difference $\Delta Freq$ is 127 MHz. Neglecting the velocity factor, we can calculate the electrical length of the cable as 1.18 meters.

Figure 5 shows a measurement of a longer cable. The frequency difference between 149.4 MHz in Figure 5A and 171 MHz in

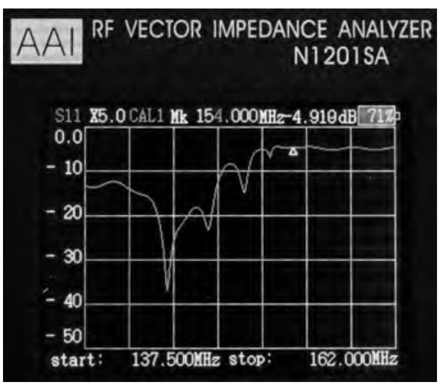
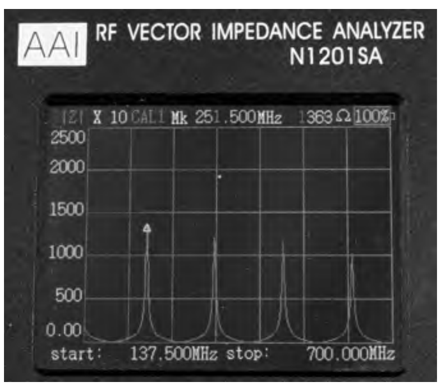
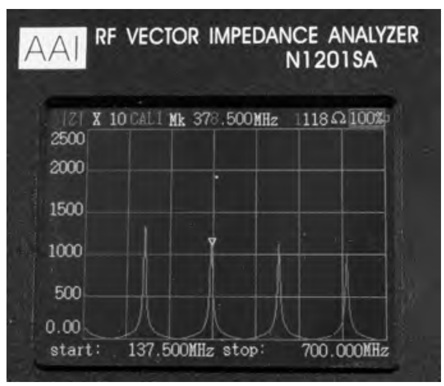


Figure 3 — VSWR sweep over wider frequency range to find cable loss.



(A)



(B)

Figure 4 — Using Markers to find frequencies of impedance peaks for estimating cable length. The marker is set at (A) to 251.500 MHz, and at (B) to 378.500 MHz.

Figure 5B is 21.6 MHz, for an electrical length of 6.94 meters.

Velocity Factor

Common coaxial cable is constructed with a plastic dielectric, which slows down the RF, so it does not travel at the speed of light through the coax, but at a slightly slower speed, typically about 2/3 of the speed of light, so the velocity factor would be 0.66 or 66%. Cables with foam dielectric have less plastic and more air in the foam, so they have a velocity factor that is a bit higher. For an unknown cable, we can estimate the velocity factor V_f if we know the physical length:

$$V_f = \frac{\text{Physical Length}}{\text{Electrical Length}}$$

But if we don't know either, say a roll of cable too big to unwind, a good guess for the velocity factor would be 66%, a common value for ordinary coaxial cable. Then we can roughly estimate the length:

$$\text{Physical Length} = V_f \cdot \text{Electrical Length}$$

For the cable in Figure 4, 0.66 × 1.18 meters = 0.78 meters — my tape measure says 0.79 meters — and about 4.9 meters for the cable in Figure 5.

Cable Characteristic Impedance

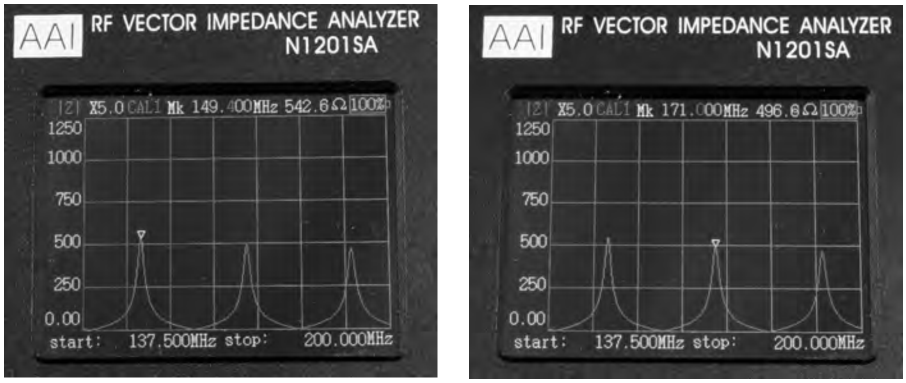
I bought a bag of nice jumper cables at a ham fest, but when I used them, results were sometimes strange. So I tested one with the antenna analyzer, with a 50 Ω termination on the end, resulting in the plot shown in Figure 6. The plot of *R*, the resistive part of the impedance, for a standard 50 Ω cable would be 50 Ω at all frequencies, so this cable has a different characteristic impedance Z_0. The *R* plot shows a repeating pattern — it repeats every half-wavelength, like the cable length measurement. At frequencies where the cable is ½ λ long, the impedance Z_{in} seen at the input is the same as at the load at the far end: $Z_{in} = Z_{load} = 50\ \Omega$. Halfway between the 50 Ω frequencies, the impedance is much higher: $Z_{in} = 108.7\ \Omega$ at the marker. At these ¼ λ frequencies, the impedance Z_{in} is resistive, so the calculation for the impedance transformer is simple:

$$Z_{in} = \frac{Z_0^2}{Z_{load}}$$

then,

$$Z_0 = \sqrt{Z_{in} \cdot Z_{load}} = \sqrt{50 \cdot 108.7} = 73.7\ \Omega$$

and the cable characteristic impedance Z_0 is calculated as 73.7 Ω. This is obviously a 75 Ω cable, another common cable impedance.



(A) (B)

Figure 5 — Impedance peaks for a longer cable; (A) at 149.400 MHz, and (B) at 171.000 MHz.

Figure 6 — Swept frequency plot of R, for calculating cable characteristic impedance.

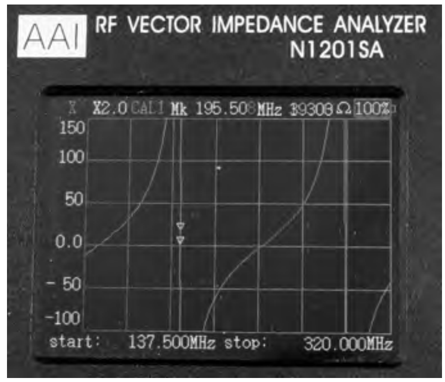


Figure 7 — Use the frequency where reactance X changes abruptly to match cable lengths.

Phasing Line Matching

Phasing lines with matched electrical length are often needed for antenna arrays. We can calculate lengths as we did above, but it is possible to match the lengths more accurately. At the frequencies where the impedance peaks, the reactance changes from inductive to capacitive. We can see this by switching the display to show reactance *X* as in Figure 7 — the reactance changes abruptly from positive (inductive) to negative (capacitive). We can measure the frequency of the abrupt change in sign accurately, and trim each cable to the same frequency and have identical electrical lengths. The exact frequency doesn't matter.

Tuning Stubs

Quarter-wavelength stubs are often used as filters for harmonic reduction or as traps for removing a specific unwanted frequency. As we saw when measuring cable length, quarter-wavelength of transmission line with a short circuit on the far end looks like a very high impedance at the input. But at twice the frequency, it is a half-wavelength long, so the input also looks like a short circuit — the second harmonic is shorted out by a ¼ λ stub.

The stub is connected to a T-connector at the antenna analyzer with a 50 Ω termination on the other leg, as shown in Figure 8. Then the stub can be tuned for maximum VSWR at the desired frequency, and the display switched to show *X* for fine tuning. We can also estimate how sharp the stub response is, by the frequency range with very high VSWR, and see whether the stub has any effect on VSWR at operating frequencies.

Several stubs can be used to reduce multiple harmonics, with the stubs separated by lengths of transmission line or, at lower frequencies, by inductors[1]. The whole assembly can be tested just like a single stub, to see that the VSWR is low at the desired frequency and high at the harmonic frequencies.

A final test would be to replace the 50 Ω termination with the antenna.

