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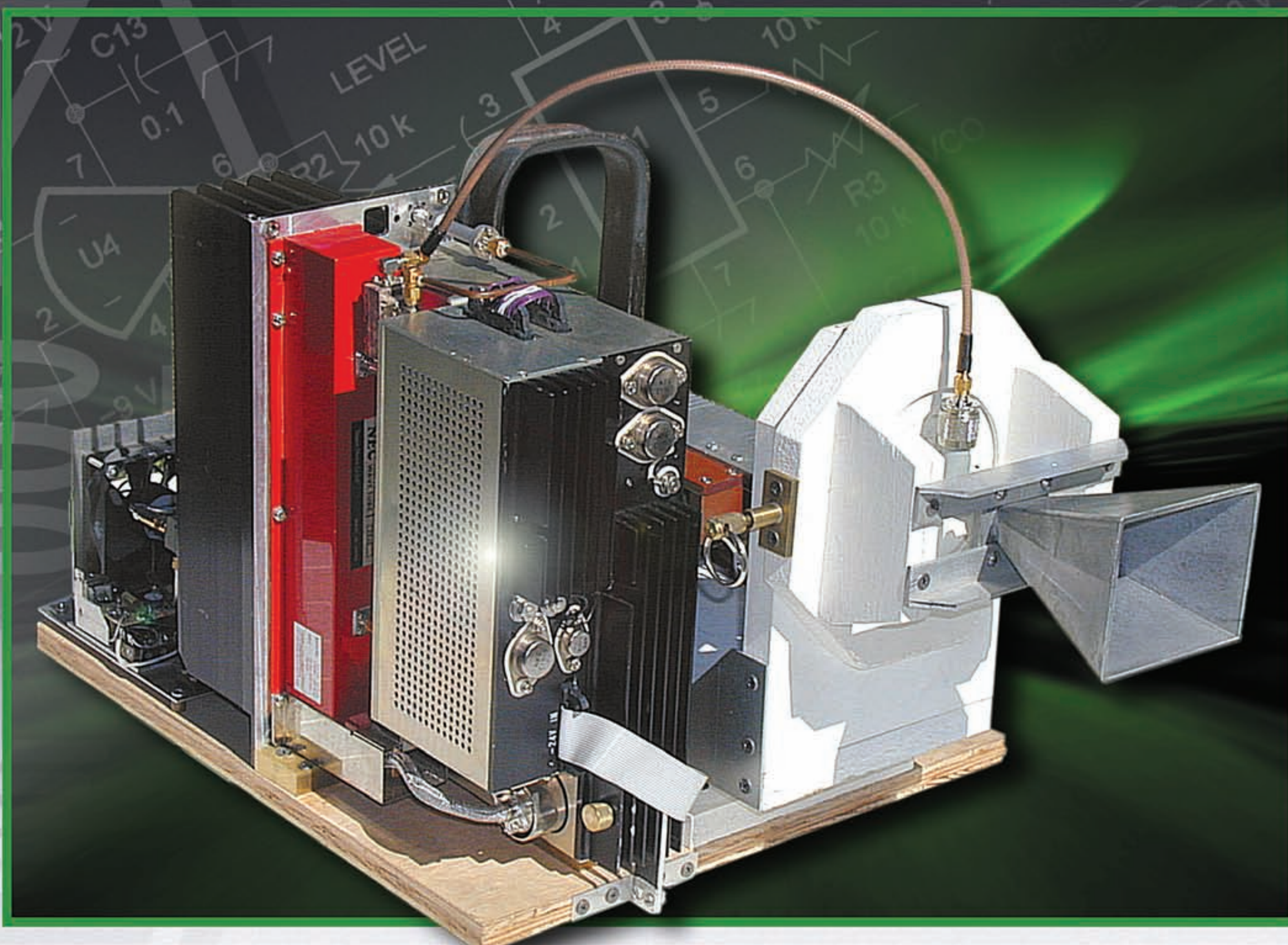
May/June 2009

www.arrl.org

A Forum for Communications Experimenters

Issue No. 254



W6VSV built this phase-locked crystal-controlled 10 GHz signal source and feed horn antenna for testing 10 GHz antennas on his UHF and microwave backyard antenna test range.

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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 254

Harold Kramer, WJ1B
Publisher

Larry Wolfgang, WR1B
Editor

Lori Weinberg, KB1EIB
Assistant Editor

Zack Lau, W1VT
Ray Mack, W5IFS
Contributing Editors

Production Department

Steve Ford, WB8IMY
Publications Manager

Michelle Bloom, WB1ENT
Production Supervisor

Sue Fagan, KB1OKW
Graphic Design Supervisor

David Pingree, N1NAS
Senior Technical Illustrator

Advertising Information Contact:

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

Circulation Department

Cathy Stepina, *QEX Circulation*

Offices

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

Subscription rate for 6 issues:

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

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

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

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

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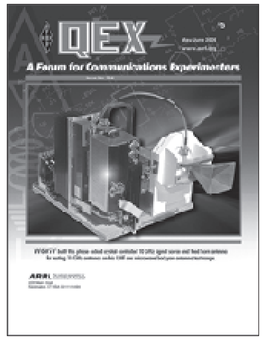
About the Cover

Bob Melvin, W6VSV, built this phase-locked crystal-controlled 10 GHz signal source and feed horn antenna for testing 10 GHz antennas on his UHF and microwave backyard antenna test range. Bob uses separate signal sources and transmitting antennas for each band from 432 MHz through 10 GHz. A computer rotates the test antenna and collects received signal data to plot patterns and calculate gain.



In This Issue

Features

Columns

Index of Advertisers

Larry Wolfgang, WR1B

lwolfgang@arrl.org

Empirical Outlook

New Hams and Upgrades

A couple of things happened recently that started me thinking about the role we all play in recruiting new hams and helping them learn about our passions in Amateur Radio. Sometimes seeds are planted where we least expect.

Around the beginning of the year I received an e-mail from my wife's youngest sister. Joyce wanted to know what she would have to do to earn an Amateur Radio license, and what license she would need to operate "SSB on a boat." This caught me by surprise, since I had never heard Joyce indicate any interest in Amateur Radio. A group of friends are planning to sail a private sailboat to Bermuda this summer, and invited her to go along. Joyce learned that there will be an HF ham rig on board the boat, so she started thinking about being able to operate that radio during the trip.

Of course I obtained a copy of *The ARRL Ham Radio License Manual* and *The ARRL General Class License Manual*, and shipped them to her. Joyce began studying for her Technician exam on her own, and a few weeks ago she sent another e-mail saying she was going to be taking the exam that weekend. My wife and I were thrilled with her phone call to tell us that she passed the exam on her first try! She has started studying for her General license, and I am confident she will pass that one in a short time, too.

In the meantime, there are questions. "Can I use a computer and *EchoLink* rather than a 'real radio' to start with?" (The Bermuda trip isn't free!) "What are the license restrictions to using *EchoLink* to connect to other stations?" "Do I have to know what frequency they are on before I can call them?" And so on.

I have not tried *EchoLink* myself at this point, so I may have to learn something new to be the most helpful, although I suspect Joyce is way ahead of me on this. I'll probably learn more from her on the topic than she from me! In fact, when I called to talk to her about some of the questions and admit that I was not all that familiar with the details, Joyce told me she has already registered her call sign to be able to access the nodes with her computer. I told her that I have talked to stations that were logged on via *EchoLink* to repeaters I was using, and that I've had some interesting DX chats on 2 m. I also pointed out that I have heard stations connect to one of our local repeaters, although the link does not seem to be available all the time. Talking with Joyce over the local repeater, and hearing her talk with my wife will be fun!

One Saturday afternoon recently, my friend Jeff called. He was quite excited and when I calmed him down enough to understand what he was saying, I learned that he had just passed his Amateur Extra class exam! Jeff's enthusiasm reminded me of how I felt when I passed my Extra exam so many years ago. Have you experienced that feeling, and do you remember it? It's quite a high, isn't it?

I've known Jeff for about 15 years or so. He had talked with me about his desire to become a ham many times. In fact, I had given him several copies of ARRL's *Now You're Talking!* over the years. I always offered to help, and to answer any questions, but life, as they say, intervened. Eventually I encouraged Jeff to join one of the Technician license classes I was helping the local radio club to teach. That was just the help he needed, and Jeff was soon a proud Technician. Not one to be satisfied with the limited operating privileges of the Technician license, Jeff was soon studying for his General license. I heard him ask many questions over the local repeater, and answered a few of those questions myself.

Jeff has really thrown himself into Amateur Radio, learning about and operating APRS, becoming very active in our local ARES group and participating in many public service events. With his new Extra class privileges, he is looking forward to being an active Volunteer Examiner for all exams.

I had a call from a ham in Ohio who is also involved with Scouting. He wanted some pointers on what to say at his District Roundtable, where he will be giving a presentation to his local Scout leaders, inviting them to bring their Cub Scouts and Boy Scouts to a local Field Day operation. He is also looking forward to being involved with the Jamboree on the Air in October. Obviously, he and the other hams who are helping him will have an opportunity to expose a large number of young Scouts to the thrill of Amateur Radio. Will they all become interested and earn a license? Probably not. There is a good chance, though, that a few of them will see something that sparks their interest. Even if they don't immediately start to earn a license, it may be something they will think about some day, and remember how much fun they had with those hams.

Have you been planting seeds? I hope so! You might even learn something from your new recruits.

Bob Melvin, W6VSV

1503 Black Mountain Rd, Hillsborough, CA 94010; rmelvin4@comcast.net

The W6VSV "Backyard Antenna Test Range"

Meaningful measurements can be made on UHF and microwave antennas in a backyard, with a carefully designed test setup.

As a ham dedicated to UHF and microwaves, I wanted to be able to measure the performance of the various antennas that I had built. I thought that it would be nice to be able to plot antenna gain patterns, measure maximum forward gain, front to back (F/B) ratio and –3 dB beamwidth.

Figure 1 is a block diagram of the system I built to satisy this desire. It includes a computer-controlled "Backyard Antenna Test Range," built on my wooded, hillside property. This project has developed over several years. UHF and microwave groups have met here for the purpose of pattern plotting and recording other measured values for each antenna brought for testing. Hams have come from several states to test their antennas.

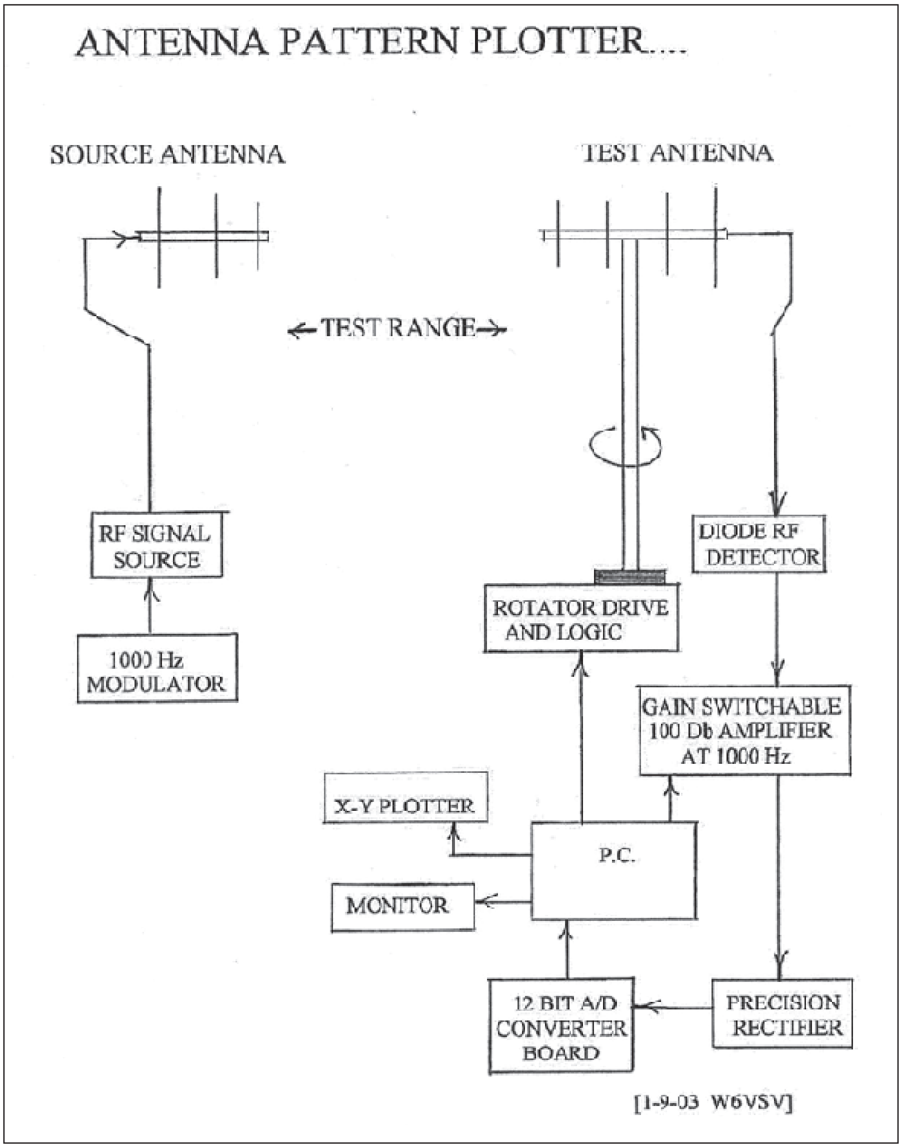


Figure 1 — Here is a block diagram of the equipment setup for my backyard antenna test range.

Hardware

The RF source antennas and the antennas to be tested are set up in the yard to form an antenna test range. An RF source, with antenna at 9 feet above ground, transmits at one end of the yard while the antenna to be tested, mounted on a 9 foot test stand at the other end reads the received signal strength in 625 computer-controlled rotation steps of 0.576° each in a complete circle (360°). The entire backyard setup is displayed in Figure 2.

I built crystal-controlled, solid-state RF signals sources for 432 MHz, 1296 MHz and 10.368 GHz, shown in Figures 3, 4 and 5. Not shown is the crystal controlled signal source at 5760 MHz built by W6DXJ nor the VCO controlled signal sources at 902, 2304 and 3300 MHz built by W6PQL. All sources are modulated at 1000 Hz by switching the RF ouptut on and off at a 1000 Hz rate. Each source has a directive antenna for transmitting its RF signal across the range to be received by the antenna under test, which is mounted

Figure 2 — This photo shows my backyard antenna test range, set up for 1296 MHz antenna testing with the RF source and a 30 inch parabolic source antenna positioned 85 feet from the test stand, which has a 25 element Yagi under test. Note the audio amplifier/loudspeaker on the wall so a tone can be heard across the range while a setup adjustment for tilt of both antennas is being made. An old computer and 4-pen Epson HI-80 digital plotter is visible on the small table. The 100 dB 1000 Hz amplifier is under the large table.

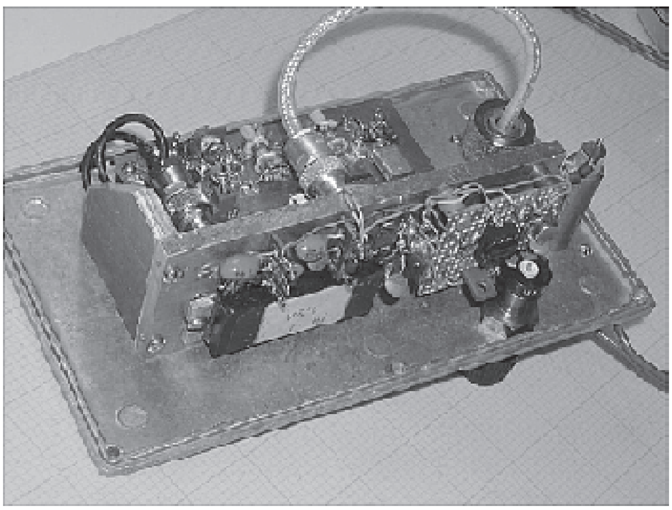


Figure 3 — This is the 432 MHz signal source for antenna testing.

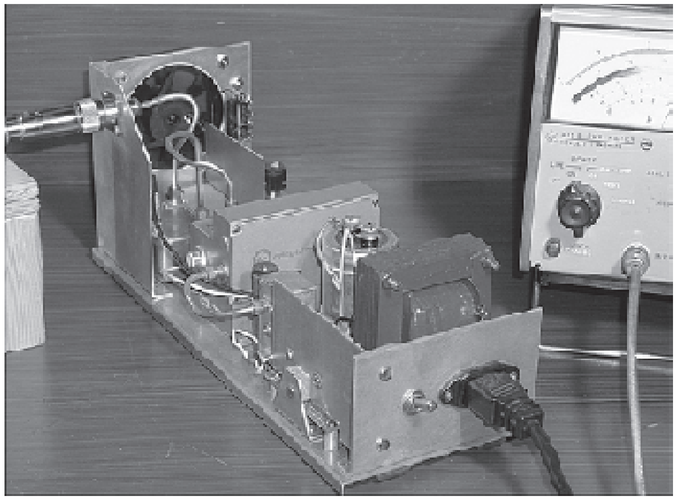


Figure 4 — Here is the 1296 MHz signal source.

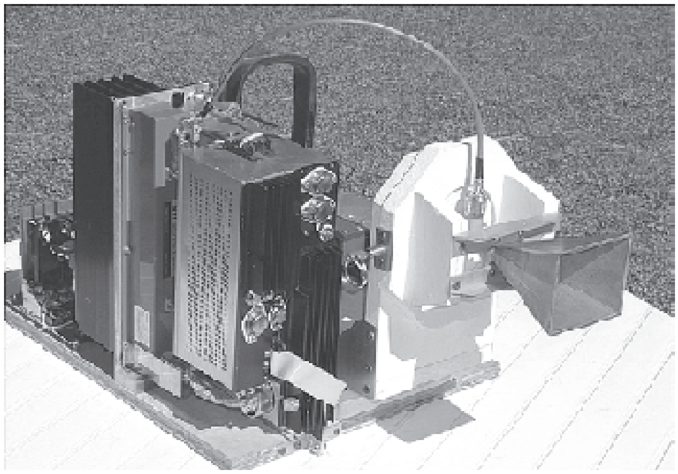


Figure 5 — This photo shows the 10 GHz signal source, along with the feed horn antenna. A 10 mW output phase-locked crystal-controlled signal source drives a TWT amplifier at 10.368 GHz, which is connected to the horn antenna. The horn can be tilt adjusted and rotated 90° for E and H plane plots.

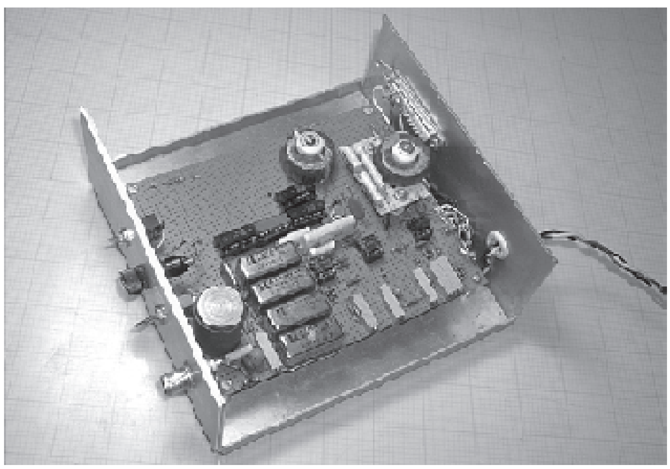


Figure 6 — The homemade high gain 1000 Hz amplifier follows the detector to produce a signal level for the 12bit A/D converter. Four reed relays switch the gain of the 1000 Hz signal into four steps of 20 dB each, with 0.1% resistors. One toroid sets the amplifier of the 1000 Hz signal to provide a –3 dB bandwidth of 55 Hz, while the other toroid is used to filter out 1000 Hz ripple in the rectified dc signal input to the A/D converter.

on the rotating test stand for plotting a 360° pattern. The received signal strength and other desired information is recorded on a chart.

Each band requires a different RF source and transmitting antenna. The receiving setup is the same for all bands.

1000 Hz Amplifier

Perhaps the most difficult portion of this project was capturing the signal strength data over a desired 50 dB range with linear circuits. Recall that 50 dB represents a power ratio of 100,000:1, or a voltage ratio of 316.23:1. This tremendous range of signal strengths would be handled best with logarithmic amplifier circuits. With no experience for such circuitry, I used readily available components and linear amplification.

To accomplish this, I chose to switch the gain of the 1000 Hz amplifier in four steps of 20.0 dB each, by using reed relays for switching voltage dividers built from 0.1% tolerance resistors. Figure 6 is a photo of the amplifier and Figure 7 shows the amplifier schematic diagram.

All program files in this system were written and compiled using the now obsolete Microsoft *QuickBasic*.

Because the RF detector is a square-law device, each of these 20.0 dB steps are computer-converted for display at 10.0 dB steps of antenna gain on the finished plot. When tests are run, 625 readings are taken and plotted in just over a minute. If it is desired to include a forward gain reading on the chart, it is necessary to plot antenna patterns in both horizontal and vertical polarization planes. Both patterns can be plotted on the same chart, and if so, plots will appear in different colors.

Pattern Generation

To record measured patterns, I built low power signal sources and directive antennas for each of the seven bands of interest. The rotating test stand, shown in Figures 8 and 9, is located across the range from the RF sources for the far-field measurements for each antenna to be tested. In my case, this antenna separation distance was limited to 85 feet due to my yard full of shrubs, pine trees and redwood trees beyond the 85 foot limit. With that limitation, far-field patterns are now being measured for all seven ham bands from the lowest frequency of 420 MHz through the highest, the 10 GHz band.

The computer rotates the test antenna by a 200 step stepper motor with a 50:1 gear box, to make 625 rotation steps of 0.576° for each reading in one 360° rotation, This large number of signal strength readings provides sufficient resolution so very sharp pattern peaks and nulls recorded on pattern charts can be clearly seen.

A microwave crystal diode detector is connected to the test antenna and its 1000 Hz output signal is fed into a high gain, narrow bandwidth amplifier. Figure 10 shows the diode detector. Reed relays switch the amplifier gain into four 20.0 db steps. At each step the ac signal is changed to dc by a precision rectifier circuit, to the input of a 12-bit analog to digital converter with its hexadecimal output into the computer for conversion into volts, and volts into dB gain for driving the four pen digital plotter, shown in Figure 11.

Particular attention is necessary when testing high gain antennas having very narrow beamwidths. For these, it is essential, before running tests, to carefully adjust the tilt of the antennas at both ends of the range. A tilt misalingment of a degree ot two can cause a gain error of several dB to be printed on the finished chart.

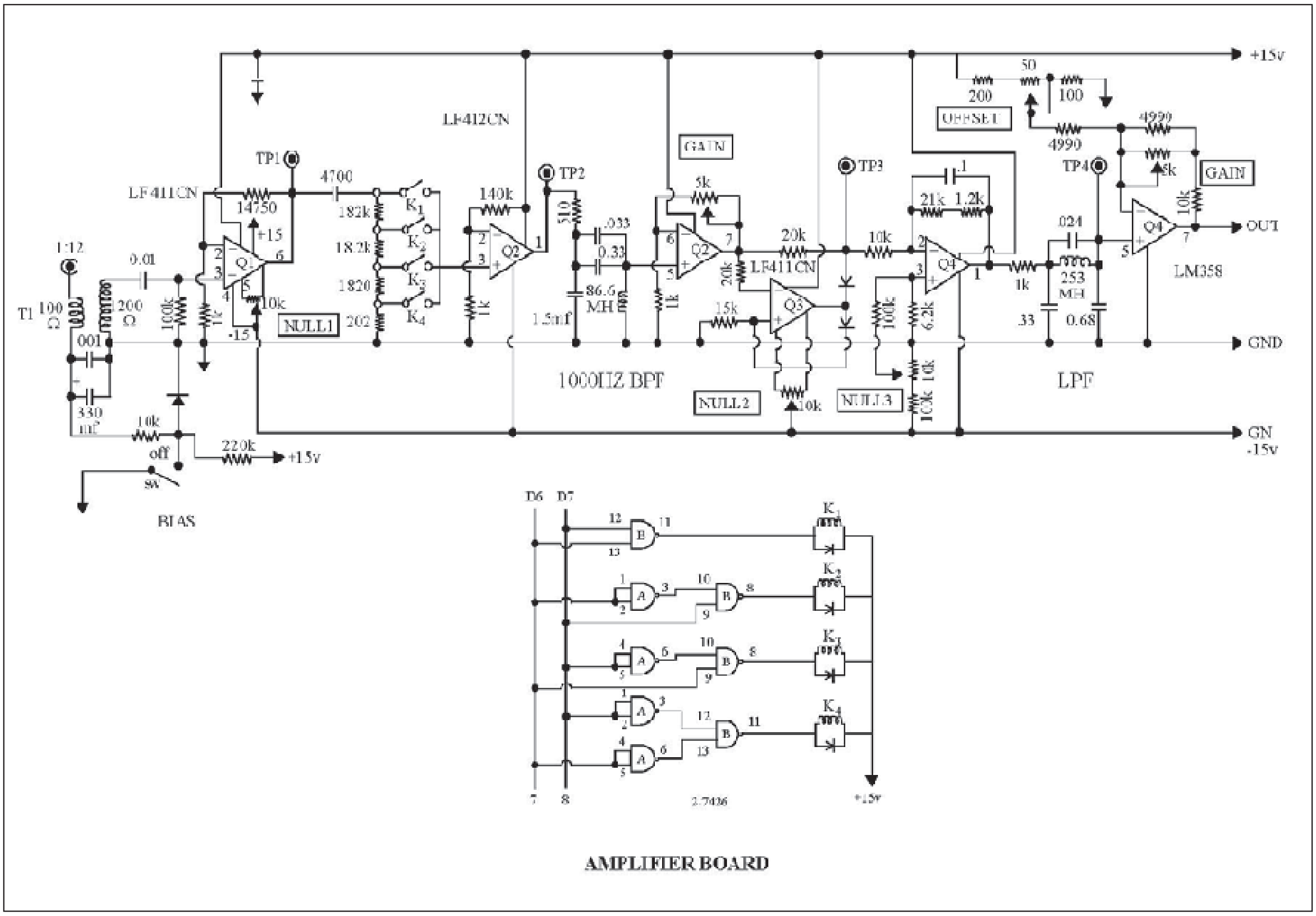


Figure 7 — This schematic diagram shows the circuit for the 1000 Hz amplifier.

Figure 8 — Here is a close-up view of the 9 foot tall rotating antenna test stand. The base frame includes leveling screws to ensure that the antenna is level and plumb. The antenna is rotated by a stepper motor with a 50:1 gearbox. At the top is a structure that can tightly clamp on Yagi booms or many other types of arrays. The clamping arrangement permits easy polarization rotation when making both E and H plane plots.

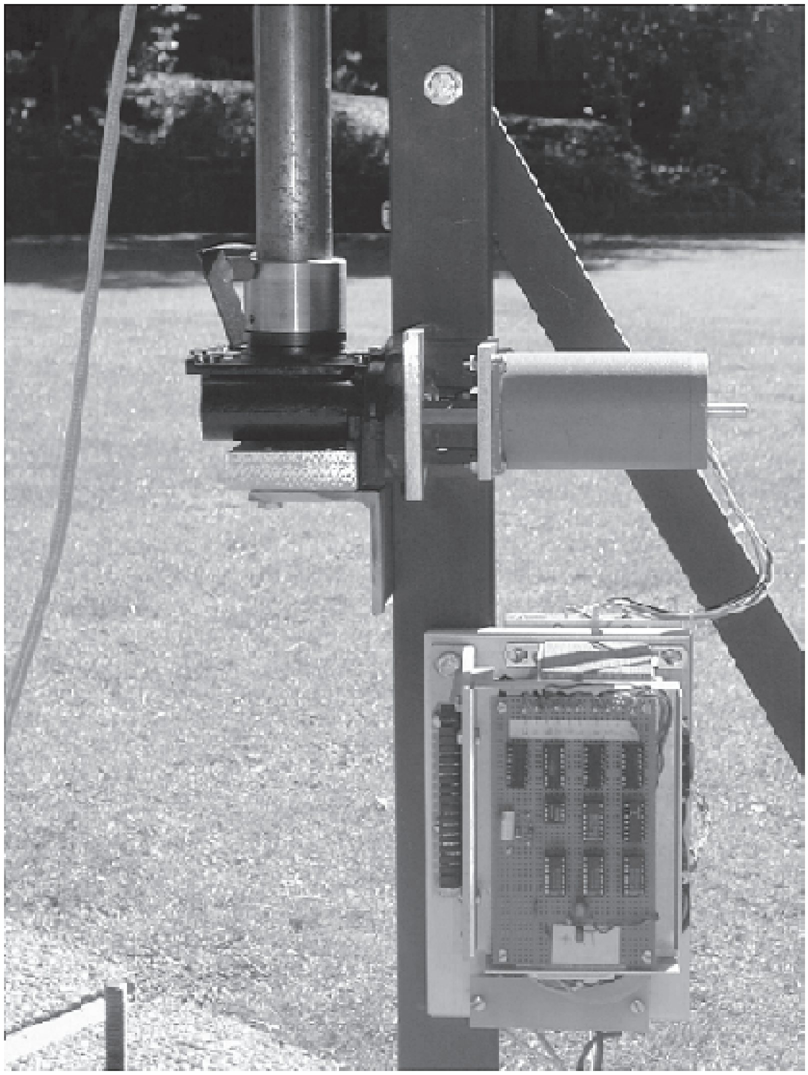


Figure 9 — This photo is a close-up look at test antenna control box with the stepper motor and 50:1 gearbox mounted to the support structure. The stepper motor drive logic circuitry is not shown, since this is not intended as a construction article.

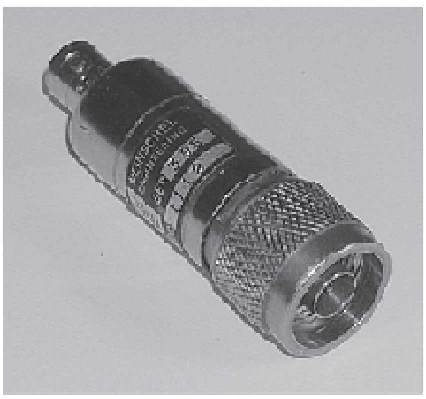


Figure 10 — This microwave crystal diode detector connects to the test antenna, and its 1000 Hz output signal is fed into the high gain, narrow bandwidth amplifier.

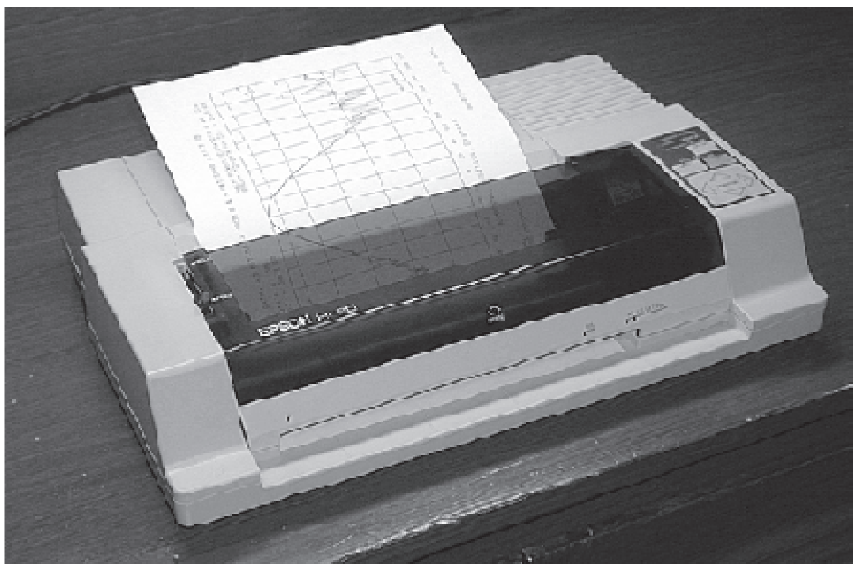


Figure 11 — This four pen Epson HI-80 digital plotter connects to an old computer and is used to create the antenna gain and the E and H plane pattern plots.

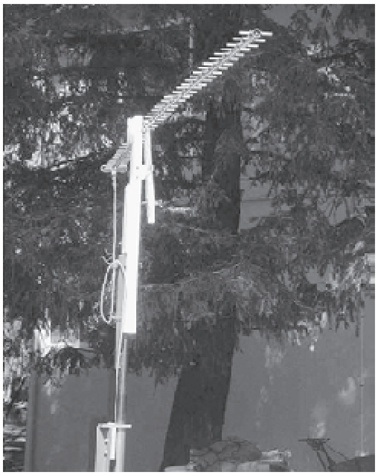


Figure 12 — A KLM 45 element 1296 MHz Yagi is attached to the antenna test fixture.

I have devised a solution to ensure precise antenna tilt adjustments.When the test antenna is mounted, it is rotated towards the signal source and the "Tilt" program is run on the computer. I wrote this program to produce an audio tone that increases in pitch as the incident RF signal increases. This permits manual adjustment of both rotation and tilt of antennas at both ends of the range to produce the highest pitch to be heard from the loudspeaker. The loudspeaker is visible on the wall in Figure 2. By making very careful setup adjustments of direction and tilt, the maximum possible signal level can be received. Once this has been accomplished, the computerized tests can begin.

The computer controls the stepper motor to first rotate the antenna clockwise by 20° from the maximum signal position, then rotate 40° counterclockwise, while searching to detect the exact direction of maximum signal strength. Antenna rotation then continues in the same direction until it is 180° from the direction for maximum signal level. Rotation is reversed for a 360° sweep for plotting the pattern. It can be noted that a pattern will be produced so the maximum forward gain will be written at the top of the plotted polar pattern or at the center of a rectangular-coordinate pattern. When the tests are completed, the antenna is returned to the heading found to be the direction for maxum signal strength.

My rather poor location for a test range is surrounded by trees, shrubs and hillside that can cause unwanted signal reflections to interfere with measurements and distort the patterns produced on the chart. Signal reflections from these objects were greatly reduced by the use of very directive RF source antennas that practically eliminate signal reflection problems.

Forward Gain Measurement

The forward gain for each antenna tested is calculated from the beamwidths that have been measured by making a reading, then rotating the antenna polarizations 90° at both ends and making a second reading. Forward gain is calculated by the computer, by entering both of the measured beamwidths, and using the formula relating forward gain as a function of the two –3 dB beamwidths that is found in Jasik's *Antenna Engineering Handbook*, page 2-14.[1]

[1] Jasik, *Antenna Engineering Handbook*, p 2-14.

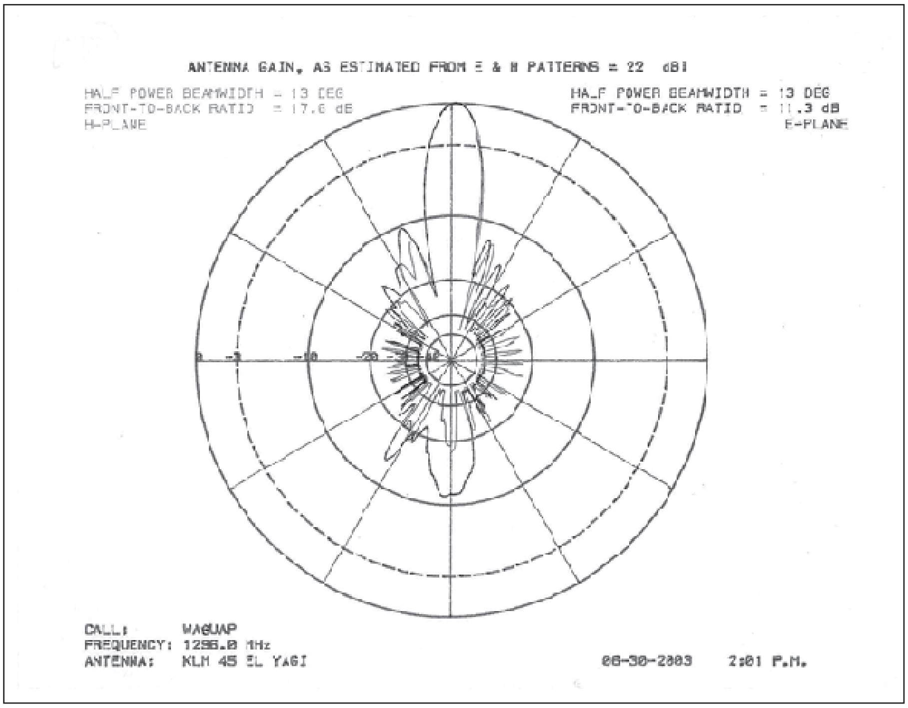


Figure 13 — This pattern plot shows the E and H plane results of measuring the KLM 45 element 1296 MHz Yagi. Note the fine resolution on the side lobes. The back lobe is only down 12 dBc for this antenna. [The original plots show the E plane pattern in green and the H plane pattern in red. — *Ed*.]