Grid Dip Meter Replacement

RF equipment is full of tuned circuits, and it is important to know the resonant frequency. Back in the Dark Ages, every home brewer had a Grid Dip Meter. This is a tunable oscillator with an external coil.

Figure 8 — Measuring resonant frequency of a stub

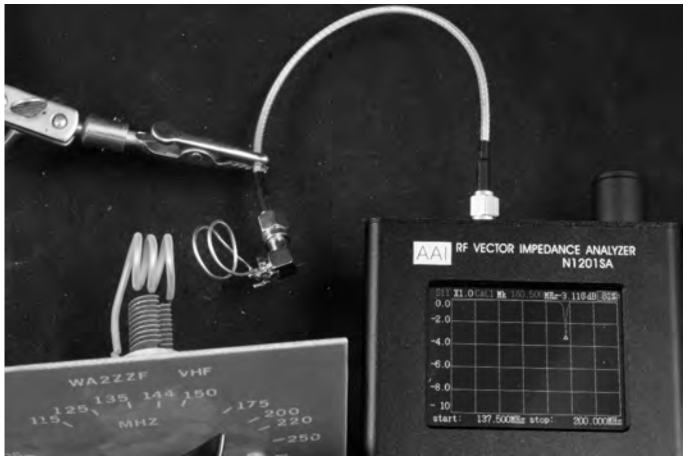


Figure 9 — Using the Antenna Analyzer as a Grid Dip Meter to find the resonant frequency of a tuned circuit. The alligator clip is for mechanical support.

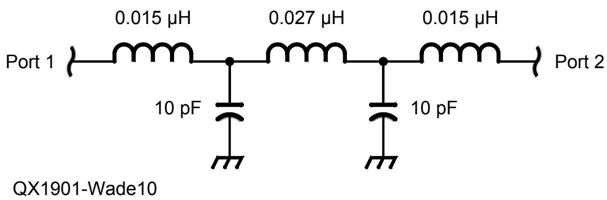


Figure 10 — Schematic diagram of a 432 MHz low-pass filter.

When the coil is held near a resonant circuit, some of the oscillator energy is sucked out of the coil when the Grid Dip Meter is tuned to the resonant frequency, and the grid current would decrease (dip). Thus we could find the resonant frequency of a tuned circuit, even when it is in equipment. The trick is to couple loosely, otherwise the oscillator circuit and tuned circuit detune each other and produce an erroneous frequency reading.

Solid-state versions lack grids, so they used diode detectors to detect oscillator energy in the coil. But all these instruments seem to have gone out of favor. Mine is more than 50 years old, and is getting finicky (like its owner).

An antenna analyzer can work as a dip meter — it has an oscillator and a detector. We attach a small coil to the analyzer, with a short cable for convenience, and place the coil near a tuned circuit to couple some energy. The maximum amount of energy is coupled to the tuned circuit at its resonant frequency, producing the dip in the |S11| or Return Loss trace in Figure 9. Tuning the circuit moves the dip, allowing us to tune the circuit to the desired frequency. Moving the coil closer or farther away can demonstrate detuning changing the resonant frequency.

While the photo shows an isolated tuned circuit, this technique will often work with components in-place in equipment. Vacuum tubes have high impedances, especially with no voltages applied, but solid-state equipment usually has lower impedances so the dip is much less pronounced.

Inductance Measurement

It is very difficult to measure the small inductances we use at VHF and higher frequencies with a meter — the inductance of test leads is greater than the desired inductance. I recently needed a low-pass filter for a small power amplifier module at 432 MHz to remove harmonics from the output. Using free software, I designed one that provides the desired performance, but the inductor values, shown in the schematic diagram in Figure 10, are quite small.

I guessed that a small 3-turn inductor might be in the ballpark for the small inductances. I wound one around a Q-tip and soldered it to a scrap of printed-circuit board in parallel with an 18 pF chip capacitor to form a resonant circuit, and added an SMA connector to connect the Antenna Analyzer.

The inductor and capacitor form a parallel-resonant circuit, which has high impedance at the resonant frequency and lower impedance at other frequencies. Figure 11 shows the impedance plot *Z* on the antenna analyzer, with the marker at the resonant frequency, 226 MHz. The resonant frequency calculation is

$$f = \frac{1}{2\pi\sqrt{LC}}.$$

Turning this around, inductance is

$$L = \frac{1}{C \cdot (2\pi f)^2}$$

The inductance to resonate with 18 pF at 226 MHz is 28 nH, close to the desired value for the center inductor — inductance can be adjusted a bit by squeezing the turns together or stretching them apart.

Figure 11 — Measuring resonant frequency of a parallel-resonant circuit to determine the value of a small inductance.

Removing one turn and re-soldering changed the resonant frequency to 289 MHz, for an inductance of 17 nH, which is in the ballpark for the smaller values.

Filter Measurement

I assembled the low-pass filter on the amplifier printed-circuit board with temporary SMA connectors to measure the VSWR that the amplifier module will see with a 50 Ω termination at the filter output. A low-pass filter should act like a transmission line below the cutoff frequency, but have a high VSWR at frequencies above the cutoff frequency so that those frequencies are reflected rather than passing through.

The VSWR plot in Figure 12 for the low-pass filter shows a low VSWR, like a transmission line, below 432 MHz, and very high VSWR above 432 MHz, with a 50 Ω termination at the filter output. Note that I designed the filter for optimum performance at the operating frequency at the expense of a slightly higher VSWR at lower frequencies — a very good 432 MHz filter rather than a good low-pass filter. Also, for each 10 pF capacitor, I used two 5 pF chip capacitors in parallel for lower loss.

I must admit that the plot in Figure 12 wasn't the first try. The first one, with the inductors from Figure 11, was slightly high in frequency even after squeezing the inductors — I hadn't allowed for the stray capacitance of the PC board. Winding new inductors with a slightly larger diameter got it right.

A band-pass filter may also be measured with the Antenna Analyzer. With a 50 Ω termination at the filter output, the VSWR should be low in the passband, where the filter is acting as a transmission line, and high at other frequencies. Figure 13 shows the VSWR plot of a band-pass filter for 432 MHz.

This simple filter measurement is also useful for unknown filters at ham fests. I occasionally see a box of filters with cryptic markings. A quick test with the Antenna Analyzer might find a useful one, or one at a frequency close enough to retune.

Filter Tuning

Filters with multiple resonators are valuable for removing undesired signals, either coming through the antenna or generated by our equipment. They can be devilishly hard to tune without good test equipment. But we do have good test equipment — the antenna analyzer. If the antenna analyzer has a Polar or Smith Chart display, which mine does not, we can try a method described by Martin and Ness[2].

Most of the filters I have built, and many used by hams, have only two or three coupled resonators. These relatively simple filters can be approximately tuned with the antenna analyzer. Start by detuning all the resonators, then connect the analyzer to one end and tune the nearest resonator for a dip in the |S11| or Return Loss trace, like Figure 9, at the filter center frequency. Then connect the analyzer to the other end of the filter and again tune for a dip; if there are only two resonators, there may be a double dip. Next, connect a 50 Ω termination to the far end of the filter, and the Return Loss should improve significantly for a filter with two resonators. Filters with a third resonator in the center should now be tuned for best Return Loss. This filter tuning should be close enough to fine tune for maximum signal in the system.

Figure 12 — Input VSWR of low-pass filter with 50 Ω termination on the output.

Vector Network Analyzer

Since the antenna analyzer is really a network analyzer, why not use it as one? The Vector term implies complex impedance, $R + jX$; our analyzer provides both R and X, which can be put together as a complex impedance. Then the complex impedance can be plotted on a Smith Chart or entered into a software simulator to do impedance matching. For instance, in *QST* "Microwavelengths" for July 2016, I used a network analyzer to do simple impedance matching using transmission lines and stubs — an antenna analyzer would do just as well.

Please note that I have said "about" or "approximately" quite frequently. A laboratory Vector Network Analyzer requires a careful calibration procedure to be performed often, and then uses computer corrections for high accuracy. With an antenna analyzer, accuracy to even one decimal place is optimistic, but is quite adequate for ham use.