Figure 14 — A W6PQL 45 element 1296 MHz loop Yagi is prepared for testing.

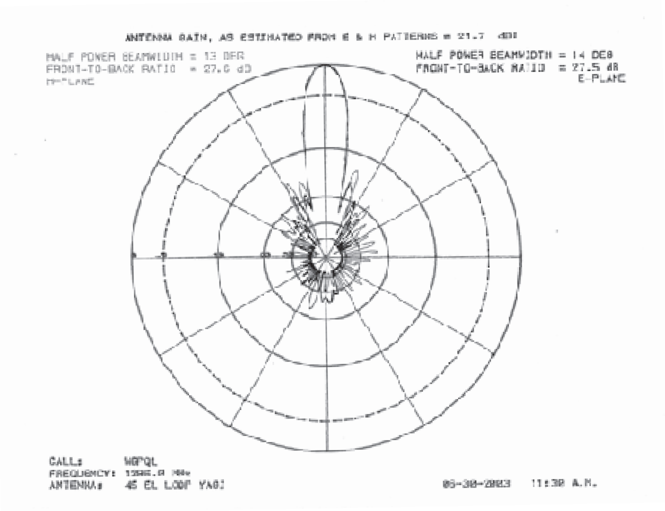


Figure 15 — This pattern plot shows the E and H plane results of measuring the W6PQL 45 element 1296 MHz loop Yagi. Note the greatly reduced side lobes and the back lobe down 24 dBc. Compare this antenna with the KLM Yagi shown in Figure 12 and the measured results of Figure 13.

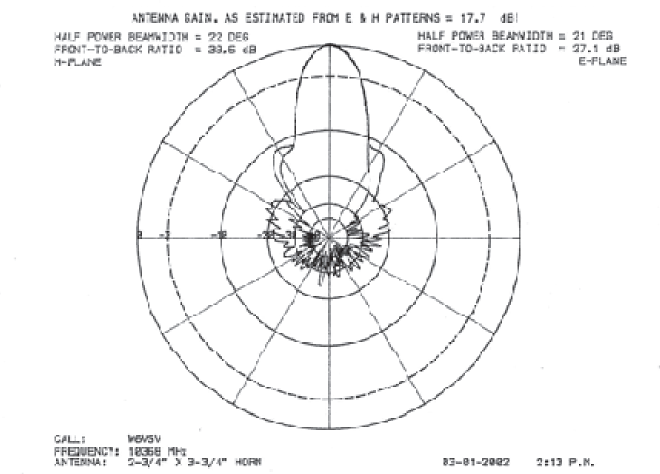


Figure 17 — The X-band radar test horn was measured at 10.368 GHz to produce the E and H plane patterns shown here. This horn now serves as the directive antenna for the 10.368 GHz RF source.

Figure 16 — A WWII X-band radar test horn is placed on the antenna test fixture.

Figure 18 — A W6VSV 18 inch off center fed satellite TV parabolic dish, with a homemade 10.368 GHz dual mode feed horn is attached to the antenna test fixture.

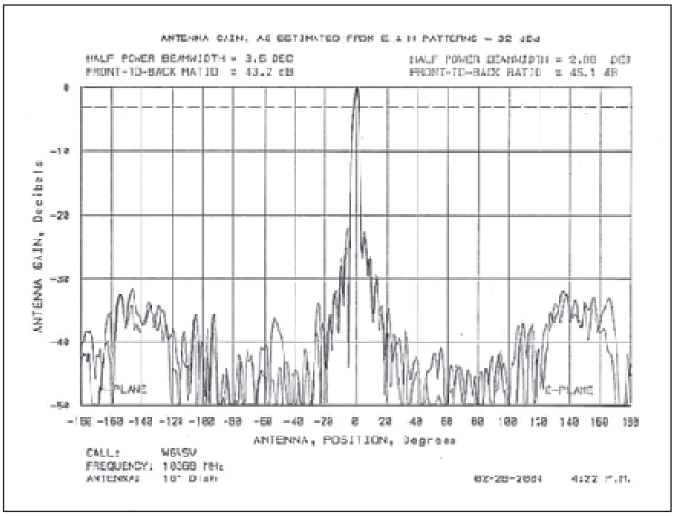


Figure 19 — This rectangular-coordinate plot shows the E and H plane estimated gain for the satellite TV parabolic dish.

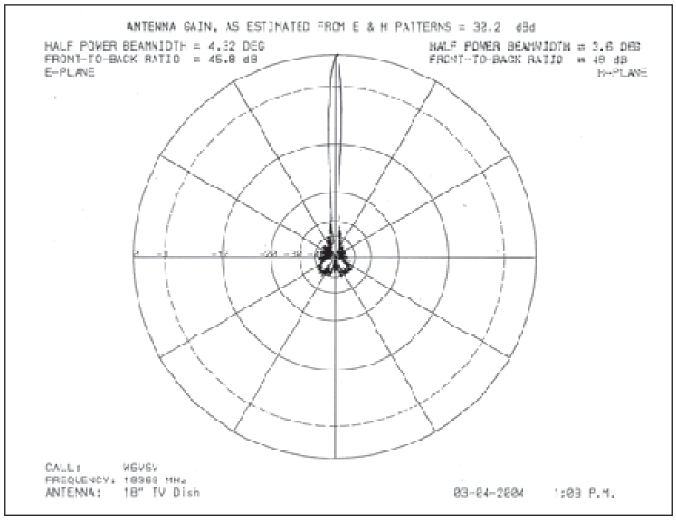


Figure 20 — This polar-coordinate pattern plot shows the E and H plane estimated gain for the satellite TV parabolic dish.

$$G \approx \frac{27{,}000}{\theta_E \theta_H} \qquad \text{[Eq 1]}$$

Note that the gain given here is a gain ratio, compared to an isotropic radiator. If you want to know the gain expressed in decibels, compared to an isotropic radiator (dBi) just take the common logarithm of the result, and multiply by 10.

$$G(\text{dBi}) \approx 10 \log \frac{27{,}000}{\theta_E \theta_H} \qquad \text{[Eq 2]}$$

The computer program I use to calculate the antenna gain based on beamwidth measurements gives a result as gain expressed in decibels compared to a dipole. To calculate that result, take the gain found by Equation 2 and subtract 2.14 dB to read again in dBd.

Problems Encountered

Early on, a problem appeared that was found to be caused by a significant 60 Hz signal in with the 1000 Hz signals. This turned out to be caused by the magnetic field surrounding the ac power wires into the tiny input transformer in the 1000 Hz amplifier circuit following the detector, and into the very high gain amplifier. The ac power lines, even at the street, produced a noticeable pick up. To solve this problem, the input transformer was first encased in a copper shield that acts as a shorted turn to reduce the magnetic pickup, and then that was encased in a thick steel shield machined from a piece of ¾ inch water pipe. You can see this shielded input transformer on the circuit board shown in Figure 2.

No other problems were found in the system. The measurement results have been very satisfactory, and have proven to be very helpful to the many Amateur Radio operators who have tested their antenna arrays with this equipment.

To test the accuracy of the decibel output scale used for the test antenna pattern plots, I used an attenuator to step the 1000 Hz tone in five 20 dB steps. If the detector diode is truly performing as a square-law device, the five 20 dB attenuation steps should appear as five 10 dB steps on the plots. Figure 21 shows the results of this test.

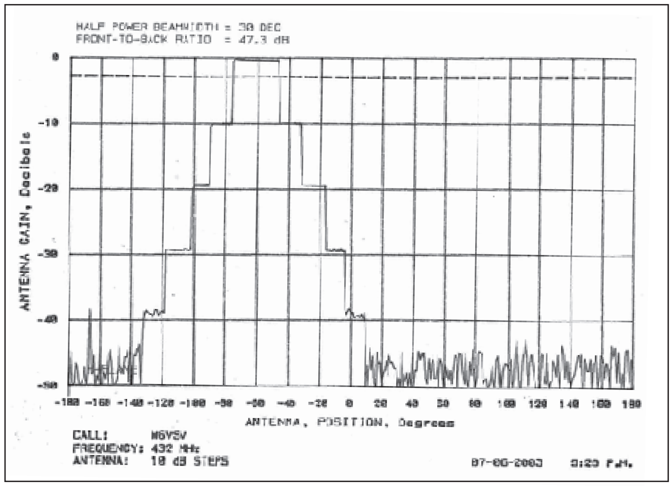


Figure 21 — This rectangular-coordinate plot shows the results of a test to prove that the diode detector is truly performing as a square-law detector. The five 20 dB steps of attenuation appear on the pattern as five steps of 10 dB attenuation.

Results

This article includes only a few of the many antennas that I have tested over several years. The plotted charts for each of these antennas are also included. All of the tests shown here were made before I began using the precise tilt adjust method described in the Pattern Generation section of the article. Therefore, the forward gain listed for narrow beamwidth antennas should actually show somewhat greater gains.

An ARRL Member, Bob Melvin was first licensed as W6VSV in 1946. In 1948, he built a TV camera and transmitter on 423 MHz and transmitted live TV from Berkeley CA from 1949 to 1951. At the end of that time he was transmitting with 200 W peak RF output atop a 1940 foot hill, just outside the Berkeley city limits, and was sending live TV pictures of the San Francisco Bay Area for 200 miles. As another project, he built and transmitted digital audio, then called "Pulse Position Modulation" on the 420 MHz band. The problem in those days, before local TV became available, was that it was extremely difficult to find hams to build receiving gear for reception of either TV or digital audio.

After 3½ years in Electrical Engineering at UC, Berkeley CA, Bob went to work for a Manufacturer's Representative firm in San Francisco for 15 years, selling electronic components in the rapidly growing Northern California territory and then spent 33 years with his own Manufacturer's Representative firm, Melvin Sales, Inc.

Bob once lost his W6VSV license when he failed to submit a timely renewal. He obtained the call KD6SZ and held that until he was able to get his old W6VSV call again. QEX

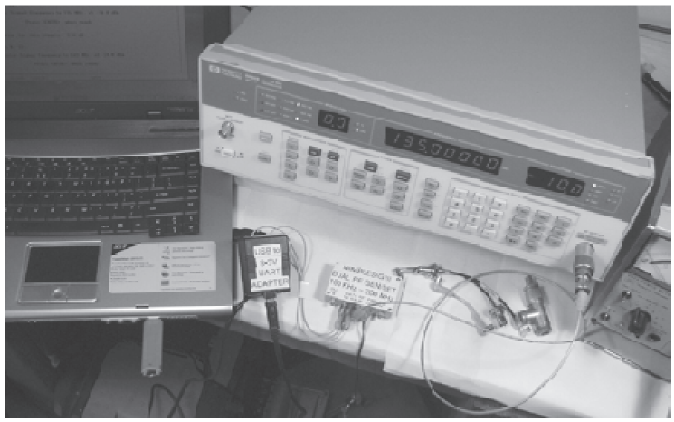
Thomas M. Alldread, VA7TA

7056 Railway Ave, Courtenay, BC V9J 1N4, CANADA; va7ta@telus.net

NimbleSig III—A Dual Output DDS RF Generator and Low Level RF Power Meter—Part 3

Learn about the control software commands and how to calibrate the various sections of NimbleSig III.

Parts 1 and 2 of this article described the hardware design, construction details, software design, computer interfacing, MPU programming and initial testing procedures for NimbleSig III (NS3). NS3 is a frequency agile, direct digital synthesizer (DDS) based RF signal source with independent dual outputs that have a frequency range of 100KHz – 200 MHz. Included within the module is a wide dynamic range RF level meter. Part 3 of this article will describe the NimbleSigIII command set in detail and provide the final calibration procedures.



The NimbleSig III connection daisy chain shows the parallel port JTAG interface connected to the 20 pin side of the cross-connection board and the NimbleSig III 10 pin JTAG programming cable connected to the 10 pin side.

NimbleSig III Command Set

At first glance the nearly 140 NimbleSig III commands might seem like an excessive number to memorize. More than half of the commands are seldom used, however, as they are utility in nature. They are just needed to either directly access DDS registers or to initiate calibration procedures. Thus, there is no need to memorize them. The frequently used commands are duplicated for generators A and B, so they have similar syntax. As there are eight help screens similar to Figure 1 available which display the commands alphabetically a refresher list is close at hand. Of course when NS3 is utilized as an internal signal source module as part of a larger project this somewhat cryptic syntax command set would typically be masked by a more friendly host computer user interface.

The command set is categorized by prefixes as follows:

- "A" Attenuator commands for fine adjusting output levels in 0.1dB steps.
- "B" Output level control mode commands.
- "C" DDS register single bit clear commands.
- "D" EEPROM read/write access.
- "E" Host communications echo control.
- "F" Frequency set commands.
- "H" Help screen commands.
- "K" Kill/restore generator outputs.
- "L" RF level measurement commands.
- "M" Modulation Commands.
- "P" Phase offset commands.
- "Q" Reserved prefix for bypassing a host MPU for direct external PC control.

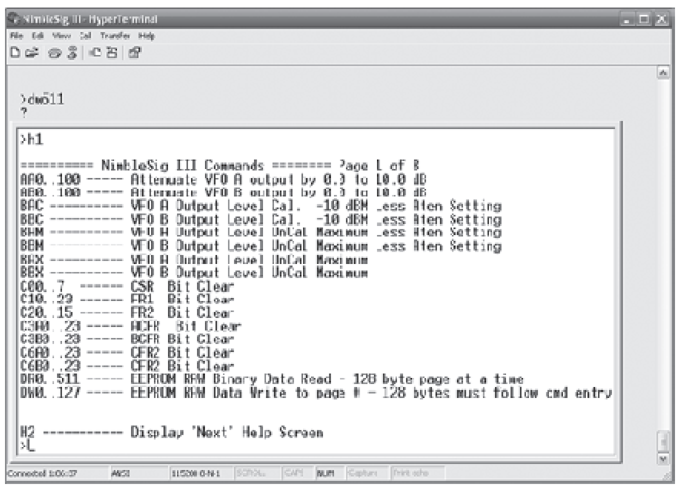


Figure 1 — NimbleSig III Help1 Screen.

[1]Notes appear on page 20.

• "R" DDS register read commands.
• "S" DDS register bit set commands.
• "W" DDS register full word set commands.
• "X" NimbleSig III calibration functions.
• "Y" Re-initialize DDS to default values.

"A" Commands (Fine Step Attenuator Control):

AA0..100 — Attenuate VFO A output by 0.0 to 10.0 dB
AB0..100 — Attenuate VFO B output by 0.0 to 10.0 dB

The format of the "A" command is typical of most of the commands that have numeric suffixes and thus the "A" command is by coincidence a good example to describe first. The "A" command provides a means to reduce the output level of either generator in 0.1dB steps, up to a maximum of 10.0 dB. The "AA" command is for generator "A" and similarly the "AB" command applies to generator "B". The 0..100 suffix indicates the desired level reduction in tenths of a decibel. Thus to reduce the output of generator "A" by 5.5 dB one would enter "AA55" followed by the "Enter" key (note that all commands must be completed with "Enter").

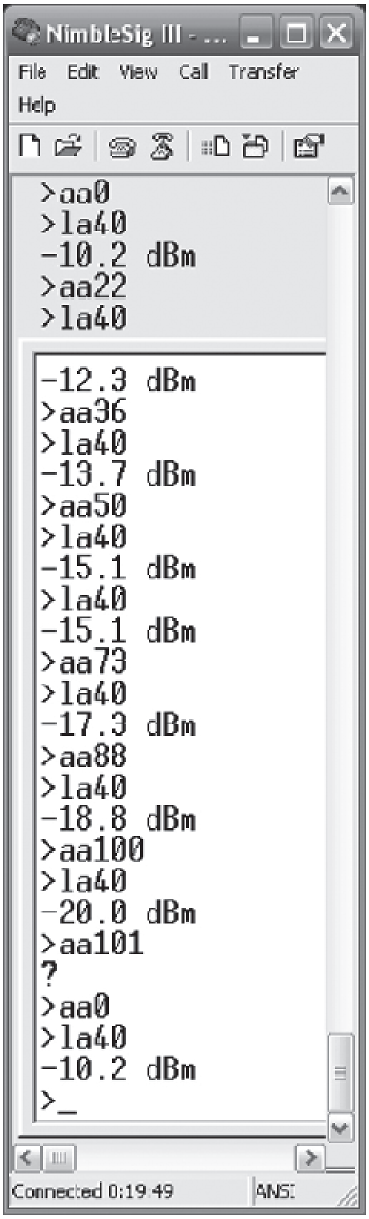


Figure 2 — "AA" command example illustrates typical NimbleSig III command entry and associated level changes.

The HyperTerminal window shown in Figure 2 illustrates some examples of the "AA" command followed by level measurements to show the resulting level shifts. The 40 MHz output of generator "A" was looped into the RF level detector input during this procedure. Note that the ">" is the command prompt sent from NimbleSig III and also note that NS3 accepts lower case alpha characters and converts them to upper case. The "aa0" entry shown at the top (in the HyperTerminal buffer zone) is entered to set the output level to the nominal –10.0 ± 0.5 dBm. The following "la40" command, which repeats throughout this example, is the command for measuring and displaying the average of 1024 power level readings. The 40 suffix is provided to select the 40 MHz calibration factor for correcting the level reading for the frequency of measurement. As shown the output with zero attenuation is measured at –10.2 dBm which is close to the expected –10.0 dBm level. I then decided to set the generator "A" output level to –12.2 dBm by attenuating the generator output by 2.2 dB with the "aa22" command. The resulting output level measurement was –12.3 dBm. As shown in the repetitive command sequences that follow, the measured level shifts remain within about 0.2 dB of the desired shift. Since the output of the DDS is controlled with digital accuracy the variation of shift from nominal is most likely due to the very slight inaccuracies of the analog level detector response in combination with the resolution limitations of the 10 bit ADC in the MPU, which is approximately 0.13 dB.

Following the 3rd command entry ("aa50") example in Figure 2, I entered two consecutive "la40" commands to demonstrate the stability of an averaged 1024 sample RF level measurement. Notice that the two readings are exactly the same at –15.1 dBm.

In the second last sequence I entered "AA101" to demonstrate the range checking done by the NS3 firmware. 101 is out of the range specified for the numeric suffix associated with the "AA" command. The acceptable range is shown in the Figure 1 help screen as 0..100 indicating the maximum value allowed is 100. NS3 performs similar range checking on all commands and retorts with a "?" in the event of a command entry error. The last "AA0" entry which completes this example restores the level back to the initial value of –10.2 dBm.

"B" Commands (Output Level Control):

BAC — VFO A Output Level Cal. –10 dBm Less Aten Setting
BBC — VFO B Output Level Cal. –10 dBm Less Aten Setting
BAM — VFO A Output Level UnCal Maximum Less Aten Setting
BBM — VFO B Output Level UnCal Maximum Less Aten Setting
BAX — VFO A Output Level UnCal Maximum
BBX — VFO B Output Level UnCal Maximum

The "B" commands provide a means to obtain the maximum possible output level from the DDS generators by removing the calibration factor that is used to flatten the output level to approximately –10 dBm across the whole output frequency range. The maximum output level across most of the output range is about –4 dBm however this rolls off as the corner frequency of the associated low pass filter is approached. The maximum output level can be useful for applications where the maximum dynamic range or extra signal power is more important than a calibrated absolute level. For example the maximum output level would usually be desirable for sweeping the response of a relatively narrow passive bandpass filter. In this case the output level from the DDS would typically be sufficiently flat across the bandwidth of interest without level correction and the extra generator level could be useful for extending the dynamic range of the measurement. Here the relative loss between the passband and the rejection bands would be the criteria of importance whilst the absolute level used for the measurement would be of little interest.

There are three pairs of "B" commands provided. "BAM" and "BBM" switch generator A and generator B respectively to maximum level with the "A" command (previously described) left active to provide a means to step the level down in 0.1 dBm increments. The "BAX" and "BBX" commands simply run the respective generator output levels at maximum irrespective of the "A" command setting. The "BAC" and "BBC" commands restore the respective generator output levels to the initial default –10 dBm calibrated output operation.

"C" and "S" Commands (Bit Clear and Bit Set):

C00..7	—	CSR Bit Clear	S00..7	—	CSR Bit Set
C10..23	—	FR1 Bit Clear	S10..23	—	FR1 Bit Set
C20..15	—	FR2 Bit Clear	S20..15	—	FR2 Bit Set
C3A0..23	—	ACFR Bit Clear	S3A0..23	—	ACFR Bit Set
C3B0..23	—	BCFR Bit Clear	S3B0..23	—	BCFR Bit Set
C6A0..23	—	CFR2 Bit Clear	S6A0..23	—	CFR2 Bit Set
C6B0..23	—	CFR2 Bit Clear	S6B0..23	—	CFR2 Bit Set

The NS3 firmware is designed to provide the user full control of the AD9958 DDS register contents. Some of the DDS registers use individual bits to configure, set modes or in general control the device

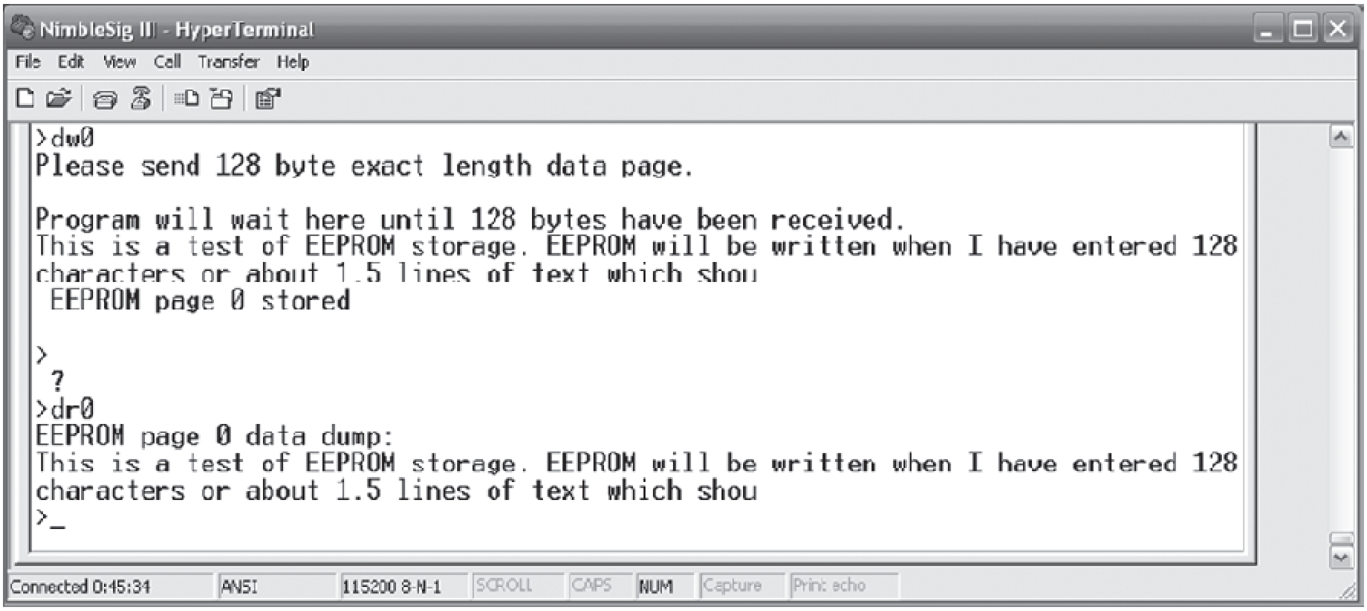


Figure 3 — Non-volatile EEPROM storage "DW" and "DR" command examples.

operation. Other registers are word registers that, when changed, require that the whole word be replaced. Thus there is need for two types of register content adjustment commands – "single bit toggle" or "word replacement." The "C" and "S" commands are used to toggle single bits whilst the "W" command described below has been provided for the replacement of full register words.

NS3 permits the changing of all the control bits except those bits that are deemed unchangeable by the DDS data sheet or those bits that could cause the MPU-DDS control communications to fail.[1] Any attempts to change the bits critical to the operation and control are ignored by the NS3 firmware.

The "C" command is a single bit command used to clear single bits in select DDS registers. The "S" command, which is provided for setting single bits, is the complement to the "C" command. As described in detail within the AD9958 data sheet the CSR, FR1, FR2 and CFR registers use single bits to control DDS configurations and functions. (See Note 1.) Thus there are "C" and "S" commands for each of these registers. Note there are separate CFR and FR2 registers for each DDS generator thus there are "A" and "B" selection suffixes added to the "C" and "S" commands for unique selection of these registers.

The second character in the syntax for the "C" and "S" commands is numeric and it refers to the register address within the DDS register bank. The remaining numeric suffixes for the "C" and "S" commands point to the bit to be changed. Note the length of these AD9958 control registers varies from 8 to 24 bits thus the maximum value of the suffix varies accordingly. The following are examples of the "C" and "S" commands:

"C122" would clear bit 22 in register FR1 which is at DDS register bank address 1.

"C3A14" would clear bit 14 in the generator A, CFR register which is at address 3 of the "0" bank.

"s06" would set bit 6 in the CSR register which is at address 0 in the DDS register bank.

"C01" which points to bit 1 of the CSR register would do nothing as this bit must be left in the one state as it configures the serial communications to the DDS.

"S3B2" which points to bit 2 of the generator "B" CFR register would also do nothing as according to the DDS data sheet this bit must be left in the zero state.

Refer to the "**Register Map**" section of the AD9958 data sheet for the bit function assignments, address assignments and other detailed information on the DDS register bank which contains a total of 45 registers.

"D" Commands (EEPROM Data Read/Write):

DR0..511 — EEPROM RAW Binary Data Read - 128 byte page at a time

DW0..127 — EEPROM RAW Data Write to page # - 128 bytes must follow cmd entry

The "D" commands permit the storing/recalling of non-volatile data to/from the EEPROM. This can be useful for initialization data such as last used frequencies or channel frequency assignments. Although the intent is to provide non-volatile storage for a host controller this feature could be useful for storing any data. However since the EEPROM device is organized with page lengths of 128 bytes any entries must also have a length of exactly 128 bytes. Thus one must use fill characters as needed. Please note the entire destination page will be overwritten.

The "DW" command permits the non-volatile storage of data to any one of the first 128 pages of the EEPROM. Each page is 128 bytes long thus 16 KB of external data can be stored. The remaining 384 upper pages are reserved for internal use. However the data they contain can be dumped one page at a time with the "DR" command (described below). Please note as mentioned above the EEPROM access is restricted to one full page at a time thus to save just one byte 127 additional fill bytes must be written. To save more than 128 bytes the data must be broken up into 128 byte segments and the last segment appended to 128 bytes as necessary. In the process the data must be saved one page at a time.

The "DR" command may be used to dump any of the 512 pages of data, one page at a time. Please also note that the data dumped is in binary format thus if there are no ASCII equivalents the characters may not print or may appear as garble on a typical ASCII terminal screen. The binary data may be viewed properly with the very powerful and excellent terminal program called *RealTerm* that is available free on the Internet.[2] *RealTerm* can be used to display the data in hexadecimal format as shown in Figure 4. Note that in this example the NimbleSig user friendly verbose mode (to be described next) was previously disabled, which prevented the human interface frill "EEPROM data page dump" text title from being sent first. In cases of nontext data, such as program initialization variables that often need to be stored in nonvolatile memory, the ability to read only the 128 bytes of raw data from the EEPROM page without any extra clutter is often desirable.

"E" Commands (Echo control):

EA	—	Enable All Human Interface functions
ED	—	Disable Character Echo
EE	—	Enable Character Echo
EP	—	Enable Prompt
EQ	—	Disable Verbose Mode
ES	—	Enable Space Bar Command Repeat
ET	—	Disable Space Bar Command Repeat
EV	—	Enable Verbose Mode
EX	—	Disable All Human Interface functions
EZ	—	Disable Prompt

Although the plain language messages and unsolicited prompts are usually very desirable when directly controlling NS3 from a keyboard they can get in the way and slow down communications to a host controller. In some cases the echo should be limited to just the prompt which could be utilized as an indicator that NS3 is ready for the next command. In other cases even the prompt character string might disrupt the host which may not always be able to tolerate the incoming prompt messages. The "E" commands provide a means to disable some or all of the human interface text statements for streamlining machine-to-machine communications. These commands, which are listed on the help screen, are pretty much self explanatory. The "EX" command may be used to disable all the human interface frills while "EA" may be used to re-enable them all. With the human interface message frills disabled, NS3 will still respond to all valid commands and respond with data when appropriate.

"F" Commands (Frequency Entry):

FAH1.200000000	—	VFO A Freq = # Hz
FAK1..200000	—	VFO A Freq = # kHz
FAM1..200	—	VFO A Freq = # MHz
FBH1.200000000	—	VFO B Freq = # Hz
FBK1..200000	—	VFO B Freq = # kHz
FBM1..200	—	VFO B Freq = # MHz
FXH1.200000000	—	VFO A/B Freq = # Hz
FXK1..200000	—	VFO A/B Freq = # kHz
FXM1..200	—	VFO A/B Freq = # MHz

There are three groups of the "F" frequency entry commands with three flavors each. The "FA" group is for entering generator "A" frequencies, "FB" for generator "B" and "FX" for entering the same frequency into both generators with a relative phase offset defined by the phase offset register values. The "H" suffix specifies frequency entry in hertz, the "K" suffix in kilohertz and the "M" suffix in megahertz. For example the commands:

- "FAH201015" — would set the generator "A" frequency to 201,015 Hz
- "FBK455" — would set the generator "B" frequency to 455 kHz
- "FAM146" — would set the generator "A" frequency to 146 MHz
- "FXM10" — would set the frequencies of both generators "A" and "B" to 10 MHz with a defined relative phase offset.

"K" Commands (RF Output Kill/Restore Control):

KA	—	Kill VFO A Output
KB	—	Kill VFO B Output
KK	—	Restore VFO A Output
KL	—	Restore VFO B Output

The "K" commands are used to toggle either output off and on.

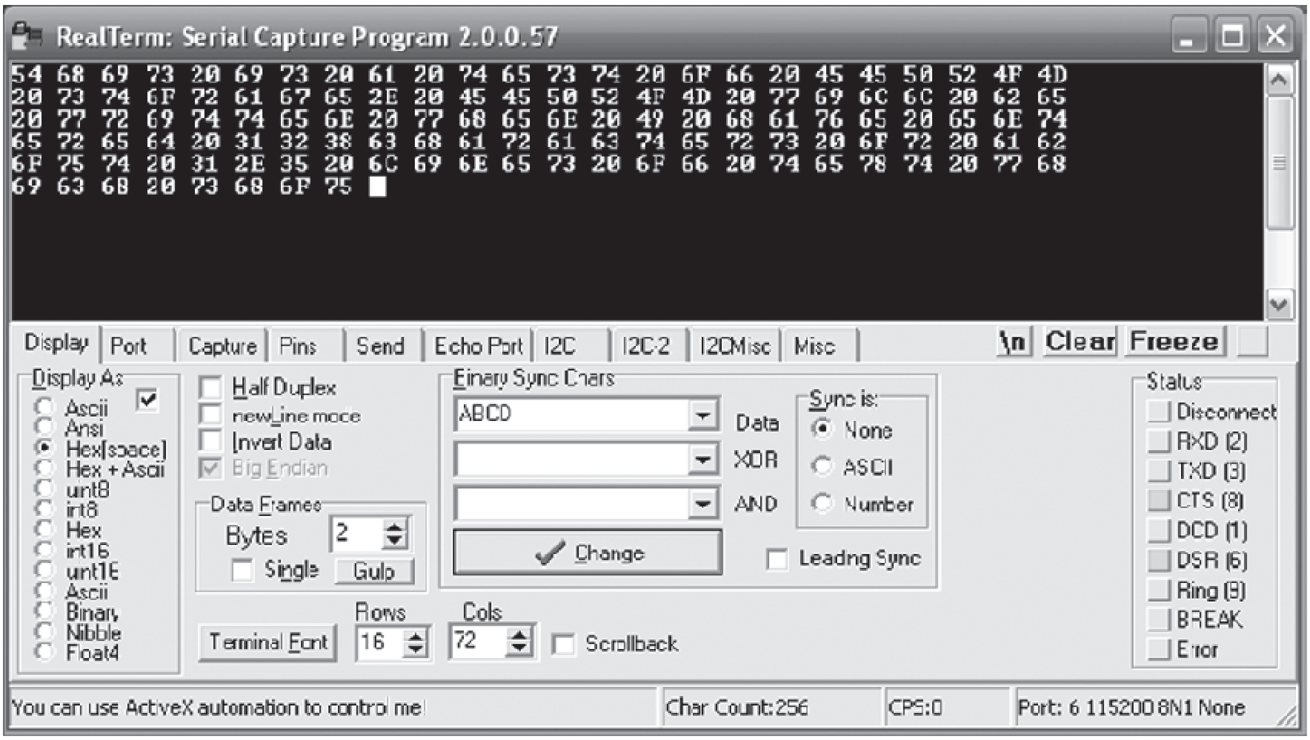


Figure 4 — RealTerm HexDump of EEPROM page 0 text shown in Figure 3 with Verbose mode disabled.

"L" Commands (RF Power Level Measurement):

LA1..500	—	RF Detector Average Level 1024 samples
LC1..500	—	RF Detector Average Level 128 samples
LL1..500	—	RF Realtime Detector Level 1 sample
LM1..500	—	RF Detector Minimum Level 1024 samples
LP1..500	—	RF Detector Peak Level 1024 samples

The suffix for all of the "L" commands is provided to specify the frequency of the signal being measured to the nearest MHz. The value given is used for the selection of the appropriate frequency response calibration correction factor. If the suffix is left blank the measurement will be flagged as uncalibrated.

The "LA" command averages 1024 samples taken over a period of about 60mS and responds with the output level in dBm. The "LC" command is similar to the "LA" command but averages 128 samples taken over a period of about 0.75mS. This command is intended for swept frequency response measurements where faster, but reasonably well averaged measurements are needed. The LL command provides the level of a single sample which maybe useful for sampling the amplitude of a signal for digital processing. The "LM" and "LP" measure the minimum and peak power of a signal over a period of 1024 samples taken in 60 mS. The two measurements could be used to measure the percentage of amplitude modulation. The "LP" command could be used to measure the peak envelope level of an SSB signal source.

"M" Commands (Modulation Control):

MA	—	vfoA AM Mod. On
MB	—	vfoB AM Mod. On
MD1..100000	—	FM Deviation 1 Hz-100 kHz
MF	—	vfoA FM Mod. On
MG	—	vfoB FM Mod. On
MP1..99	—	AM Mod. Depth 1-99
MQ1..20000	—	Mod. Freq. 1 Hz – 20 kHz
MX	—	Modulation Off

The "M" commands, which provide modulation control, are pretty much self explained by the help page information as shown above. Note that modulation is only permitted on one channel at a time. If one channel is currently modulated and a command request is sent to modulate the alternate channel, the modulation is first turned off and then applied to the most recently requested, alternate channel.

"P" Commands (Phase Offset Control):

PA0..360000	—	Absolute Set VFO A phase offset millidegrees
PB0..360000	—	Absolute Set VFO B phase offset millidegrees
PP	—	Print VFO A phase Offset in degrees
PQ	—	Print VFO B phase Offset in degrees
PS22..10000	—	Set A/B phase step size – millidegrees
P+A	—	Increment VFO A phase offset by one step
P+B	—	Increment VFO B phase offset by one step
P-A	—	Reduce VFO A phase offset by one step
P-B	—	Reduce VFO B phase offset by one step

The "P" commands are provided for controlling the relative phase of the two generator outputs after the generator pair has been set to the same frequency with the "FX" command. The "FX" command initially places the two generators into relative phase with each other in accordance with the values within the phase offset registers. Thus if the generator A offset is set at 90 degrees and the generator B offset is set at 0 degrees when the "FX" command is implemented then generator A output will lead generator B by 90 degrees following the "FX" command execution. Consequently the "FX" command can be used to change the frequency whilst keeping the phase offset between the two RF outputs constant.

I decided to use millidegrees for the phase offset entry "PA" and "PB" command suffixes to provide an easy data entry means for harnessing the 22 millidegree step resolution, the finest step resolution offered by the AD9958. The entry in millidegrees is rounded off to the nearest step. The "PP" and "PQ" commands provide a means to interrogate the current phase offset values (in degrees decimal format) for either generator. The reply format is in conventional, easy-to-read degrees, decimal degrees notation.

The "P+A"/"P-A" and "P+B"/"P-B" command pairs provide a means to tune the offset in steps. The size of the steps is defined by the "PS" command and is entered in millidegrees. The step size can be varied from 22 millidegrees to 10 degrees. These commands are useful for optimizing the phase offset for the best results. For example, in the case of a phasing type of single sideband generator, one could use these commands to find the optimum phase offset for the best cancellation of an opposite sideband.