Summary

An antenna analyzer is too useful a piece of test equipment to be limited to antenna measurements. As a network analyzer, it can do much more, and is far more convenient, portable, and affordable than a commercial network analyzer. These are a few of the things you can do with one, but hams are resourceful and can probably come up with others. So teach your analyzer some new tricks.

Figure 13 — Input VSWR of a band-pass filter with 50 Ω termination on the output showing the filter frequency range.

Acknowledgements

Thanks to Don Twombly, W1FKF; Chip Taylor, W1AIM; and Matt Reilly, KB1VC; for help with this project.

Paul Wade, W1GHZ, previously N1BWT and WA2ZZF, has been licensed since 1962 and has never made a contact below 50 MHz. He has been a microwave experimenter for years and has published numerous articles as well as the "Microwavelengths" column in QST. He is active in the Vermont 10 GHz group and is past president of the North East Weak Signal Group. An ongoing project is the "W1GHZ Microwave Antenna Book", online at ***www.w1ghz.org****.*

In 1997, he was honored by the Central States VHF Society as the recipient of the Chambers Award. He has been honored by the ARRL with the 2000 Microwave Development Award, in 2001 with the Thomas Kirby Eastern VHF/UHF Society Award, and in 2009 by Microwave Update and the North Texas Microwave Society with The Don Hilliard Award for Technical Contributions to the Microwave Community.

After a long career in electrical engineering, he and Beth, N1SAI, are now happily retired in Vermont with a new puppy named Hannah. Paul was also a ski instructor for a time, and now enjoys skiing on a new bionic knee, and skijoring with Hannah.

Notes

[1]John Regnault, G4SWX, "Coaxial Stub Filters," **www.ifwtech.co.uk/g3sek/swxfiltr/swxfiltr.htm**.

[2]Peter Martin and John Ness, "Coupling Bandwidth and Reflected Group Delay Characterization of Microwave Bandpass Filters," **www.rfshop.co.uk/Martin.pdf**.

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Novel Method for Impedance Measurements

A Howland Current Source forms the basis of an experimental impedance measuring instrument.

Impedance measurements are basic to components and other circuits. These methods can be found in numerous references and in the Keysight Technologies "Impedance Measurement Handbook 6th Edition".[1] Commonly used are the bridge method, resonant method, I-V method, and network analysis method. The auto-balancing bridge method is often used in impedance measuring instruments.

The method that I used is similar to the I-V method except that the current is constant and instead the voltage varies with the impedance. The constant ac current source is a Howland current pump or source. A complete theory and circuit of the Howland current source can be found in the Texas Instruments application note, AN-1515 A "Comprehensive Study of the Howland Current Pump".[2] There are other sources for the description of the Howland current source as well.

The Howland Current Source

The basic Howland current source (Figure 1) was invented by Prof. Bradford Howland at MIT about 1962. This circuit was never patented and appears to be an open-source circuit. In this application however, the theory is straight-forward in the basic form, but the practical implementation problems involved are not simple or obvious.

According to AN-1515 the output current is,

$$I_{out} = \frac{E_{in}}{R} \tag{1}$$

Voltage E_z across the load impedance Z is,

$$E_z = I_{out} Z \tag{2}$$

Manipulating Eqs (1) and (2), we get

$$\frac{E_z}{E_{in}} = \frac{Z}{R} \tag{3}$$

from which,

$$Z = R\frac{E_z}{E_{in}}\,. \tag{4}$$

Z can be expressed as the magnitude of Z and the phase of Z. We get the phase of Z if we measure the phase shift between E_z and E_{in}. With the value of Z magnitude and phase angle, we can calculate $Z = R + jX$. Looks easy!

One nice feature of this method is that one side of Z is grounded, making it easier to measure components. Many measurement methods require both ends of the component floating. This can be a negative point for the user. In addition the circuit is dc-coupled, allowing operation from dc to the high frequency limit of the op-amp.

The Op-amp is the Major Limitation of the Circuit

The resistor values in Figure 1 must be equal and matched very closely in tolerance. This is easily accomplished because tight tolerance resistors are easy to obtain. The real challenge is the op-amp. The selected op-amp must have a low input bias current, high input impedance, low noise, high gain-bandwidth product, large voltage output swing, smooth open-loop phase shift, and reasonable output current.

These characteristics will become more obvious as you use the circuit. Some characteristics, such as the input noise of the op-amp will limit how small a signal can be measured at E_z, affecting the dynamic range. The op-amp gain-bandwidth product will determine the frequency range and the shunting capacitance. The OPA211 is the best op-amp that I have found after working with this circuit for a number of months.

Another requirement is that the generator for E_{in} should be 0 Ω source impedance at all frequencies. This is attainable at lower frequencies with an op-amp driver, but impedance will increase at the higher frequencies, introducing some error. This is one of many sources of error that can occur when using real op-amps.

Measuring E_{in} and E_z

Measure E_{in} and E_z with voltmeters and then calculate the value of Z using Eq (4). Voltmeters do not usually show phase so the answer would be magnitude only. Knowing phase is important to determine if the component is inductive, capacitive or somewhere in-between.

The value of R sets the reference for the Howland current source. This value is a compromise of dynamic range and shunt capacitance of the current source. I have found that 1000 Ω is a good choice. Thus all the resistors in the circuit of Figure 1 are 1000 Ω within very tight tolerances. The measurement of the ratio of E_z / E_{in} multiplied by R will now yield the value of Z.

Recently Keysight introduced a low cost digital storage oscilloscope (DSO) with a software program that would measure gain

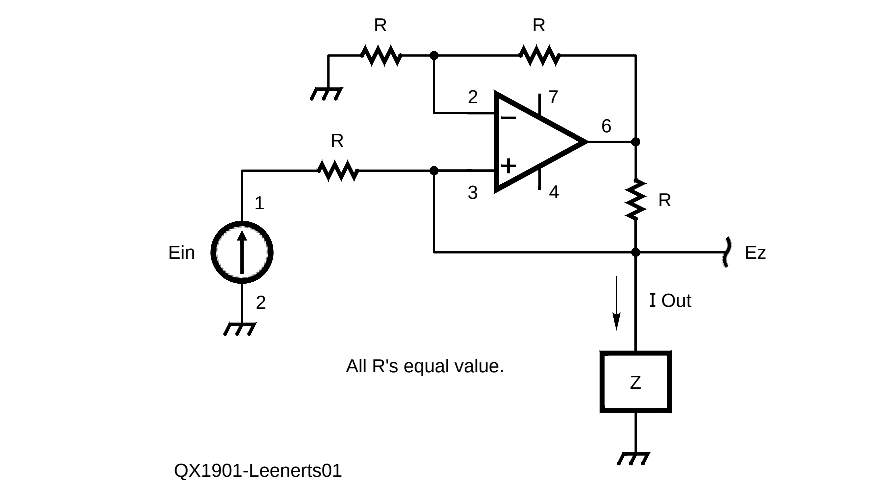

Figure 1 — Basic circuit of the Howland current source.

and phase. The software is called *Frequency Response Analysis* (*FRA*). When used with the built-in waveform generator, it plots out the gain and phase as a function of frequency. I used the Keysight Model DSOX1102G DSO. This was a good solution to measure E_z and E_{in} for the impedance measuring instrument using the Howland current source. The *FRA* software displays the ratio (Gain) of channel 1 and 2 in a log scale calibrated in decibels. Phase is displayed in +/- degrees as a function of frequency. The frequency is also displayed on a log scale calibrated in frequency decades. The log frequency scale is a little lacking as it displays only the frequency decade markers, not points in between, such as 2 and 5. However, there is a marker that can be moved to a selected point to display the ratio in dB, phase in degrees and frequency in Hertz.

The measurement and display of the ratio E_z / E_{in} in dB does not give the answer in ohms, the desired unit. Thus we need to know that dB is the scaled log ratio of two numbers. When the two numbers are equal, the ratio is 1 and the log of 1 is 0. Using one of the numbers as a reference, we then know the level difference between the reference level and the measured level. By definition 1 mW into 50 Ω is 0 dBm. Spectrum analyzers and signal generators are often calibrated in dBm.