"R" Commands (Register Read):

RA0..24	—	VFO A DDS Register Read in Hex
RB0..24	—	VFO B DDS Register Read in Hex
RX0..24	—	Current VFO DDS Register Read in Hex

The "R" commands are provided for reading the DDS register values. The displayed value is in hexadecimal format. The numeric suffix refers to the internal address of the register to be read within the DDS register address map described in the AD9958 datasheet. (See Note 1.) As the first three registers, addresses 0 to 2, are common to both generators, the "RA0"/"RA1"/"RA2" and "RB0"/"RB2"/"RB3" commands would respectively access the same registers. For the higher addresses in the range of 3 to 24 the "RA" and "RB" commands uniquely select the individual registers within the banks assigned to generator A and generator B respectively. As described in the data sheet the selection of register bank 0 or register bank 1 (corresponding to generators A or B respectively) depends on the setting of steering bits in the CSR register at address 0. The "RX" command is provided to read the currently selected register bank without changing the steering bits in the CSR register. This can be useful for determining which register bank is currently selected as the "RX" command does not disturb the steering bits.

"S" Commands (Register Bit Set):

S00..7 — CSR Bit Set

The "S" commands are described above with the complementary bit clear "C" commands.

"W" Commands (Write Register Word):

WA0H0..FF	—	ACSRW0 Hexadecimal Word Set
.		
.		
WBOH0..FFFFFFFF	—	BPROF15 Hexadecimal Word Set

The 48 variations of the "W", "write register word" commands, permit one to write any DDS register with a full data word with a bit length equal to the size of the target register. As described in the "C" command section above any attempt to change bits critical to the control and operation of the DDS will be blocked. Otherwise any register content can be changed. To determine the desired register data to be written by this command one should refer to the "Register Maps" section on page 36 of the AD9958, "Rev. A"(July 2008) data sheet. (See Note 1.)

The generator A or B register bank is selected with the second character of the command.

The register address within the generator A or B register bank is selected with the third character. This third character is assigned numeric values in the range of '0' to '9' for pointing to the first ten registers which are described within tables 28 and 29 of the AD9958 "Rev. A" data sheet. To target the profile registers (data sheet table 30) which have addresses from 10 to 24 the third character is assigned an alphabet character in the range 'A' to 'O' as defined in the associated help page list.

The binary data to be written must be entered in the hexadecimal format with the 'H' prefix. Thus the fourth character of the command is always an H as the prefix identifier for the hex format. The length of the hexadecimal data must match the length of the target register.

For example to write the value 19,462,008(128F778 hex) to the generator B, profile register 10 one would first look at the NS3 help page 7 (by entering the “h7” command at the NS3 prompt) which lists the command prefix for the BPROF10 register as “WBJH”. This prefix would be followed with the 128F778 hexadecimal data value to be written. However note that a leading 0 must be added to make the value match the length of the 32 bit profile register which requires 8 hex character words to be fully defined.

(Note that a hex character represents 4 bits, thus it takes 8 hex characters to make up a 32 bit word. To write decimal number 10 to an NS3 32 bit register in hexadecimal, one must enter seven leading zeros, resulting in “0000000A” hex.)

For this example the required command syntax is: WBJH0128F778

“X” Commands (Setup and Calibration):

XA — CALIBRATION - VFO A Freq Response
XB — CALIBRATION - VFO B Freq Response
XC — Set DDS Reference Clock Nominal Frequency in Hz
XD — Dump Power Meter Freq Response Cal Factors
XE — Activate New EEPROM with Approximate Power Meter Data
XF — CALIBRATION - Power Meter Freq Response
XHz — CALIBRATION - Generator Frequency
XI — DDS Register Initialization SetUp
XK — Activate New EEPROM with Approximate VFO A/B Data
XL — CALIBRATION - Log Detector Amplitude Response
XP — Dump Log Detector Amplitude Response
XR — Dump DDS Registers All Current Data

The “X” commands are used to perform the initial setup and calibration (or recalibration) of the NimbleSig III module. The use of these commands will be described in detail in the EEPROM initialization and calibration sections below.

“Y” Command (Restore DDS Settings):

Y — Re-Initialize the DDS to EEPROM stored values

In case of difficulties, the “Y” command can be used to restore the DDS to the initialized state.

“Space Bar Repeat”

The space bar may be used to repeat any previous command. This is very useful for phase tuning and for the calibration procedure, which can be quite repetitive. Striking the space bar repeatedly can increment or decrement a value repeatedly as defined by the previous command.

EEPROM Initial Programming

As the NimbleSig III firmware needs initialization values from EEPROM during start up or after reset it will not function properly if the EEPROM is blank. The only time this will normally happen is when a new NimbleSig III module is run for the first time. It is thus necessary to pre-program the EEPROM to default values prior to calibrating a new module for the first time. The XC, XE, XI and XK commands are provided for this purpose.

Enter the “XC” command which will ask for the nominal DDS clock frequency which normally is 500 MHz. This frequency must be entered in Hz. When requested enter 500000000 (a number five followed by eight zeros). The software will acknowledge that the clock value is saved.

Next enter the “XE” command and confirm that you wish to proceed. Again the software will acknowledge the saving of approximate calibration data.

Next enter the “XI” command and confirm that you wish to proceed with a “c” when requested followed by “19” to load the default register settings. Again the software will acknowledge the saving of approximate calibration data.

Finally enter the “XK” command and again confirm you wish to proceed. And again acknowledgement of data saved will be given for approximate calibration of both generators.

Once the above EEPROM initialization steps are completed, restart the NS3 module by either power cycling or reseting. This re-start is needed to load the default data from the EEPROM into the MPU data memory.

For confirmation enter the “XR” command which provides a screen dump of the DDS registers. A screen dump similar to that shown in Figure 15 should follow. Note values for registers A and B at addresses 4h should be 147AE148h and 1999999Ah respectively. The values for registers at 6h are output level calibration sensitive thus may vary but they should be in the range of 1100h to 1200h. There should be 40 and 50 MHz signals present at the outputs of generators “A” and “B” respectively at levels close to –10 dBm. The “LA10” power level command should report a level less than –50 dBm. If both generator

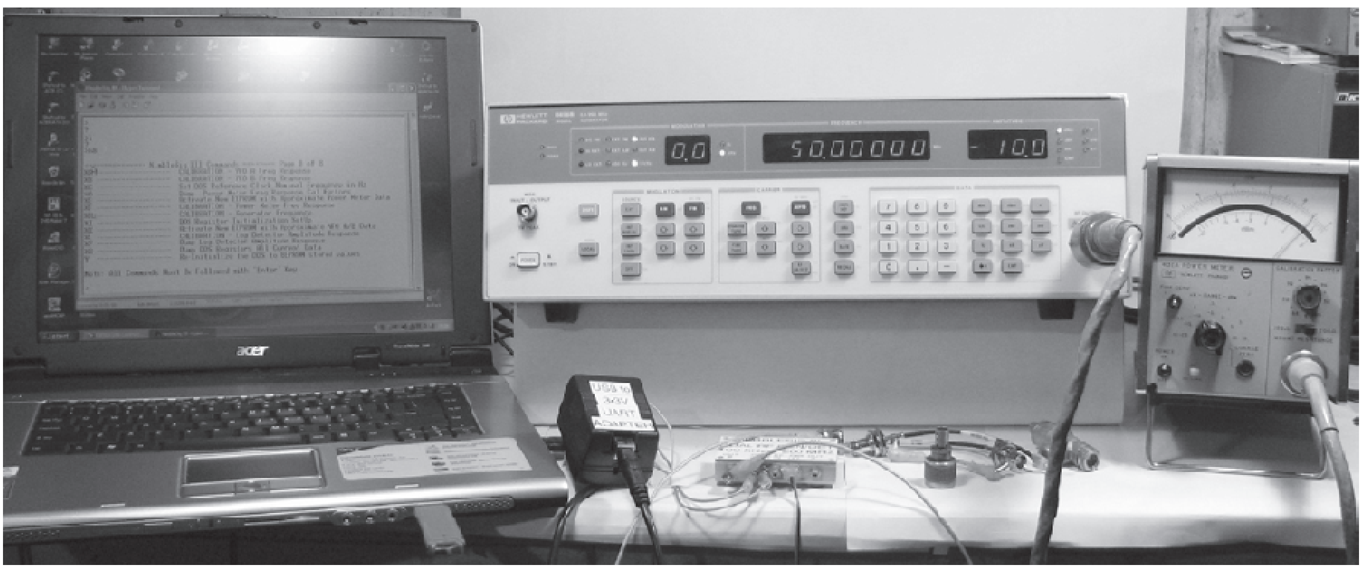


Figure 5 — Calibration check of HP8656B signal generator.

outputs and power meter indications are normal, plus the screen dump compares favorably to Figure 15 the module is ready for calibration.

Calibration Procedure

Figure 5 is a view of my signal generator being checked for output level accuracy against my RF power meter. I had the good fortune of being able to acquire and successfully repair this surplus HP8656B lab-quality signal generator. It works well for performing the NS3 calibration procedures. The HP8656B permits one to increment or decrement either the output level or frequency by a specified amount with just the single push of a button which is certainly a convenient feature for calibrating an NS3 module. As an attempt to confirm the absolute level accuracy of my generator I decided to measure the output level with my rather old HP432 power meter. As shown in Figure 5 the power meter measured the 50 MHz, –10 dBm output level of the signal generator at –10.5 dBm. When two instruments with different calibration lab history match this close I think one can be fairly confident in the accuracy. I am not sure which instrument is closest but considering the difference was minor and my generator is decades newer than my power meter I decided to place my trust in the generator.

```
>XL
NimbleSig IIIRF Power Detector Calibration procedure
!!! CAUTION !!! - this will change instrument calibration !!!
Be prepared to inject RF signals at known levels into RF power detector input
 Press 'c' to continue - any other key to abort.
c

Set 10 MHz generator output level to pre-defined calibration step amplitude

Power Level Step # 1: Set RF Power Detector input level to -30.00 dBm
  Press <ENTER>  when ready

Power Level Step # 2: Set RF Power Detector input level to -28.00 dBm
  Press <ENTER>  when ready

Power Level Step # 3: Set RF Power Detector input level to -26.00 dBm
  Press <ENTER>  when ready
...
...
...
Power Level Step # 14: Set RF Power Detector input level to -4.00 dBm
  Press <ENTER>  when ready

Power Level Step # 15: Set RF Power Detector input level to -2.00 dBm
  Press <ENTER>  when ready

Power Level Step # 16: Set RF Power Detector input level to 0.00 dBm
  Press <ENTER>  when ready

Power Meter Amplitude Response Calibration Values Table

 Value    Power    Slope
(Volts)   (dBm)   (mV/dB)
1.431    -30.00
1.479    -28.00    24.2
1.528    -26.00    24.2
1.579    -24.00    25.8
1.631    -22.00    25.8
1.685    -20.00    27.4
1.740    -18.00    27.4
1.792    -16.00    25.8
1.840    -14.00    24.2
1.892    -12.00    25.8
1.943    -10.00    25.8
1.995    -8.00    25.8
2.050    -6.00    27.4
2.104    -4.00    27.4
2.156    -2.00    25.8
2.204    0.00    24.2

A = -85.100487
B = 0.124075

Press 'y' if you wish to SAVE this calibration - any other key to discard it.
Y
NimbleSig IIIRF Detector Amplitude Response Cal Saved to Non-Volatile Memory!
```

Figure 6 — NimbleSig III power detector logarithmic response calibration dialog.

Power Detector Logarithmic Response Calibration

The NS3 calibration is done by first calibrating the internal power detector. Once calibrated the power detector will later be used for automatically determining the frequency dependent output level response flattening factors for the two generators. There are two calibration procedures that must be done for the power detector. Both require the reference signal generator output to be connected to the detector input. First the logarithmic response of the detector is calibrated at 10 MHz from –30 dBm to 0 dBm in 2 dB steps. Figure 6 is an abbreviated capture of the dialog which occurs during the calibration procedure.

The logarithmic response calibration sub routine is started with the "XL" command which, as shown in Figure 6, first requests confirmation from the operator that this procedure is in fact desired. Upon confirmation the procedure requires that the reference signal generator be set to 10 MHz at –30 dBm. Once the input signal is set up accordingly the operator simply presses the <Enter> key then NS3 makes the level measurement and moves on to the next step with a request for a 2 dB increase in signal level. Once the level is increased <Enter> is again pressed and the process repeats until all 16 steps are completed at which point the level will have reached 0 dBm. When this last signal level setting is measured, the calibration subroutine displays, for review purposes, the table of values acquired during the procedure. Upon confirmation from the operator the new calibration values are then saved to the EEPROM for non-volatile storage.

The specified nominal output slope for the AD8307 is 25 mV per dB. As shown in Figure 6, in this case the steps varied from 24.2 mV to 27.4 mV. Most of this variation is probably caused by the resolution of the 10 bit A-D converter used here which is (3.3 V / 1024) = 3.22 mV per step. This corresponds to about 0.13 dB, or on a power scale 3% per step, which is the resolution limit for this power meter. This is usually good enough for practical purposes, but I think if I were to do this project over again I would look for an MPU with a 12-bit ADC, which if set up with a 2.5 V reference could offer 0.025 dB or roughly 0.5% power scale resolution. Nevertheless, the accuracy as show in Figure 7 seems more than adequate for most purposes.

Figure 7 shows the dynamic range accuracy obtained when I checked this unit against the step attenuator of my signal generator. At the top end of +10 dBm the detector appears to be on the verge of satu-

Input Level dBm	NS3 Read-ing	Error dB	Input Level dBm	NS3 Read-ing	Error dB	Input Level dBm	NS3 Read-ing	Error dB
+10	+9.5	-0.5	-18	-17.9	+0.1	-46	-45.7	+0.3
−8	+7.7	-0.3	-20	-20.0	0.0	-48	-47.7	+0.3
−6	+5.7	-0.3	-22	-22.1	-0.1	-50	-49.8	+0.2
−4	+3.7	-0.3	-24	-24.1	-0.1	-52	-51.7	+0.3
2	+1.7	-0.3	-26	-26.1	-0.1	-54	-53.9	+0.1
0	0.0	0.0	-28	-27.9	+0.1	-56	-55.7	+0.3
-2	-1.9	+0.1	-30	-29.8	+0.2	-58	-57.7	+0.3
-4	-3.9	+0.1	-32	-31.8	+0.2	-60	-59.6	+0.4
-6	-6.0	0.0	-34	-33.9	+0.1	-62	-61.7	+0.3
-8	-8.1	-0.1	-36	-35.9	+0.1	-64	-63.8	+0.2
-10	-10.1	-0.1	-38	-37.9	+0.1	-66	-65.6	+0.4
-12	-12.0	0.0	-40	-39.8	+0.2	-68	-67.4	+0.6
-14	-14.0	0.0	-42	-41.7	+0.3	-70	-68.9	+1.1
-16	-15.9	+0.1	-44	-43.7	+0.3	-72	-70.0	+2.0

Figure 7 — Power detector 10 MHz dynamic range accuracy.

ration with a 0.5 dB error. From +8 dBm down to –58 dBm, the unit is within 0.3 dB of the step attenuator. It remains within 0.4 dB down to –66 dBm (about ¼ of a nanowatt) below which the noise floor starts to contribute. At –72 dBm the error is a full 2 dB, indicating a large proportion of the power being measured is noise.

Note that since the power meter was calibrated at 10 MHz, it will be most accurate in the HF part of the spectrum. According to the AD8307 specification sheet, the dynamic range logarithmic tracking varies somewhat with frequency. Additionally the detector loses sensitivity as the frequency increases. Although the frequency calibration to be done next compensates for the sensitivity drop off, the net effect is that the power meter is about 10 dB less sensitive at 500 MHz than at the HF end of the spectrum.

Power Detector Frequency Response Calibration

The next step is to calibrate the frequency response of the power detector by entering the "xf" command. I decided to calibrate the detector all the way up to 500 MHz which is well beyond the frequency range of the two generators. Please note that although the detector input SWR should remain at less than 1.5:1 up to 200 MHz at 500 MHz the SWR will deteriorate to about 2:1. This may affect the measurement accuracy within the UHF range. At a cost of input sensitivity, the UHF input mismatch could be improved with a 6 dB attenuator connected directly to the input connector. As mentioned above, the detector sensitivity rolls off by about 10 dB at 500 MHz and although the internal NS3 firmware calibration factors correct for this roll off the sensitivity of the detector is correspondingly less at higher frequencies. Thus don't expect to be able to measure levels below about –50 dBm at 500 MHz without a preamplifier.

The frequency response measurement is done in 10 MHz increments starting at 5 MHz, thus there are 50 steps to this procedure by the time 495 MHz is reached. This multi-step procedure is similar to that for the dynamic range logarithmic response calibration except in this case the level is left constant and the frequency is stepped. The NS3 procedure dialog is shown in Figure 8. The generator output level is first set to –10.0 dBm and the frequency is set to 5 MHz. Then step-by-step the dialog asks the operator for frequency increments of 10 MHz until 495 MHz at step 50 is reached.

Once step 50 is completed the frequency calibration values table shown in Figure 9 is generated for review. If the operator chooses to accept the calibration, the new calibration values are saved in the external EEPROM. This completes the calibration of the NS3 power detector.

RF Generator Frequency Response Level Calibration

The frequency response flattening procedure is done automatically which, unlike the power meter calibration, requires very little operator intervention. All the operator needs to do is connect the generator outputs, one-at-a-time, to the power detector input as requested by the sub routine. Then the embedded NS3 software completes the calibration.

The dialog for the generator A calibration procedure is shown in Figure 10. After entering the "xa" command and providing confirmation to proceed, the operator must ensure the Generator A output is connected to the power detector input. Then after the <Enter> key is pressed the NS3 firmware will step the Generator A frequency, in 1 MHz increments, across the 200 MHz spectrum. At each step it will hesitate to measure the maximum output level from the DDS engine and then calculate the required level reduction factor needed to produce the desired –10 dBm output level for the current 1 MHz frequency segment. As this process proceeds, the measured output levels scroll down the terminal program screen which provides the operator a view as shown in Figure 10 of the calibration in progress.