The selected1000 Ω reference resistor is the value used in the Howland current source for the impedance measurement instrument. A measurement of 0 dB means that *Z* equals 1000 Ω. A measurement of +20 dB means that *Z* is 10 times larger than 1000 Ω or 10 kΩ. Conversely, if the measurement is -20 dB, *Z* is 10 times smaller than 1000 Ω or 100 Ω. If, for example, the measurement is -8 dB, use the antilog (10^{-Z}) function in a math calculator for the answer. So, -*Z* = -8/20,which equals -0.40. The antilog of -0.40 multiplied by 1000 Ω is 398 Ω.

Impedance Measurement Instrument Circuit

The circuit for a functioning Howland current source consists of three op-amps (Figure 2). The first op-amp U1 buffers the input from the signal generator so that the input resistance of the current source is driven as close to 0 Ω as possible over the frequency range. It also drives the E_{in} reference output. The second op-amp U2 is the Howland current source for the current through *Z*. The third op-amp U3 is the output buffer to measure the voltage E_z with as little loading as possible and drives the E_z output.

The condition for the circuit is that the ratio of R11/R12 = R13/R14. Even with 0.1% resistors, a slight ratio mismatch will degrade the current source output impedance. Trimmer resistor R9 in series with R10 will allow the R11 value to be adjusted to the exact value that achieves a high output impedance. The circuit is powered by a +/-15 V lab power supply. The current draw is approximately 15 mA.

Construction Notes

I built my instrument in a small Bud Industries cast aluminum box #CU-124. Since this is an experimental project and I am not set up for working with surface mount parts, I did my own point to point wiring. I used proto boards (Jameco part number 2130263) to connect to the surface-mount OPA211 op-amps.

I used Grayhill 29-100 series spring test clips available from Mouser. The ground test clips were spaced differently so that I could better accommodate different size parts for testing. The RF jacks are standard BNC connectors. My power input connector is a 5-pin DIN connector. Figure 3 shows an inside view of my instrument.

Adjusting the Z_{out} Trimmer

Before using the instrument we must adjust the Z_{out} trimmer for the output impedance of the current source. First, connect the output of the built-in DSO generator to the input E_{in}. Next, connect the input channel 1 to the E_{in} output, and the input channel 2 to the E_z output. Figure 4 shows my setup.

Adjust the signal generator to a level of 6 mV P-P at a frequency of 100 Hz. The setting is not critical. Since the gain is high, the input level must be small to not be limited by the op-amp. Since the corner frequency is in the kilohertz range, the frequency must be set low so that the output impedance is resistive. Figure 4 shows that E_{in} and E_z are in phase. I adjusted the Z_{out} trimmer for slightly less than maximum output impedance, which corresponds to an output voltage of 12 V P-P. The output impedance is now,

$$Z = R\frac{E_z}{E_{in}} = 1000\frac{12}{0.006} = 2\ \mathrm{M\Omega} \qquad (5)$$

2 MΩ is high enough for the typical measuring range using a 1000 Ω reference.

The output capacitance of the current source is of interest, and can be both calculated and measured. The equation for the capacitance from Appendix D of AN-1515 is,

$$C = \frac{1}{2\pi BW}\frac{1}{\mathrm{R2}}\left[\frac{\mathrm{R4}+\mathrm{R3}}{\mathrm{R3}}\right]^2 \qquad (6)$$

where BW = 45 MHz is gain-bandwidth product of the OPA211 op-amp and the values of the resistors are 1000 Ω each, so *C* is 14 pF.

The measurement connections are the same as for adjusting of Z_{out} trimmer. Set the frequency to 100 kHz, well above the corner frequency, where the output *Z* is capacitive. Set the generator to 60 mV P-P for E_{in} and the measured E_z is 5 V P-P. From these values, Z_{out} is 83 kΩ. This is the reactance of the capacitance in parallel with the Z_{out} resistance. By calculation *C* = 19 pF. Taking into account a little for the stray capacitance for the input terminal, this is in very good agreement with the Eq (6) value of 14 pF.

Using the Impedance Measuring Instrument

This instrument is experimental and not a precision laboratory grade measuring instrument. The accuracy depends on more than just the 0.1% resistors. It also depends on other factors such as the impedance

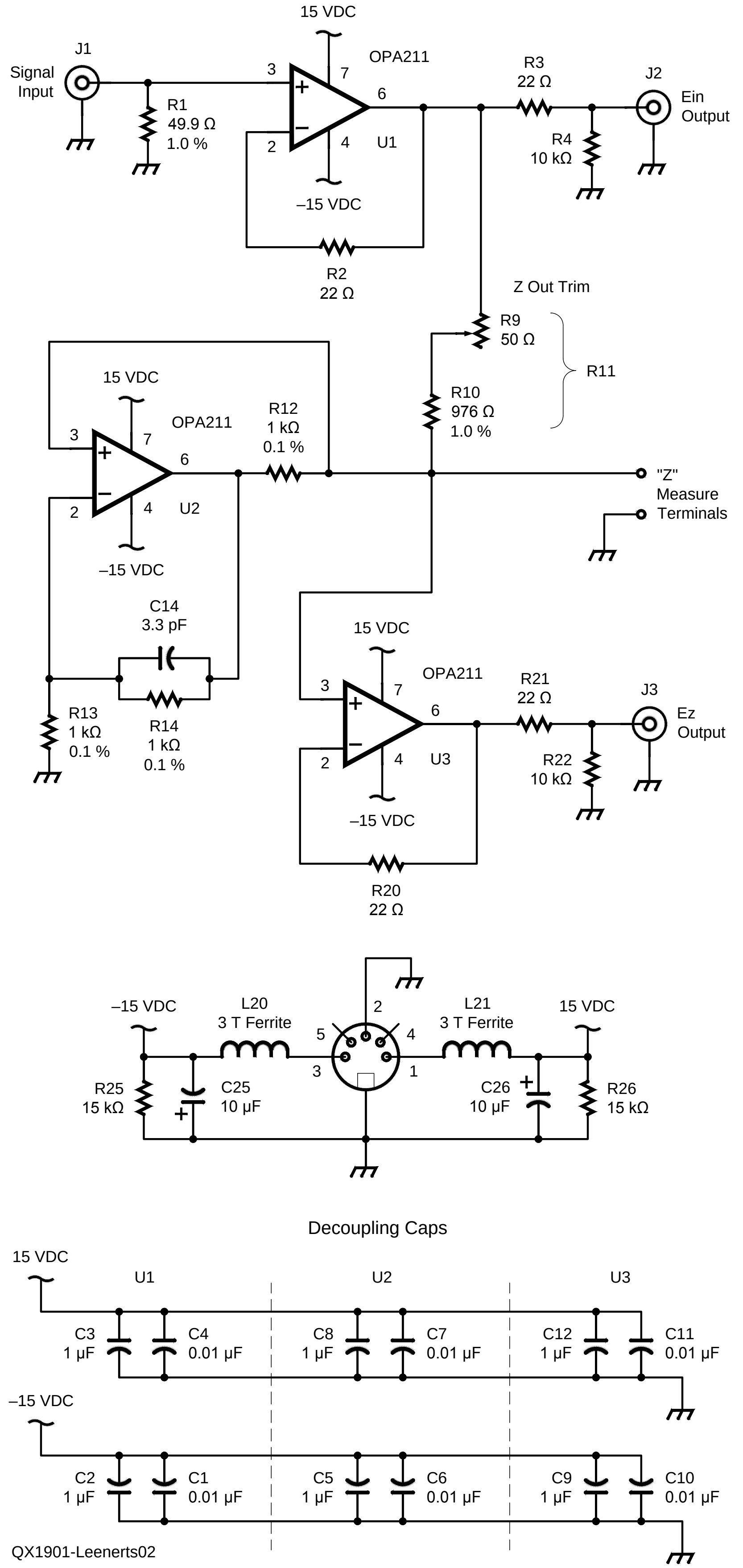

Figure 2 — The circuit for a functioning Howland current source consists of three op-amps.

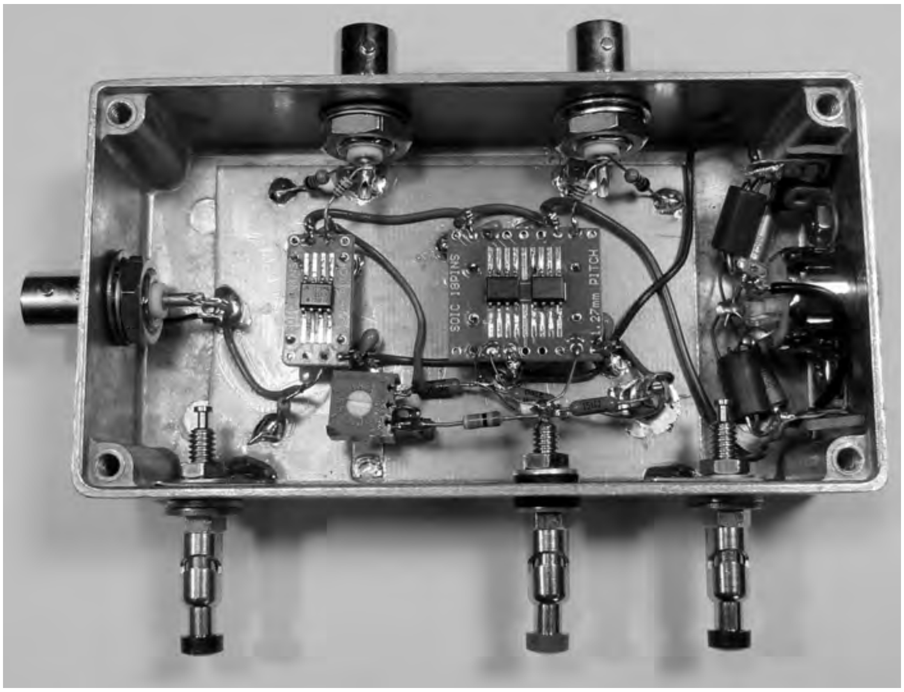

Figure 3 — Inside view of the instrument.

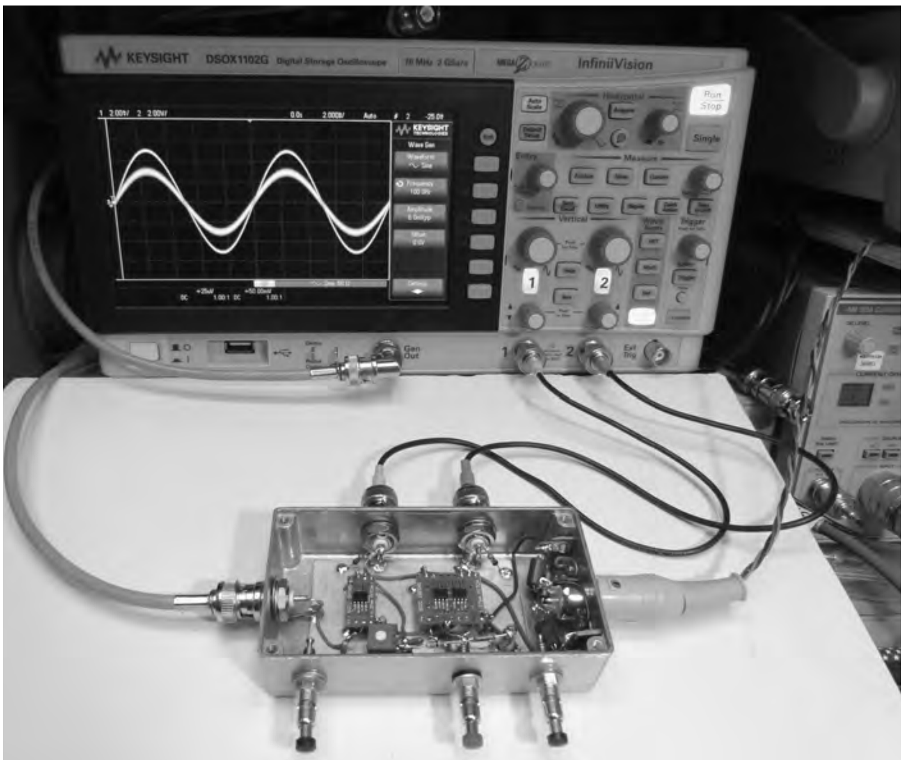

Figure 4 — Adjust the Z_{out} trimmer for max Ez.

driving the input resistor R11, which should be zero. There are also errors affecting accuracy in the measuring DSO. The accuracy varies as a function of level and frequency, and needs to be factored in when seeking an accurate measurement. I estimate that the nominal accuracy varies around 5%.

The goal of this instrument is to be able to plot the impedance of components or circuits as a function of frequency using log scales as seen in text books and specification sheets. This is an educational experimental instrument, thus accuracy is not the major issue.

Although the *FRA* software in the Keysight Model DSO1102G is straight forward to use, I will cover some of the details related to the impedance measuring instrument. The frequency range of the software is 20 Hz to 20 MHz. The frequency buttons set the *min* and *max* frequency in decade steps. For the *min* frequency of 20 Hz, the display will start at 10 Hz. For the *max* frequency of 20 MHz, the display will show the *max* frequency of 100 MHz. The amplitude level is in V P-P, and has plenty of output range. When using the impedance measuring instrument, the output level setting is one of the most important settings to track. Depending on what is being measured, setting the level too high will cause the current source to limit and clip its output signal. Setting the level too low can limit the dynamic range of the measurement. An additional aid in setting the amplitude is the display of the two channel signals between each frequency step as well as the main plot screen. This feature will aid in setting the level for good measurement results. Experience will inform the user on level setting for good results.

The number of steps per decade was an initial issue with the DSO as it had far too few steps per decade to give a reasonable display. This was not a problem for a slowly changing response, but if there is a resonance, the peak can be easily missed. I contacted Keysight product support and managed to have the number of steps increased from 10 step increments to 50 steps per decade with a firmware update. This is now a button and is selectable. I do not know the current status of the DSO in production or if the update was made widely available. I have used *LT Spice* frequently and I have found that the minimum number of steps for simulation is approximately 100 steps per decade. The 50 steps per decade setting gives acceptable results and is the number I use for my measurements.

Once the settings have been made, start the program by pushing the Run Analysis button. The program starts by what appears to be the setting of the initial levels for the channel inputs on the plot screen. As the plot starts, it displays the results of magnitude and phase as it steps through the frequency range selected. Also the scope will display the input waveforms as an overlay on the plot screen.

After the scan is completed, only the plot screen is presented. The chart can be adjusted for the desired displayed range of magnitude and phase. The marker can also be moved along the scan to display the magnitude, phase, and frequency at each data point. I like to save plots on a USB drive for reference. I move the marker to the most likely point of interest. The log frequency scale has only decade markers and this is an issue, especially when looking at the saved plot later. The addition of frequency markers at 2 and 5 or more would be very useful. Interpolation between log decade markers is more difficult than interpolation between the linear dB scale marker lines.

There are realistic limitations that need to be noted for the operation of the impedance measuring instrument and the DSO. When measuring a small value capacitor at high frequencies, the DSO may have difficulty in discerning if the phase is really -90° and generates a spike toward +90°. When measuring an inductor, usually with small values and high self-resonant frequencies, the Howland current source oscillates making a measurement not possible. I encountered no other limiting idiosyncrasies.

Despite the limitations, the usefulness and fun of using the instrument is rewarding. Using the DSO with the impedance measuring instrument is easier as you gain experience.

Measuring a Capacitor

The following is an example of how to setup the FRA program to measure a 1 μF capacitor. Set the *min* frequency to 20 Hz and the *max* frequency to 10 MHz. Set the generator output level to 1 V P-P. The following is the same for all the subsequent example setups. The output load is set for 50 Ω. Set the number of points per decade to 50. Push the Run Analysis button to start the scan. Figure 5 shows the scan result.