Note that as shown in Figure 10 the last five measurements, in this case, fall below –10 dBm which indicates that this particular

```
>xf
      NimbleSig III RF Power Detector Frequency Response Cal procedure
     !!! CAUTION !!!  -  this will change instrument calibration !!!
Prepare to inject 5 to 500 MHz RF sigs at -10.0 dBm into RF detector input.

Press 'c' to continue - any other key to abort.
c

Set RF signal generator unmodulated output level to -10 dBm amplitude

Frequency Step #  1:
Set RF Generator Signal Frequency to 5 MHz, at -10.0 dBm
                       Press <ENTER>  when ready
Calibration factor for this step is:  0.3 dB

Frequency Step #  2:
Set RF Generator Signal Frequency to 15 MHz, at -10.0 dBm
                       Press <ENTER>  when ready
Calibration factor for this step is:  0.5 dB
.
.
.
Frequency Step #  49:
Set RF Generator Signal Frequency to 485 MHz, at -10.0 dBm
                       Press <ENTER>  when ready
Calibration factor for this step is:  9.9 dB

Frequency Step #  50:
Set RF Generator Signal Frequency to 495 MHz, at -10.0 dBm
                       Press <ENTER>  when ready
Calibration factor for this step is:  10.2 dB
```

Figure 8 — Power detector frequency response calibration dialog.

```
Power Meter Frequency Response Calibration Values Table
Freq. Factor Freq.  Factor  Freq. Factor  Freq. Factor  Freq.  Factor
(MHz)  (dB)  (MHz)   (dB)   (MHz)  (dB)   (MHz)  (dB)   (MHz)  (dB)
  5 ...0.3    15....0.5     25....0.8     35....1.0     45....1.2
 55 ...1.4    65....1.7     75....1.9     85....2.0     95....2.3
105 ...2.6   115....2.6    125....2.4    135....2.6    145....2.7
155 ...2.9   165....3.1    175....3.2    185....3.4    195....3.7
205 ...4.0   215....4.1    225....4.2    235....4.5    245....4.7
255 ...5.0   265....5.1    275....5.4    285....5.5    295....5.7
305 ...6.0   315....6.2    325....6.4    335....6.6    345....6.9
355 ...7.0   365....7.3    375....7.5    385....7.8    395....8.0
405 ...8.2   415....8.4    425....8.5    435....8.8    445....9.0
455 ...9.3   465....9.4    475....9.6    485....9.9    495....10.2

Press 'y' if you wish to SAVE this calibration - any other key to discard it.
y

NimbleSig III RF Detector Frequency Response Cal Saved to Non-Volatile Memory!

>h8
```

Figure 9 — Power detector frequency response calibration factors table.

```
>xa
      NimbleSig III RF Generator A Frequency Response Cal procedure
     !!! CAUTION !!!  -  this will change instrument calibration !!!
Prepare to loop Generator A output into RF power meter detector input.
Press 'c' to continue - any other key to abort.
c

Loop signal generator A output to RF power detector input
Starting Generator A Scan and Calibrate
-2.2 dBm*
-2.2 dBm*
-2.2 dBm*
.
.
.
-9.9 dBm*
-10.3 dBm*
-10.5 dBm*
-10.9 dBm*
-11.3 dBm*
-11.4 dBm*
```

Figure 10 — Generator "A" frequency response calibration dialog.

module is not capable of generating the desired –10 dBm level above 195 MHz. This is mainly due to the roll off of the low pass filter which may vary slightly from unit-to-unit due to component tolerances. I decided the output level of –11.3 dBm at 200 MHz is still very usable, and thus worthwhile to support with the firmware. It should be kept in mind however that the generator output levels may fall below the flatness calibration target of –10 dBm ±0.5 dB above 190 MHz. Note that tests have now been completed on a new 230 MHz LPF design which resolves this upper corner limitation. The design details along with test results for the new filter can be found on my Web page. I thank the ARRL technical review staff for suggesting this design improvement. Note that the parts list on the NS3 Web site provides ordering information for the new 230 MHz filter design. (See Note 2.)

Once this ~Frequency Step and Calibrate~ automatic sequence completes, the calibration factor table shown in Figure 11 is generated. This table of values lists the algebraic level adjustment that is needed for each 1 MHz segment up to 200 MHz.

The output levels of the DDS generators are adjusted by changing the associated Amplitude Scale Factor (ASF) which sets the maximum output voltage from the associated Digital to Analog Converter (DAC) within the AD9958 chip. In this case this is a 10 bit value thus the scale maximum is 1023. Upon displaying the calibration factor table of Figure 11 the calibration procedure pauses and prompts the operator prior to displaying the ASF factor percentage table shown in Figure 12. Note that from 196 to 200 MHz, the ASF is set to 100% which indicates the generator set to maximum output level has reached the end of its calibration range. This corresponds to the 0.0 dB correction factors listed for 196 to 200 MHz in Figure 11. With the new 230 MHz filter design this end-of-range situation has been eliminated as there is still plenty of calibration head room at 200 MHz.

Finally the operator is prompted to accept and save the new calibration values to the EEPROM chip for non-volatile storage.

Next this same procedure is implemented for Generator "B" by connecting the Generator "B" output to the power detector input and entering the "xb" command. This completes the output level calibrating for the signal generator pair.

Reference Clock Frequency Calibration

This procedure involves adjusting the actual DDS output frequency to match a frequency standard such as WWV or a highly accurate frequency counter. More traditionally the reference clock frequency is usually adjusted to achieve the desired synthesized generator output frequency. In the case of a phase lock loop based generator or a frequency counter this is the only option. However a convenience associated with a DDS generator is that one has the option to leave the reference frequency as it is and instead adjust the reference numeric value used by the DDS for calculating the frequency tuning word to match the frequency of the reference clock. For this example I am using the 25 MHz TCXO for the reference which is not adjustable. It is multiplied by 20 with the AD9958 internal PLL to obtain a nominal 500 MHz reference for the DDS engines. However after completing the frequency adjustment I found the actual DDS reference frequency needed to generate exactly 10 MHz was 500,000,120.0 Hz which indicates the actual TCXO frequency is approximately 25,000,006.0 Hz. Please note that although the NS3 firmware will support any nominal reference frequency between 400 and 500 MHz a frequency close to 500 MHz is needed to generate signals as high as 200 MHz.

The first step as shown in the Figure 14 dialog is to enter the "xc" command which will provide a means for setting the nominal design reference frequency. In this example 500000000 is entered for the nominal 500 MHz which provides the starting point.

If you wish to use the AM receiver with WWV as a reference method, tune the receiver to WWV at 10 MHz. Then set the DDS

RF Generator A Amplitude Response Calibration Values Table

Freq (MHz)	0	1	2	3	4	5	6	7	8	9
	Correction Factor (dBm)									
0	-7.8	-7.8	-7.8	-7.8	-7.8	-7.5	-7.8	-7.8	-7.8	-7.6
1	-7.9	-7.9	-7.9	-7.8	-7.8	-7.8	-7.6	-7.6	-7.6	-7.5
2	-7.8	-7.8	-7.6	-7.6	-7.6	-7.5	-7.5	-7.5	-7.4	-7.4
3	-7.5	-7.5	-7.5	-7.4	-7.4	-7.4	-7.3	-7.3	-7.1	-7.1
4	-7.0	-7.1	-7.1	-7.1	-7.0	-7.0	-7.0	-6.9	-6.9	-6.9
5	-7.0	-7.0	-7.0	-7.0	-6.9	-5.9	-6.9	-6.7	-6.7	-6.7
6	-7.0	-6.9	-6.9	-6.9	-6.9	-5.7	-6.7	-6.7	-6.7	-6.7
7	-6.9	-6.9	-6.9	-6.9	-6.9	-5.9	-6.7	-6.7	-6.7	-6.7
8	-6.9	-6.9	-6.9	-6.7	-6.7	-5.7	-6.7	-6.7	-6.7	-6.7
9	-6.9	-6.9	-6.9	-6.9	-6.9	-5.9	-6.7	-6.7	-6.7	-6.7
10	-7.0	-6.9	-6.9	-6.9	-6.9	-5.7	-6.7	-6.7	-6.7	-6.6
11	-6.6	-6.6	-6.6	-6.5	-6.5	-5.4	-6.4	-6.4	-6.4	-6.2
12	-6.1	-6.0	-6.0	-6.0	-5.9	-5.9	-5.7	-5.7	-5.7	-5.6
13	-5.7	-5.6	-5.6	-5.5	-5.5	-5.3	-5.3	-5.2	-5.1	-5.1
14	-5.1	-5.1	-5.0	-5.0	-4.8	-4.7	-4.7	-4.6	-4.6	-4.5
15	-4.6	-4.6	-4.5	-4.5	-4.3	-4.2	-4.2	-4.1	-3.9	-3.9
16	-3.9	-3.9	-3.8	-3.7	-3.7	-3.6	-3.4	-3.4	-3.3	-3.3
17	-3.3	-3.2	-3.2	-3.1	-2.9	-2.9	-2.8	-2.7	-2.5	-2.4
18	-2.7	-2.4	-2.4	-2.3	-2.0	-1.9	-1.8	-1.7	-1.4	-1.3
19	-1.3	-1.1	-0.9	-0.6	-0.4	-0.1	0.0	0.0	0.0	0.0
20	0.0									

Figure 11 — Generator "A" frequency response flattening adjustment values table.

Press <ENTER> to display Amplitude Scale Factors

RF Gen. A Amplitude Response Calibration ASF() Table

Freq (MHz)	0	1	2	3	4	5	6	7	8	9
	Amplitude Scale Factor (ASF) Percentage of Maximum									
0	40.9	40.9	40.9	40.9	40.9	40.9	40.9	40.9	40.9	41.4
1	40.3	40.3	40.3	40.9	40.9	40.9	41.4	41.4	41.4	42.0
2	40.9	40.9	41.4	41.4	41.4	42.0	42.0	42.0	42.7	42.7
3	42.0	42.0	42.0	42.7	42.7	42.7	43.3	43.3	44.0	44.0
4	43.3	44.0	44.0	44.0	44.6	44.6	44.6	45.3	45.3	45.3
5	44.5	44.6	44.6	44.6	45.3	45.3	45.3	45.9	45.9	45.9
6	44.5	45.3	45.3	45.3	45.3	45.9	45.3	45.9	45.9	45.9
7	45.3	45.3	45.3	45.3	45.3	45.3	45.3	45.9	45.9	45.9
8	45.3	45.3	45.3	45.9	45.9	45.9	45.9	45.9	45.9	45.9
9	45.3	45.3	45.3	45.3	45.3	45.3	45.9	45.9	45.9	45.9
10	44.5	46.3	46.3	46.3	46.3	46.9	46.9	46.9	46.9	46.6
11	46.5	46.6	46.6	47.3	47.3	48.0	48.0	48.0	48.0	48.7
12	49.5	50.1	50.1	50.1	50.9	50.9	51.7	51.7	51.7	52.4
13	51.7	52.4	52.4	53.2	53.2	54.0	54.0	54.8	55.6	55.6
14	55.3	55.6	56.4	56.4	57.3	58.1	58.1	58.9	58.9	59.0
15	58.0	58.0	59.8	59.8	60.7	61.6	61.6	62.6	53.4	63.4
16	62.4	63.4	64.4	65.3	65.3	66.3	67.3	67.3	68.3	69.3
17	68.2	69.3	69.3	70.3	71.4	71.4	72.4	73.5	74.6	75.7
18	73.5	75.7	75.7	76.7	79.1	80.1	81.4	82.6	85.0	86.3
19	86.3	87.6	90.2	92.9	95.7	98.5	100.0	100.0	100.0	100.0
20	100.0									

Press 'y' if you wish to SAVE this calibration – any other key to discard it.
y
NimbleSig III RF Gen A Frequency Response Cal Saved to Non-Volatile Memory!

Figure 12 — Generator "A" frequency response flattening amplitude scale factor table.

```
>Fa10
>xc

NimbleSig III DDS Reference Frequency Setup

Enter Nominal DDS Reference Clock Frequency in Hertz For this design

Example:

 If reference oscillator is 25MHz and multiplier is set to 20 enter:

        > 500000000

>500000000

NimbleSig III DDS Reference Frequency Setup Saved to Non-Volatile Memory!

DDS Reference Frequency Setup (Prior to Calibration) = 500000000.0 Hz.

>_
```

Figure 13 — DDS reference frequency setup.

Generator "A" frequency to 10,000,000 Hz with the "fam10" command. Connect a small antenna made from a jumper cord to the center conductor of a coaxial lead connected to the Generator A output. The objective is to be able to radiate just sufficient signal level from the NS3 generator to closely match the amplitude of the WWV signal at the receiver antenna terminals. This will result in cyclic cancellation of the WWV reception as the two signals beat together. You will need to be able to couple the proper amount of signal to clearly hear and/or see the cancellation of the two signals on the receiver S meter as they beat together.

Alternately if you have access to a high quality and well calibrated VHF frequency range counter, just connect the generator A output to the frequency counter input. Set the generator output frequency to 100 MHz and following the procedure below adjust the DDS reference value for an output frequency of 100,000,000 Hz.

The next step is to enter the "xhz" command which brings up the dialog in Figure 14. Press "n" at this point to use the nominal design reference frequency just previously entered as a starting point.

As shown in Figure 14 you will then receive prompts for adjusting the frequency in either ±1, ±0.1 or ±0.01 Hz steps.

If you are using the WWV technique, if the beat rate is greater than 1 beat per second (which will probably be the case) use the "u" or "d" keys to lower the beat rate. One you obtain a beat rate less than 1 per second then use the "I" and "f" keys to lower the beat rate to close to one cancellation every 5 or 10 seconds. As shown in this example, it only took about a half dozen key clicks to establish the reference frequency as 500,000,120 Hz.

The very fine "o" and "g" keys are provided for demanding frequency applications that are referenced to highly accurate frequency standards. When finished press "c" to continue and "y" to save the reference frequency, as desired. NS3 will then report the reference frequency and this completes this calibration procedure. The new reference frequency will subsequently be reported on the NS3 start up welcome screen.

Windows Mouse Button Control for NimbleSig III

At the time of this writing there is a free *Windows* platform terminal program available for download from the Internet called *uCon*, which provides a relatively user friendly interface for controlling an NS3 module.[3] The *uCon* program provides 24 mouse click buttons that are easy to customize for sending predefined character strings. I have written a configuration file for *uCon* called "NimbleSigIIIuCom.ct" that customizes the buttons with the more common command prefixes. Figure 15 is a screen shot of *uCon* illustrating a DDS register dump that was triggered by the mouse click of the F2 "DDSregDump" button. As the top row of buttons also respond to the associated function keys, a single F2 key stroke could alternately have been used to trigger the register dump.

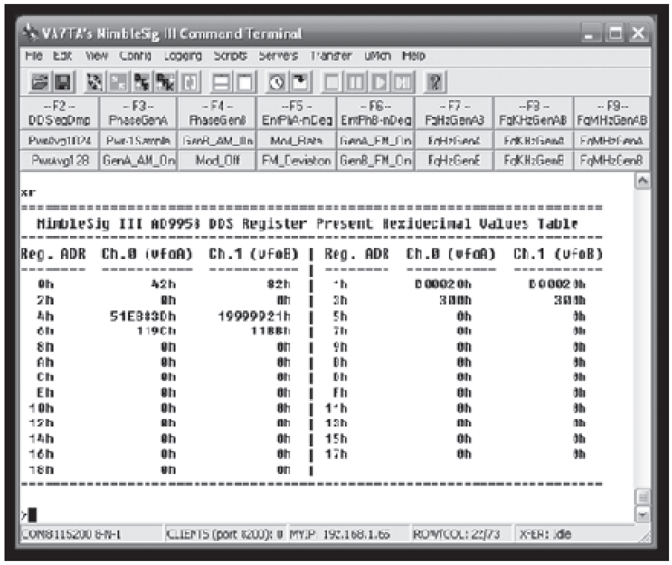
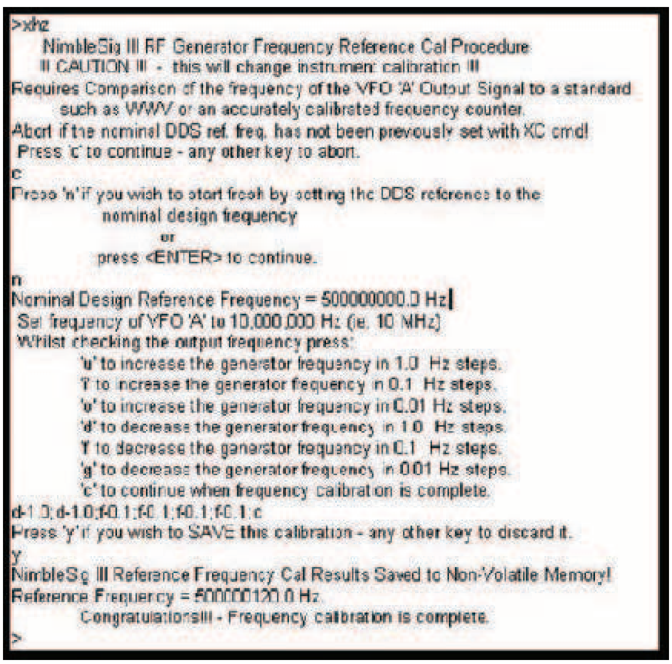


Figure 15 — The *uCon* terminal program is customized for NimbleSig III control.

The "NimbleSigIIIuCom.ct" file is available for download on my Web page.[4] It should be installed in the config folder within the *uCon* install directory on your system (most commonly "C:\ Program Files\ ucon\config\"). After starting *uCon* you then load the NS3 configuration using the "File/open" menu command, which should provide a selection display for all the ".ct" extension configuration files including "NimbleSigIIIuCom.ct." You will probably need to change my COM8 port assignment to match your NS3 interface COM port assignment. Then save your modified configuration with the "File/Save" menu item.

After installing the "NimbleSigIIIuCom.ct" configuration file and modifying it to match your COM port assignment then you may make a shortcut icon that will start up *uCon* configured and ready for NS3 control. Assuming you have installed *uCon* in the "C:\Program Files\ucon\" folder you may do this by entering the following line in the short cut icon Properties/Target window: "C:\Program Files\ucon\ucon.exe" "..\ucon\config\NimbleSigIIIuCom.ct"



```
>xhz
   NimbleSig III RF Generator Frequency Reference Cal Procedure
   !! CAUTION !! - this will change instrument calibration !!
Requires Comparison of the frequency of the VFO 'A' Output Signal to a standard
      such as WWV or an accurately calibrated frequency counter.
Abort if the nominal DDS ref. freq. has not been previously set with XC cmd!
 Press 'c' to continue - any other key to abort.
c
Press 'n' if you wish to start fresh by setting the DDS reference to the
            nominal design frequency
                     or
            press <ENTER> to continue.
n
Nominal Design Reference Frequency = 500000000.0 Hz
 Set frequency of VFO 'A' to 10,000,000 Hz (ie. 10 MHz)
 Whilst checking the output frequency press:
         'u' to increase the generator frequency in 1.0  Hz steps.
         'i' to increase the generator frequency in 0.1  Hz steps.
         'o' to increase the generator frequency in 0.01 Hz steps.
         'd' to decrease the generator frequency in 1.0  Hz steps.
         'f' to decrease the generator frequency in 0.1  Hz steps.
         'g' to decrease the generator frequency in 0.01 Hz steps.
         'c' to continue when frequency calibration is complete.
d-1.0;d-1.0;f-0.1;f-0.1;f-0.1;f-0.1;c
Press 'y' if you wish to SAVE this calibration - any other key to discard it.
y
NimbleSig III Reference Frequency Cal Results Saved to Non-Volatile Memory!
Reference Frequency = 500000120.0 Hz
         Congratulations!!! - Frequency calibration is complete.
>
```

Figure 14 — DDS reference frequency adjustment.