The marker was placed at -20dB, which is a reactance of 100 Ω and the frequency is the expected nominal 1.59 kHz. Where the marker is placed, the phase is -90° which is expected for a capacitor. The dashed trace of the gain (reactance) is the text book reactance chart sloping 20 dB per decade until around 100 kHz. Then the real world shows up with the series resonance of the capacitor of around 1 MHz, and the phase (solid trace) shifts from -90 degrees to an inductive +90°.

Measuring an Inductor

The following is an example of the setup for the FRA to measure a 10 mH inductor. Set the *min* frequency to 100 Hz and the *max* frequency to 1 MHz. Set the output level of the generator to 10 mV P-P. Figure 6 shows the scan result.

The plot is another real world scan of an inductor. The marker is at 15.85 kHz and the impedance (solid) is very close to 0 dB, which is 1000 Ω with a phase shift (dashed) of nearly +90°. The plot at the marker shows that the inductor is just starting to become inductive after being nominally a resistor at 100 Hz. The dc resistance value for the inductor is about 63 Ω. Then the plot shows the self-resonance frequency of the inductor

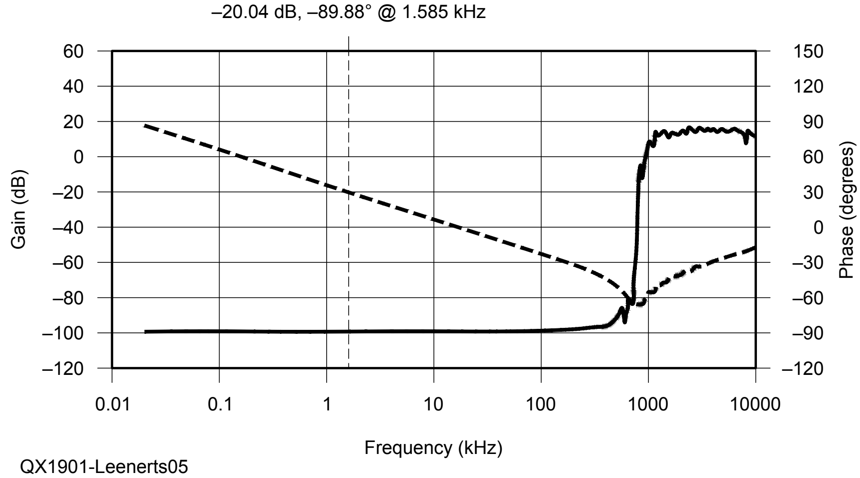

Figure 5 — Measured response of a capacitor, phase (solid) and amplitude (dashed).

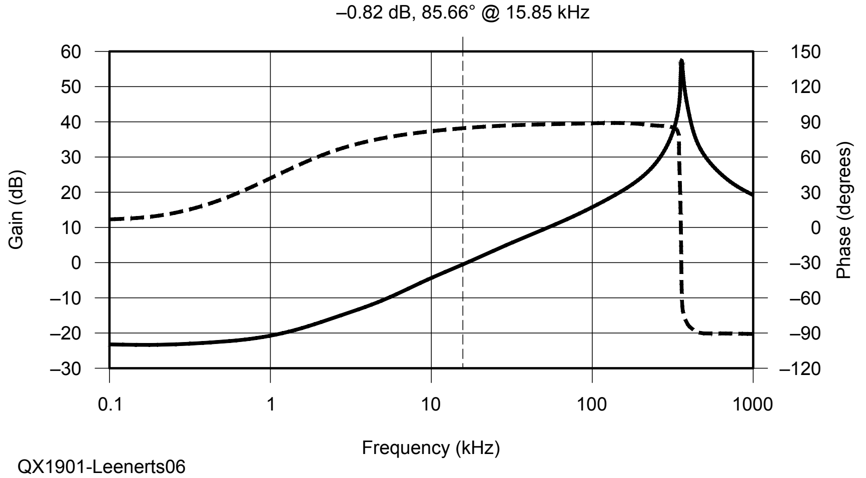

Figure 6 — Measured response of an inductor, phase (dashed) and amplitude (solid).

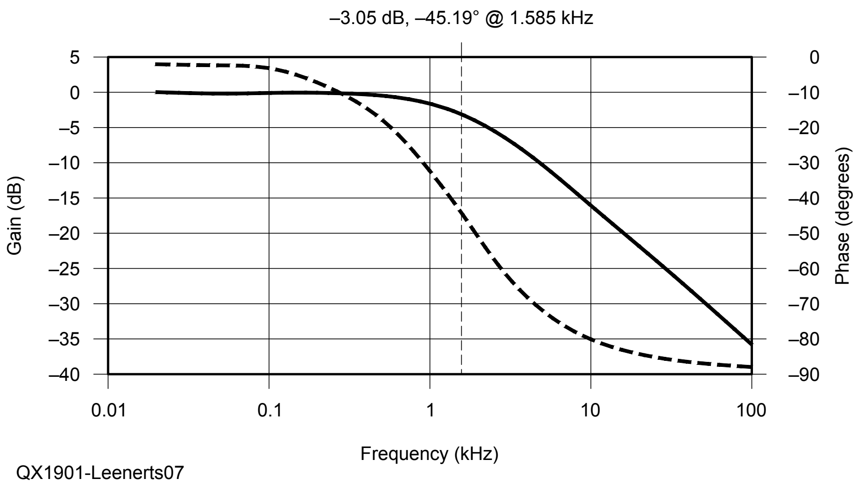

Figure 7 — Measured response of a parallel RC network, phase (dashed) and amplitude (solid).

at a nominal 360 kHz with an impedance of 500 kΩ. The frequency is not the actual self-resonance frequency as the current source shunting capacitance of 19 pF lowers that frequency. Also note that the phase changes from +90° to -90° after the parallel resonance of the inductor.

During the measuring process, oscillations on the DSO waveforms will be noted especially during the beginning of the scan. The oscillations seem to be caused by the impulse of the step frequency that causes the Howland current source to have ringing oscillations. When I sweep the frequency with the generator manually I do not see the oscillations so I have no solid conclusions. However, the *FRA* program appears to not be affected and produces good results.

Measuring a Parallel RC Network

The following is an example of how to setup the *FRA* program for measuring a parallel RC network. Set the *min* frequency to 20 Hz and the *max* frequency to 100 kHz. Set the generator output level to 2 V P-P. Figure 7 shows the resulting scan. The plot is the classic text book curve for a one-pole low-pass Butterworth filter. The plot starts out with the resistor value of 0 dB or 1000 Ω (solid trace). Then at the cutoff frequency of 1.59 kHz the response is down the expected 3 dB with a phase shift of -45 degrees (dashed line). After the -3 dB frequency point, the slope between 10 kHz and 100 kHz follows the expected 20 dB per decade.

Measuring Transformer Input Impedance

The following is an example of how to setup the FRA program to measure a telephone coupling audio transformer. The transformer primary winding is connected to the instrument and the secondary winding is connected to a 600 Ω resistor. Set the *min* frequency of 20 Hz and the *max* frequency to 100 kHz. Set generator output level to 1 V P-P. Figure 8 shows the scan result.

The plot shows how the transformer input impedance (solid line) changes with frequency. The frequency scan starts at 20 Hz and shows that the impedance is dominated by the magnetizing inductance and resistance of the transformer. As the frequency increases to 1 kHz, the magnetizing inductance is high enough that the 600 Ω load resistor is the dominant impedance (solid trace) with phase (dashed trace) close to 0°. As the frequency continues to increase, the leakage inductance in series with the 600 Ω load resistor shows increasing impedance and shifting in phase to become more inductive. The calculated impedance at the 1 kHz marker is -4.67 dB (584 Ω) at 8.43°.

Measuring a Tubular Ferrite

The following is an example of how to setup the *FRA* program to measure the frequency response of a tubular ferrite. The ferrite is a Fair-Rite #2675540002 with one turn. Set the *min* frequency to 100 Hz and the *max* frequency to 10 MHz. Set the output level of the generator to 2 V P-P. Figure 9 shows the scan result.

The plot shows the typical impedance of a ferrite for EMI use. The impedance is the solid trace and the phase is the dashed line. At 100 Hz the impedance is essentially resistive, calculated as 0.25 Ω. As the frequency increases to around 50 kHz, the ferrite is inductive. The calculated reactance is 8 Ω inductive or 25 µH. As the frequency continues to increase the ferrite impedance at 1.5 MHz has a phase of zero and a magnitude of 150 Ω resistive. As the frequency continues to increase the ferrite impedance becomes capacitive but is mostly resistive.

Measuring a 100 µF Capacitor Without the FRA Software

The following is an example of how to measure a 100 µF capacitor with the DSO and the impedance measuring instrument. With connections the same as in the original description of using the *FRA*, turn on the wave generator. Set the generator to 800 mHz and the amplitude to 400 mV P-P. Adjust channels 1 and 2 to 200 mV per division and set for dc coupling. Set sync to normal trigger on channel 1. There is a small dc offset at the current source and will cause the capacitor to charge up in voltage. To

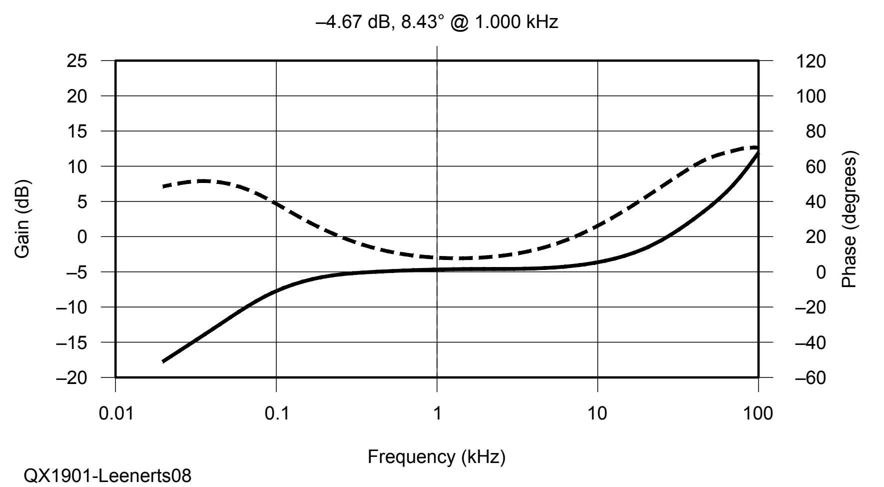

Figure 8 — Measured response of an audio transformer, phase (dashed) and amplitude (solid).

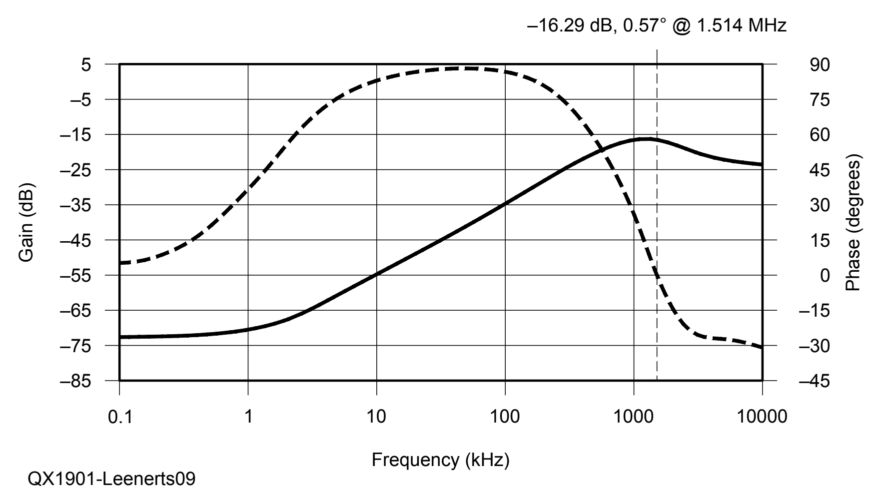

Figure 9 — Measured response of a tubular ferrite, phase (dashed) and amplitude (solid).

slow this charging, connect a 100 kΩ resistor across the capacitor. Figure 10 shows the capacitor measurement result.

The 400 mV P-P lower amplitude is E_{in} signal from the generator. The higher amplitude trace E_z from the instrument shows 800 mV P-P. The ratio is 2 and thus the impedance of the capacitor with respect to the 1000 Ω is 2000 Ω, and the calculated reactance of a 100 µF capacitor at 800 mHz is 1989 Ω.

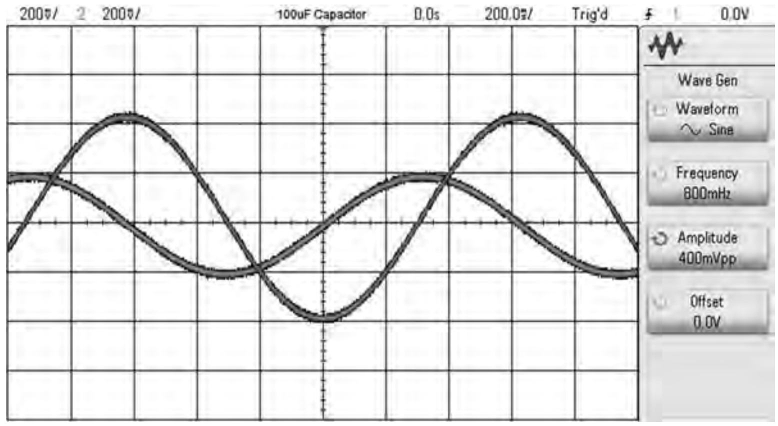

Figure 10 — Measured response of a capacitor without using FRA software. The amplitude ratio (here 2) multiplied by the 1000 Ω reference gives the reactance of the capacitor.

Conclusion

It's been a new, interesting and rewarding experience to work with this circuit. In researching articles, I did not find anything about using the Howland current source as an ac current source to measure impedance. The Keysight DSO *FRA*, a waveform generator, and the Howland current source makes this method of measuring impedance possible. One attractive feature is that the Howland current source measurement method needs no calibration. The basic measurement process depends only on the measurement of ratios with a known value of a reference resistor. Another desirable feature is that one end of the component to be measured is connected to ground.

The Howland current source does have numerous issues. Also, this is a not fully developed experimental method. I plan to make further improvements to this, one of my most rewarding projects.

Virgil Leenerts, WØINK, is an ARRL member and an ARRL Technical Specialist for Colorado. He was licensed in 1954 and earned his BSEE degree from the University of Illinois in 1963. Virgil retired as an electrical engineer with Hewlett Packard and Agilent Technologies after 38 years of service, but continues as a part-time engineer for SCOM designing power supply and audio circuits. He has given many presentations to radio amateurs, including a discussion about switching power supplies at the 2009 ARRL Rocky Mountain Division convention. He is author of "Loop Antennas, the Factor – N", QEX – Jan./Feb., 2011. He is now relearning vacuum tube radios where his interest in electronics began during his high school years. The restoration of WWII radios has been a focus for the last few years. The restoration of BC-611's (Handy Talkies) has been especially rewarding.

Notes

[1]Keysight Technologies, "Impedance Measurement Handbook 6th Edition", **https://literature.cdn.keysight.com/litweb/pdf/5950-3000.pdf**.

[2]Texas Instruments application note, AN-1515, "A Comprehensive Study of the Howland Current Pump", **www.ti.com/lit/an/snoa474a/snoa474a.pdf**.

Trimming the Wire Dipole Antenna Revisited

The procedure[1] in "Trimming the Wire Dipole Antenna" by Luiz Duarte Lopes, CT1EOJ, for referring the impedance at the input to a transmission line to an antenna at the terminating end of the line can be very useful. The same result that is obtained by the author's Eqn. (4) can be implemented by Eqns. (2) and (3) making length *S* negative in value so that statements written in code that implement the equations need not be modified to change the direction of movement of the reference location.