Recent NimbleSig III Update Information

Recently I have completed network analyzer transmission test results on the new 230 MHz low pass filter design. Photos of the 0 to 750 MHz swept frequencies responses for four filters I built up can be found on my Web page.[5] Although I did not find it practical to tune the traps in general I am very pleased with the responses of the filters. The sharpness, depth and consistency between filters within the 200 to 300 MHz cutoff frequency portion of the responses seem quite remarkable.

I have provided some guideline photos for the mounting of the

NS3 PCB into an off-the-shelf die cast aluminum enclosure manufactured by Bud. This information can also be found on my Web page.[6]

There is a minor update to the firmware, which improves the calibration accuracy of the power meter by a few tenths of a dB. The new hex file is available for download on my Web site.[7]

Conclusion

I hope this multi-part article has been of interest to many *QEX* readers. I also hope that RF equipment construction enthusiasts who decide to build a NimbleSig III module enjoy a rewarding experience! If there is sufficient interest shown for NS3 I may be motivated to design a slightly revised PC board that would correct a couple of minor glitches and make ISP programming less awkward to initiate. Bare PC board availability information may be found on my Web site, where I plan to add Nimble Sig III information and software updates as time permits. (See Note 2). I would be pleased to receive comments, questions, bug reports or suggestions.

I wish to again express my appreciation to Analog Devices Inc. for their innovative and excellent RF integrated circuit product line which provided the foundation for this project.

Notes

[1]The AD9958 data sheet is available at: **www.analog.com/static/imported-files/data_sheets/AD9958.pdf**

[2]NimbleSig III Web site: **www3.telus.net/ta/NimbleSig%20III/**

[3]You can download the uCon-Embedded Console Terminal Program at: **www.microcross.com/ucon_install.zip**

[4]You can download the uCon configuration file at: **www3.telus.net/ta/NimbleSig%20III/uConTerminal/**

[5]The 230 MHz bandwidth low pass filter design is given on the NimbleSig III Web site at: **www3.telus.net/ta/NimbleSig%20III/NS3_230%20MHz%20LowPassFilter/NS3_230_MHz_LPF_Network_Analyzer_Test_Results.html**

[6]There are more details about the BudBox enclosure at: **www3.telus.net/ta/NimbleSig%20III/NS3_BudBoxConstruction/index.html**

[7]The latest version of the NimbleSig III program files are available for download from the Nimble Sig III Web site at: **www3.telus.net/ta/NimbleSig%20III/NS3%20HEX%20Code/**

Tom Alldread, VA7TA, became interested in electronics when still in grade school, repaired radio and television sets in his teens, obtained his Amateur Radio ticket at the age of 19, and went on to obtain commercial radio operator certification a year later. He subsequently graduated from the Capitol Radio Engineering Institute Engineering Technology program. His career was in the telecommunications industry where he worked in microwave and VHF radio equipment maintenance, training, engineering standards and design, and long distance network operations management.

Tom is now retired and lives on Vancouver Island with his wife Sylvia, VA7SA. Tom has a history in repeater building and emergency communications, and he enjoys operating mobile CW, building equipment and is currently the net manager for the 20 meter Trans Canada Net. His other interests include microcontroller programming, circuit design, computing, hobby farming, digital photography, bicycling and sailing.

QEX

Gary Steinbaugh, PE, AF8L

9534 Appomattox Court, Loveland, OH 45140-7140, gsteinbaugh@yahoo.com

A Cybernetic Sinusoidal Synthesizer: Part 2

A system of four modules produces high purity, frequency- and amplitude-stabilized sine waves.

In the first part of this series, we celebrated some of the pioneers in cybernetics, and took a look at a few of the inherent difficulties in automatic control. Now let us apply this knowledge to a real RF system.

A Low Distortion, Ovenized, Crystal-Controlled Oscillator (OXCO)

First, consider the oscillator proper. Figure 4 is a photograph of the oscillator with the heated crystal. Figure 5 shows its schematic, and Figure 6 is a list of components. This is a conventional Colpitts oscillator, but it employs the trick of extracting the output signal from the crystal itself, in effect using the crystal as a high-Q bandpass filter.

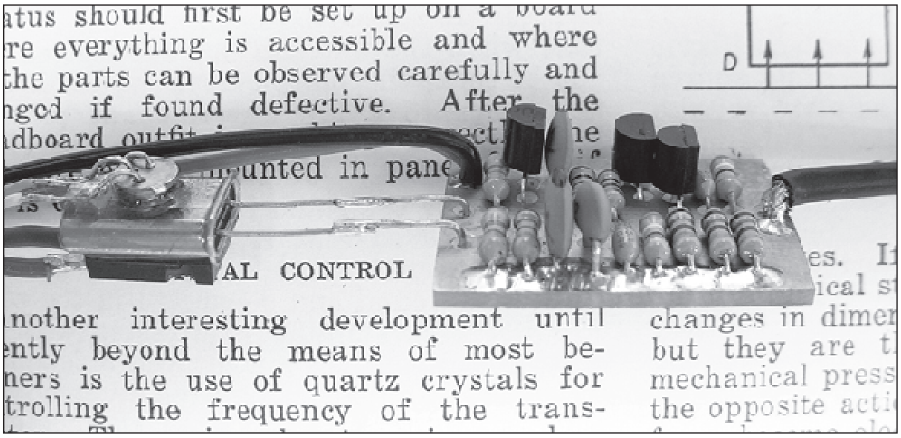


Figure 4 — This photo shows the oscillator with the heated crystal. The circuit is lying on a 1926 *ARRL Handbook*, open to the section about crystal oscillators.

Returning to the bathtub and shower examples of Part 1, an oscillator is similar to the shower in that the frequency of oscillation, ω, is a function of time constants and delays, in this case provided by the crystal.[9] (You might also think of a time delay with respect to the periodic output sinusoid as a phase shift.) The OXCO loop diagram is shown in Figure 7. To clarify the equivalence of the schematic with the loop diagram, Figure 8 shows the basic schematic. Figure 9 shows the ac equivalent (without biasing or other circuit refinements; note that the parallel-resonant crystal is modeled by an inductance), and Figure 10 shows the ac equivalent, rearranged to reveal the feedback path. In essence, the transistor is trying to adjust the "temperature" of its base, but the LC "pipe" time lag results in constant hunting, as in the shower example.

An important concept in control engineering is the damping factor, ζ (represented by a lower case Greek zeta), which is analogous to the shock absorbers on an automobile suspension system. If ζ is too low (for example, 0.3), the automobile will spring up and down at each bump in the road; if ζ is too high (for example, 3.0), every disturbance will be coupled through the automobile to its occupants. An oscillator with a constant output amplitude has $\zeta = 0$, and it actually *depends* on noise in the amplifier to start its oscillations. Say, noise isn't all bad! (Note that ζ is dimensionless; Bode plots measure stability with phase margin φ_M in radians.) In Figure 11, the response of an undamped loop is diagrammed; similar damping graphs will be provided for the control loops of other modules.

The output is buffered by a MOSFET / bipolar transistor combination, and passes through a 6 dB, 50 Ω resistive attenuator. Any external signal with the impudence to enter this output will be attenuated, and then, upon reflection, will be attenuated again. The result is a lowered SWR, or in other words, a good output impedance match.

Figure 12 is a photograph of the completed oscillator circuit. While prototyping, I typically use solderless breadboards for frequencies below about 100 kHz, and ground-plane construction for higher frequencies. I appear to have inherited "the little-old-watchmaker gene," however, and simply cannot suffer these ugly beasties to be seen … once a circuit is debugged, I typically redo it using perfboard for low frequencies, or wired-trace construction for RF. A salutary side effect of this approach is verification of the circuit's reproducibility; I have built half a dozen of these oscillators, and all are well-behaved. Figure 13 shows the wiring plan that was developed by drawing on a scrap of perfboard, which was in turn taped to a piece of copper-clad printed circuit board and used to guide a no. 64 drill. The board was sanded to size, copper was removed around non-grounding holes with a no. 40 drill in a pin vise, and the copper was protected from

[1]Notes appear on page 27.

tarnishing with electroless tin plating (see Figure 12). Figure 14 depicts the completed wiring.

I eschew etched boards unless forced to use near-microscopic (translation: surface mount) components, as with the upcoming RF power level controller project. I find the toner-transfer method of fabrication to be the most elegant approach.[10] Although etchable, I usually avoid traces smaller than ½ mm, as they are susceptible to damage during soldering.

The output is trimmed to –10 dBm by varying the oscillator supply voltage with an LM317 voltage regulator (in itself a closed-loop system!); in lieu of a laboratory RF wattmeter, a calibrated oscilloscope can be used; terminate the oscillator output at the oscilloscope input with a 50.0 Ω coaxial resistor, and adjust the peak-to-peak output amplitude to 0.200 V.[11]

Home Cooking a Crystal

Now consider the crystal oven. Crystals are not ovenized merely to keep them at a constant temperature. A standard AT-cut crystal has a temperature dependence as shown in Figure 15; at about room temperature, the rate of temperature variation (slope) is small, but not zero. By slicing the crystal at a slightly greater angle, the modified curve in Figure 15 is obtained; note that the deviation is zero at about 75°C (called the upper turning point). Thus temperature excursions of several degrees in either direction will have almost no effect on frequency. (Note that there is also a lower turning point, useful only to those who crave the complications that would accompany an RXCO (refrigerator-stabilized crystal-controlled oscillator).) Such non-AT-cut crystals must be special-ordered; ask for a 75°C calibration at 10.000000 MHz with a load capacitance of 28 pF. International Crystal produced mine for $24.92 plus shipping.[12]

Similar to a bathtub, an oven is an easily controlled process. The oven in your kitchen

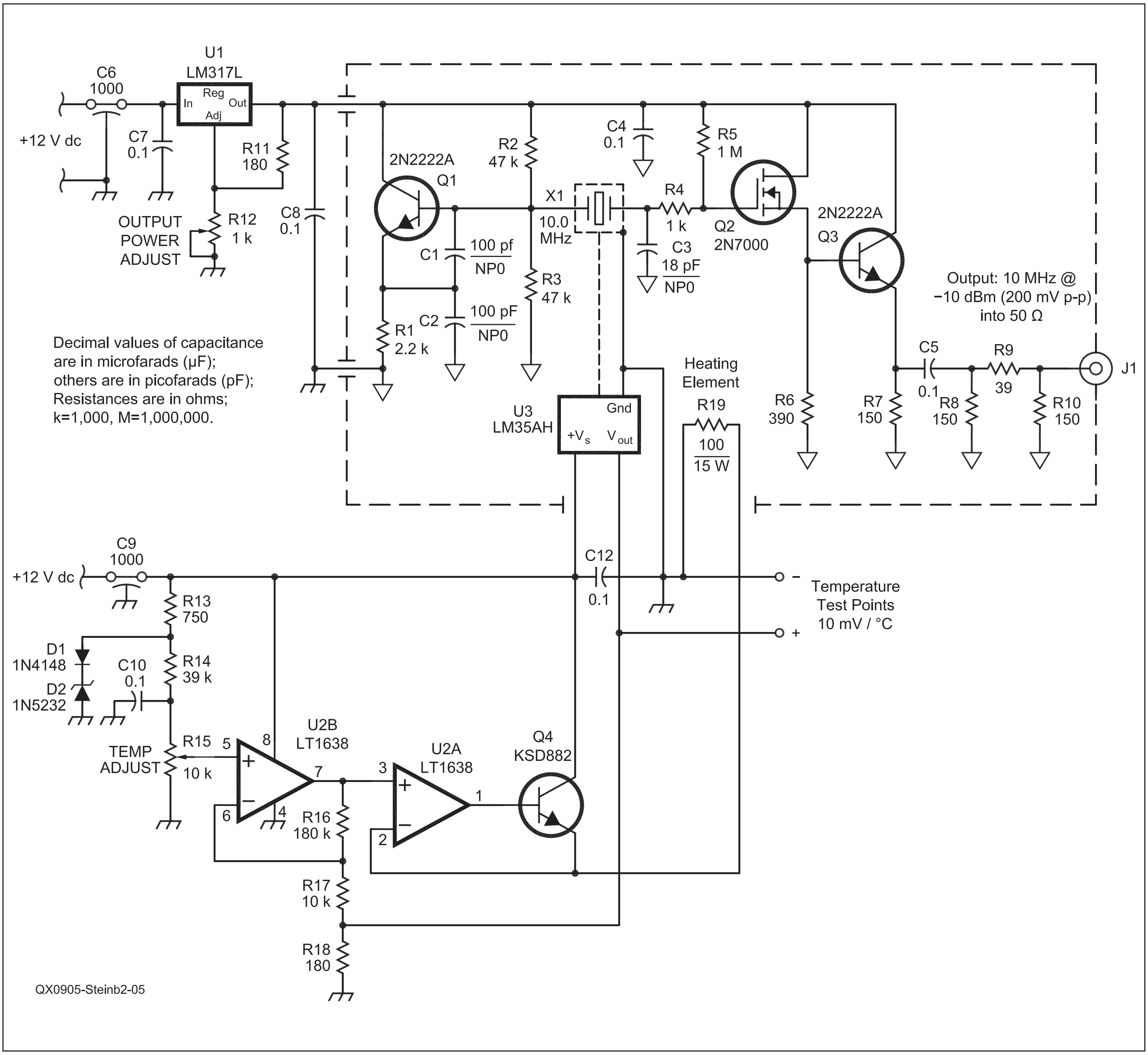


Figure 5 — Here is the schematic diagram of the ovenized crystal controlled oscillator.

LOW DISTORTION OVENIZED CRYSTAL CONTROLLED OSCILLATOR COMPONENT LIST

R1	2.2 kΩ	¼ W, 5% Carbon Film
R2	47 kΩ	¼ W, 5% Carbon Film
R3	47 kΩ	¼ W, 5% Carbon Film
R4	1 kΩ	¼ W, 5% Carbon Film
R5	1 MΩ	¼ W, 5% Carbon Film
R6	390 Ω	¼ W, 5% Carbon Film
R7	150 Ω	¼ W, 5% Carbon Film
R8	150 Ω	¼ W, 5% Carbon Film
R9	39 Ω	¼ W, 5% Carbon Film
R10	150 Ω	¼ W, 5% Carbon Film
R11	180 Ω	¼ W, 5% Carbon Film
R12	1 kΩ	Trimpot
R13	750 Ω	¼ W, 5% Carbon Film
R14	39 kΩ	¼ W, 5% Carbon Film
R15	10 kΩ	Trimpot
R16	180 kΩ	¼ W, 5% Carbon Film
R17	10 kΩ	¼ W, 5% Carbon Film
R18	180 Ω	¼ W, 5% Carbon Film
R19	100 Ω	Power Film, Caddock Part # MP915-100-1%
C1	100 pF	NP0 (C0G) Disc
C2	100 pF	NP0 (C0G) Disc
C3	18 pF	NP0 (C0G) Disc
C4	0.1 µF	Ceramic
C5	0.1 µF	Ceramic
C6	1000 pF	Feedthrough
C7	0.1 µF	Ceramic
C8	0.1 µF	Ceramic
C9	1000 pF	Feedthrough
C10	0.1 µF	Ceramic
C11	0.1 µF	Ceramic
C12	0.1 µF	Ceramic
X1	10.000000 MHz	HC49U Quartz Crystal, Custom Calibration (see text)
D1	1N4148	100 V 500 mW
D2	1N5232	5.6 V 500 mW Zener
Q1	2N2222A	NPN Bipolar
Q2	2N7000	N-Channel Enhancement-Mode
Q3	2N2222A	NPN Bipolar
Q4	KSD882	NPN Bipolar Power
U1	LM317L	Adjustable Voltage Regulator
U2	LT1638	Low Power Rail-to-Rail In/Out Dual Operational Amplifier (Linear Technology)
U3	LM35AH or equiv.	Celsius Temperature Sensor (National Semiconductor)

Hammond Model 1590E Cast Aluminum Enclosure
Hammond Model 1590H Cast Aluminum Enclosure
Phoenix MPT 0,5/ 7-2,25 7-position Terminal Block

Misc.: BNC Jack, Polystyrene Foam Insulation, Fiberglass Insulation, 8-pin DIP Socket, Hardware.

Figure 6 — The parts list for the oscillator is shown here.

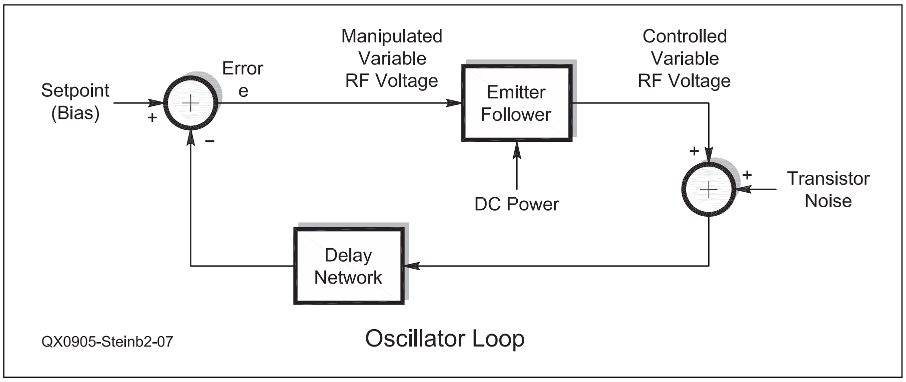


Figure 7 — This block diagram shows the ovenized crystal oscillator (OXCO) loop.

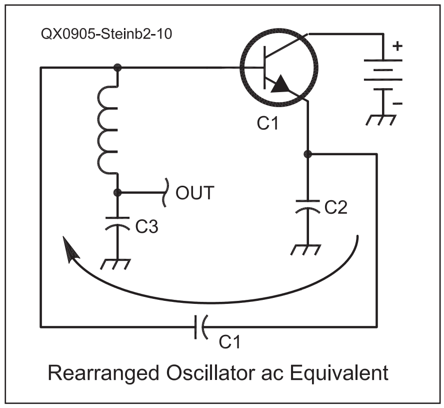


Figure 10 — This drawing shows the ac equivalent circuit, rearranged to better show the feedback path. In essence, the transistor is trying to adjust the "temperature" of its base, but the LC "pipe" time lag results in constant hunting, as in the shower example from Part 1.