This procedure — making the length negative — is used in referring the input impedance of the line to the antenna impedance[2] in "Octave for Transmission Lines". [*TLW*, distributed with the *ARRL Antenna Book*, however, does not accept negative lengths. — *Ed*.]. Lopez correctly notes that the procedure he describes applies when the attenuation of the line may be neglected. One should, though, be careful about that. If we duplicate the Lopes calculation in his Table A (reproduced here) for just the longest distance, 18 meters, using the *GNU Octave* code from Note 2 with an attenuation of zero, we get an input impedance[3] of 23.454 + j0.57586 Ω. This impedance is close to that of Lopes, the difference probably arising from round-off errors, conversion from meters to feet for input to the script in Note 2, and the fact that the *Octave* code in Note 2 uses hyperbolic rather than circular trigonometric functions for the input impedance equation[4, 5]. Applying Eqns. (6) and (9A) from Chapter 20 of the *2018 ARRL Handbook*[6], we get an SWR of 2.13, matching the calculation in the Table A.

Now let's assume that the transmission line exhibits a loss at 14.250 MHz of 2 dB per 100 feet, typical for some RG-58 coaxial lines at that frequency. The loss itself is probably of little concern, representing only a fraction of an S-unit for this line length. When we recalculate using the *GNU Octave* code from Note 2 with the added attenuation, though, we get an input impedance of 28.412 + j0.49994 Ω. This appears at first glance to be not too far from the lossless value, but when we recalculate the reflection coefficient and then the SWR, we get SWR of 1.76:1, a 17% difference that may be significant to users of transceivers that prefer an SWR under 2:1.

This difference results because the loss in the line will cause the SWR to decrease as we move away from the antenna. The reflection coefficient will approach zero and the SWR will approach 1:1 asymptotically as the distance from the antenna (or the loss per unit length) becomes very large. The increasingly higher loss will effectively mask the input from the load so that the input impedance will approach Z_0. — *Best regards, Maynard Wright, W6PAP,* ***m-wright@eskimo.com***.

Table A, reproduced from Lopes, *QEX*, Sept./Oct., 2018.

Input impedance at the end of a length *S* of transmission line for an antenna impedance Z_a = 84 + *j*37 at 14.250 MHz.

S, m	Z_i, Ω	*SWR*
6	42.7 + j35.1	2.13
9	46.7 – j 37.2	2.13
12	27.2 + j 17.3	2.13
15	91.1 – j 32.4	2.13
18	23.5 + j 0.25	2.13

Notes

[1]L. D. Lopes, CT1EOJ, "Trimming the Wire Dipole Antenna – Technical Note," *QEX*, Sept./Oct. 2018, p. 26.

[2]M. Wright, W6PAP, "Octave for Transmission Lines," *QEX*, Jan./Feb. 2007, pp. 3-8.

[3]To be mathematically correct, we should round off to two significant figures in these calculations to correspond with the precision of the distance specifications in Table A. We will, though, retain additional figures of precision as we are working with small differences between larger numbers and the values here are hypothetical anyway.

[4]β in Note 1 represents the phase shift of the entire line. In Notes 2 and 5, β represents the phase shift per unit length.

[5]R. A. Chipman, "Schaum's Outline Series: Theory and Problems of Transmission Lines," McGraw-Hill, 1968, Eqn. 7.17.

[6]*The ARRL Handbook for Radio Communications,* 2018 Edition [also in the 2019 Edition]. Available from your ARRL dealer or the ARRL Bookstore, ARRL. Telephone 860-594-0355, or toll-free in the US 888-277-5289; **www.arrl.org/arrl-handbook-2018** and **www.arrl.org/arrl-handbook-2019**.

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Upcoming Conferences

New Mexico TechFest

Albuquerque, New Mexico
February 23, 2019
www.rmham.org/wordpress/new-mexico-techfest

The 2019 New Mexico TechFest will be Saturday February 23, 2019 at the New Mexico Veterans' Memorial Event Center located at 1100 Louisiana Blvd SE, Albuquerque, NM 87108. The event is sponsored by Rocky Mountain Ham Radio, New Mexico.

Join fellow Amateur Radio operators for a day of quality presentations, demonstrations, and instruction provided by some of New Mexico's most experienced technical hams on a variety of emerging and relevant technical topics within Amateur Radio today. The New Mexico TechFest is designed to provide a unique opportunity for all hams interested in the technical aspects of our hobby to advance and expand their technical knowledge and to facilitate technical discussion, collaboration, and ideas with one another.

Ham-specific and presentation-related prizes will be drawn throughout the TechFest. Refreshments including coffee, water, and light snacks will be available. Lunch, catered by local small business, will be available at an optional cost.

Additionally, a limited number of tables with ac power will be provided for attendees to set up and conduct informal ham-related technology/project demonstrations and technical poster-board presentations between the formal presentations, during lunch, and at the conclusion of the event. Due to space limitations, attendees interested in providing a demonstration or poster-board presentation should submit details to TechFest organizers for review and consideration. Sales and promotions are not permitted.

See website for complete details.

Southern California Linux Expo 2019 (SCALE 17x)

Pasadena, California
March 7 – 10, 2019
www.socallinuxexpo.org/scale/17x/cfp

The 17th Annual Southern California Linux Expo – SCALE 17X – will be held on March 7 – 10, 2019, at the Pasadena Convention Center in Pasadena, California.

SCALE 17X will have the following specialized tracks: Open Source in Enterprises; Containers and Orchestrator; Open Data; Security; Developer; Next Generation; Mentoring; Embedded hardware and software; Libregraphics; UbuCon; Observability; The Next Generation, and ham radio. In addition, there will be general tracks covering a wide range of topics related to Free/Open Source Software (FOSS) and open hardware. We invite you to share your work on FOSS programs and open hardware projects with the rest of the community, as well as exchange ideas with some leading experts in these fields.

See the website for complete details on activities, programs, registration, and how to submit a paper or presentation.

2019 Tenth SARA Western Conference

Boulder, Colorado
March 23 – 24, 2019
www.radio-astronomy.org

The 2019 (Tenth) SARA Western Conference will be held at the University Corporation for Atmospheric Research (UCAR) Center Green conference facility in Boulder, Colorado, from March 23 – 24, 2019.

The Conference will be preceded by a tour of the NCAR Mesa High Altitude Observatory (HAO) on Friday, March 22, 2019.

Papers and presentations on solar astronomy, radio astronomy hardware, software, education, research strategies, philosophy, and observing efforts and methods are welcome. The deadline for submitting a letter of intent to the conference coordinator including a proposed title and informal abstract or outline is January 15, 2019. Formal proceedings will be published for this conference. Presenters should prepare an electronic copy of their presentation, so that we can make them available to attendees on a CD.

Registration is $80.00. This includes lunch and refreshments on Saturday and Sunday. Current membership in SARA is required for the conference, and non-members will be able to attend upon payment of an additional $20.

On Sunday afternoon, registrants at the conference can travel to Haswell, CO, to visit the Paul Plishner Radio Astronomy and Space Sciences telescope. This trip will require an additional overnight stay in the Haswell area, before visiting the Plishner site.

See website for complete details.

Aurora Conference

White Bear Lake, Minnesota
April 27, 2019
www.nlrs.org/home/aurora

The 2018 Aurora conference will be held April 21, 2019, at the Community of Grace Lutheran Church, 4000 Linden Street, White Bear Lake, MN 55110. Aurora, the largest annual gathering of weak-signal VHF'ers in the Upper Midwest, is the annual gathering of the Northern Lights Radio Society.

In the morning hours the club runs an outdoor antenna range from 0900 to 1130. The morning also allows for socializing, show & tell, and casual demonstrations. And ACØRA will be providing us a live demonstration of 6 meters MSK144. Members are then on their own time for lunch. The technical programs start at 1300, and typically run until 1700 or 1730. If you have a weak signal VHF topic that is of interest to you, or that you would like to present, please contact our Technical Chairman, Jon Platt, WØZQ (**w0zq@aol.com**). If a technical presentation is not to your liking, Aurora also features a poster session.

See website for details — full information will be added to as we draw nearer to the conference date.

VHF Super Conference

Sterling, Virginia
April 26-28, 2019
vhfsuperconference.com

Sponsored by Southeastern VHF Society, North East Weak Signal Group and the Mt. Airy VHF Radio Club. Hosted by the Grid Pirates Contest Group and Directive Systems and Engineering. Conference will be at the Holiday Inn Washington-Dulles International Airport. Please check website; details will be added as available.

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