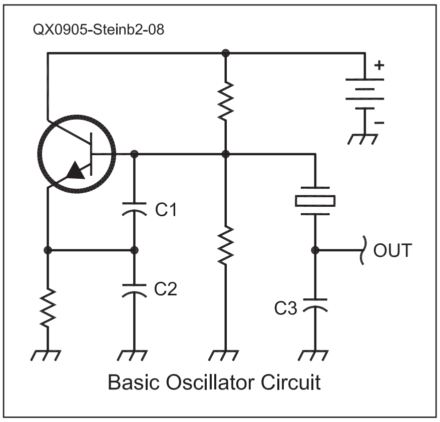


Figure 8 — This is the basic schematic diagram of the OXCO.

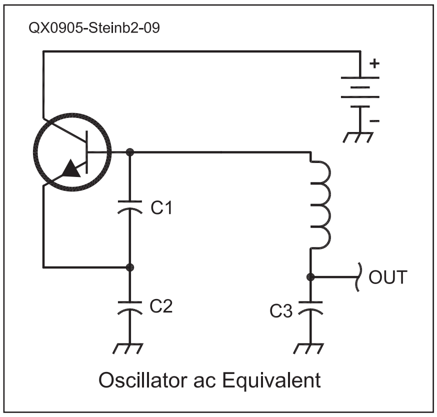


Figure 9 — Here is the ac equivalent circuit (without biasing or other circuit refinements). Note that the parallel-resonant crystal is modeled by an inductance.

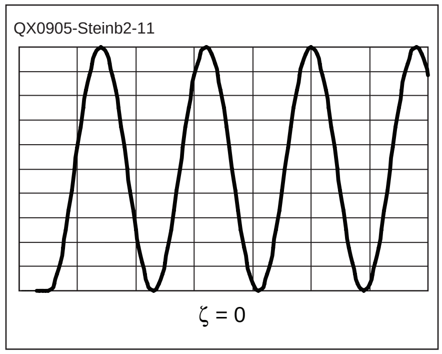


Figure 11 — This graph represents the response of an undamped control loop.

uses a simple *on-off* controller: the heating element is on if the temperature is below the setpoint (325°F, for example), and off if above. This produces an oven temperature that constantly fluctuates around an average temperature, however. This crystal oven design incorporates the more refined *proportional* control action, which holds a constant oven temperature near the desired 75°C setpoint. Even more sophisticated approaches are available, and we will examine the widely used proportional-integral-derivative (PID) controller in the next part of this series, but the added complexity is unwarranted for this application's control requirements (I strive to follow Einstein's advice to make everything as simple as possible … but not simpler).

The loop diagram is shown in Figure 16, and the step response of the closed loop is graphed in Figure 17. Corresponding to a ζ of 3, this response, termed *overdamped*, is appropriate for a process that does not require a quick response to noise (ambient temperature changes, power supply voltage

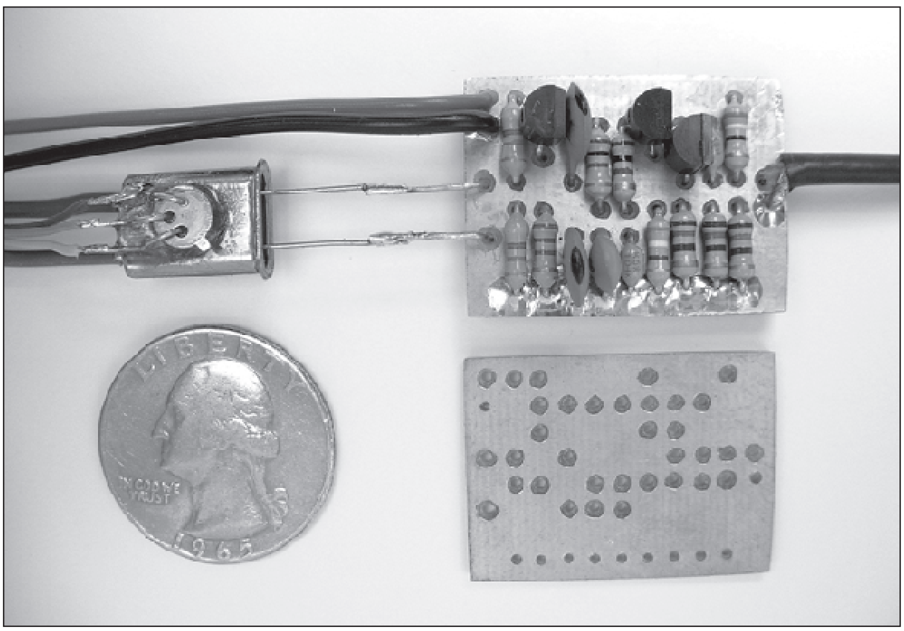


Figure 12 — This photo shows the completed oscillator circuit, along with another circuit board ready for the components to be installed. The circuit board copper cladding was protected from tarnishing with electroless tin plating.

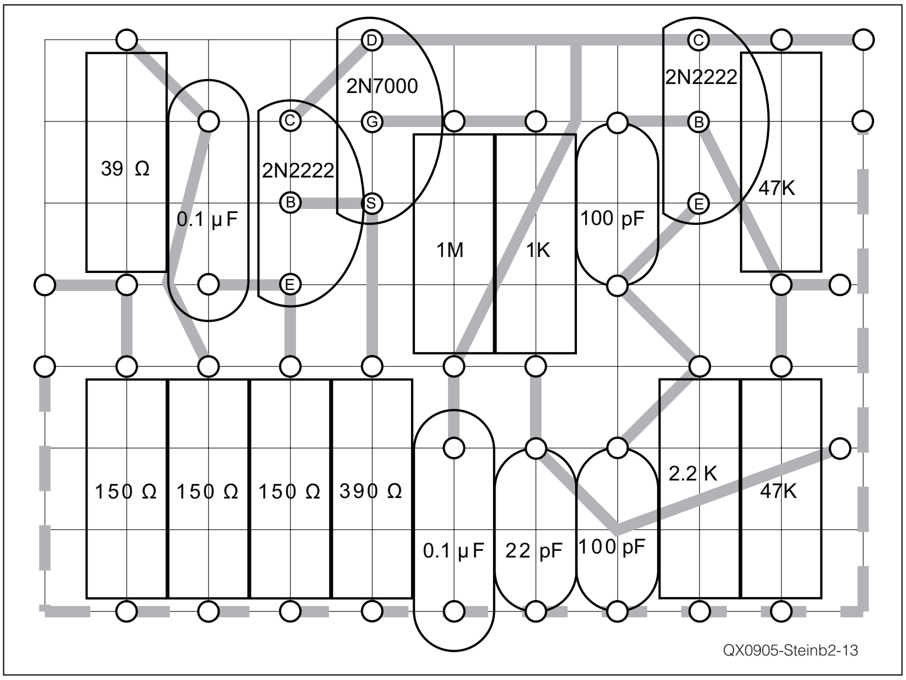


Figure 13 — This pattern shows the wiring plan that I developed by drawing on a scrap of perfboard, which was then taped to a piece of copper-clad circuit board material. I used the perfboard to guide a no. 64 drill for the component holes. I used a no. 40 drill in a pin vise to remove copper around non-grounding holes.

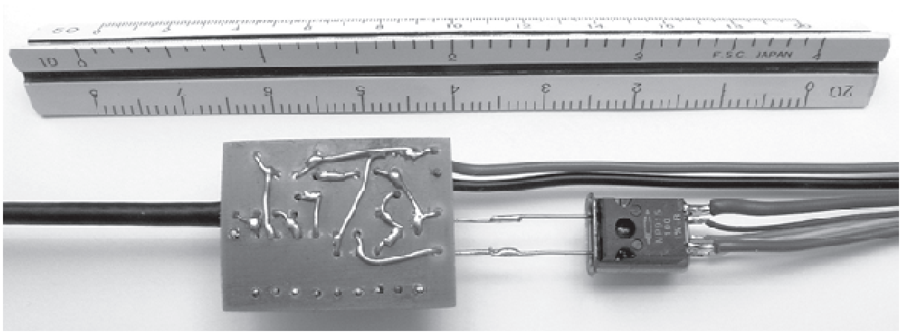


Figure 14 — Here is the completed wiring side of the circuit board.

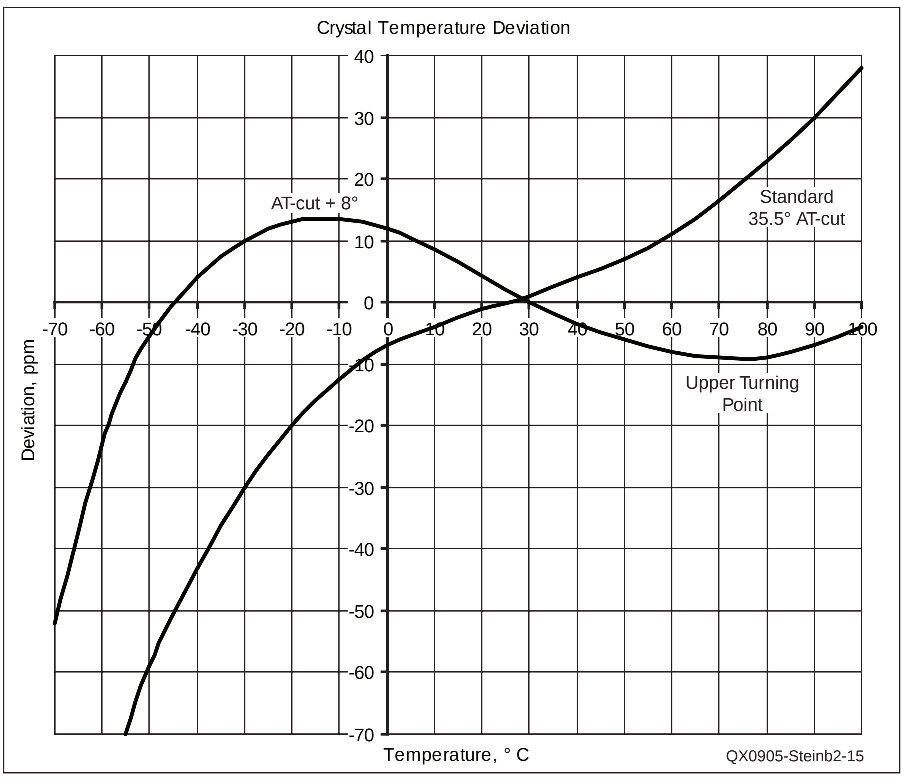


Figure 15 — This graph shows the temperature dependence of a standard AT-cut crystal. Notice that at about room temperature (20°C), the rate of temperature variation (slope) is small, but not zero. By slicing the crystal at a slightly greater angle, the modified curve of the AT-cut + 8° crystal is obtained; note that the deviation is zero at about 75°C (called the upper turning point).

drift, and so on).

For proportional controllers, I frequently use the subtracting (differential) amplifier circuit shown in Figure 18, which works as well with just one operational amplifier as the usual three-op amp circuit, except that the inverting input does not present a high impedance. Although it produces an output $V_O = V_1 + (V_1 - V_2)(R_F / R)$, if $R_F >> R$, the output becomes $V_O \approx (V_1 - V_2) A_V$ (substituting A_V for R_F / R). The oven controller error amplifier subtracts the temperature sensor signal from the setpoint, and multiplies the resulting error by 18. (I habitually use LT1638 dual op amps, but even µA741s should work acceptably in this application.)

Here is an oft-cited truth: a control loop is only as good as its sensor. In this instance, an LM35 Celsius temperature sensor produces a voltage output (0.01 V per 1°C) for easy setpoint calibration. Note that it requires a relatively low impedance load; 180 Ω is used here. Its ground lead is soldered to the crystal case, and its case and leads are epoxied to the crystal case; be sure to use an adhesive that is rated for at least 95°C (200°F). (A less expensive, plastic-packaged version might be substituted, but be mindful that these obtain essentially all of their thermal input through their leads.) A Caddock part number MP915-100-1% power film resistor, epoxied to the opposite side of the crystal case, is used as a heating element. A photograph of the temperature controller circuit board is shown in Figure 19.

The setpoint is produced with a trimpot and a stabilized voltage source, using the trick of combining a 5.6 V Zener diode (actually an avalanche diode, whose temperature coefficient is +2 mV / °C) in series with a silicon signal diode (whose temperature coefficient is –2 mV / °C) to obtain a 6.3 V temperature-compensated voltage reference. The second amplifier of the LT1638 drives a KSD882 power NPN transistor to create a current output amplifier that powers the resistive heating element. The oven controller circuit is powered by an external regulated 12 V dc at 200 mA wall-mounted dc supply, separate from the main power supply for the network analyzer, so that the oven can maintain the crystal temperature when the analyzer is switched off. Be careful to check the polarity of the power connector; I had to rewire mine to make the barrel negative. Calibration is simple, performed with a digital voltmeter connected to a two-pin header, as pictured in Figures 19 and 20.

To increase the crystal temperature to 75°C, the heat from the element is confined by two pieces of 1½ inch thick expanded polystyrene insulation, readily available at home improvement stores. This foam is easily cut with a fine-tooth razor saw (see Figure 21),

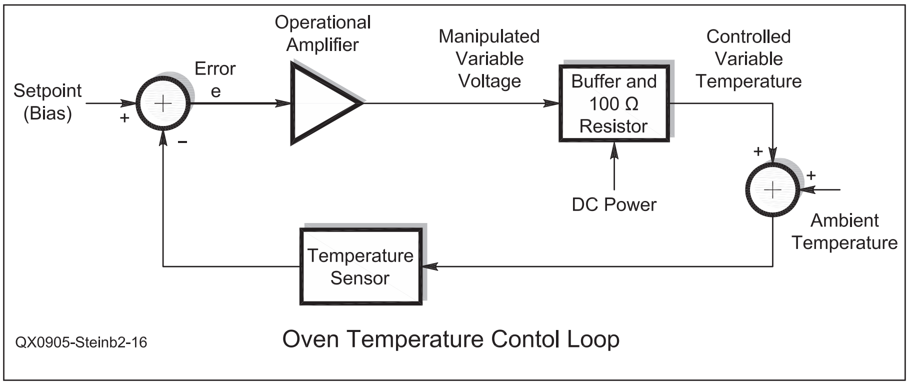


Figure 16 — The block diagram of an oven temperature control loop is shown here.

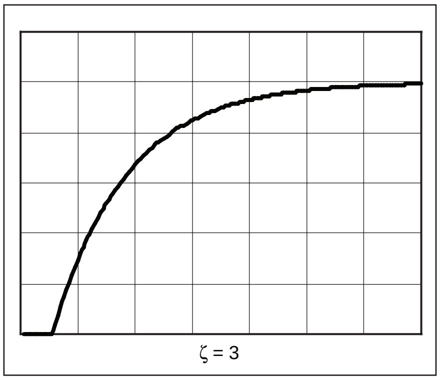


Figure 17 — This graph, with a damping factor, ζ = 3, shows the oven temperature stabilizing over time. This response shape is called *overdamped*, and is appropriate for a process that does not require a quick response to noise (ambient temperature changes, power supply voltage drift, and so on).

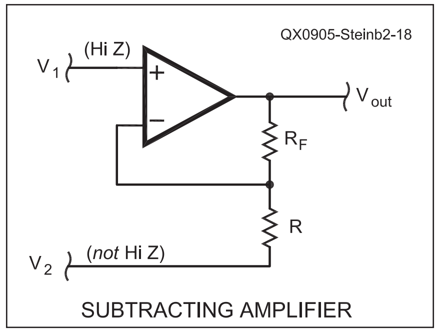


Figure 18 — For proportional controllers, I frequently use the subtracting (differential) amplifier circuit shown here.

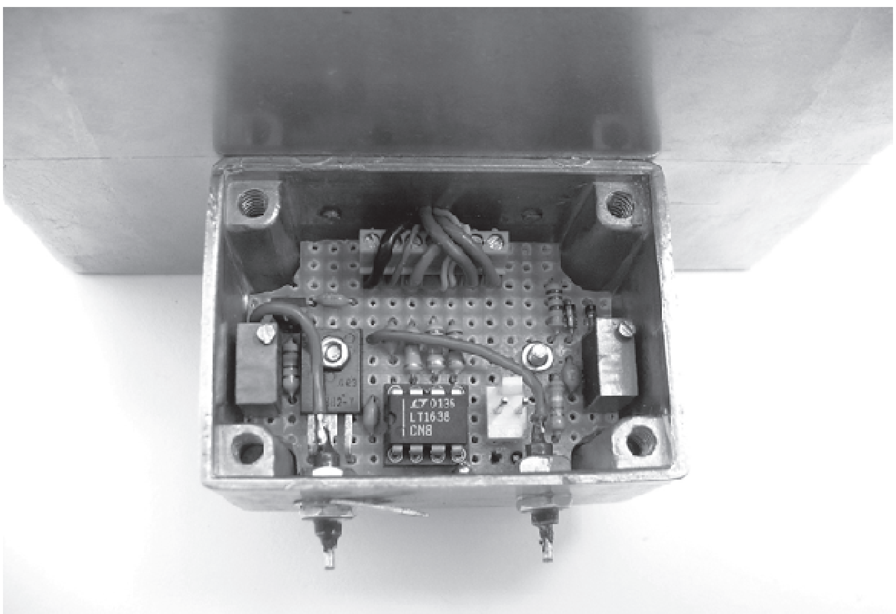


Figure 19 — This is a photo of the temperature controller circuit board mounted inside of a Hammond model 1590H cast aluminum enclosure.

and with care it can be shaped and smoothed with sandpaper. A rectangular cavity was cut in the center of the bottom piece with a hobby knife, a hole was drilled from the side into the cavity, and a plastic drinking straw was inserted as a conduit for wiring to the oven controller. Figure 22 shows the crystal / sensor / heater surrounded by some scraps of fiberglass insulation, both to reduce convection currents and for some operator-error insurance for the polystyrene, which begins to melt at about 120°C.

A graph of the oven power-up current and temperature versus time is shown in Figure 23; note that the supply current quickly drops to 75 mA for a steady-state power requirement of just 900 mW (divided between the heater and the power transistor). Figure 24 is an enlarged view of the first two minutes of the power-up graph, clearly showing the bathtub-like 1 s time delay and 48 s time constant that make the oven easy to control.

I was able to get access to an Agilent 4395A network / spectrum / impedance analyzer; Figure 25 shows its display of the OXCO output spectrum. The first harmonic is –45 dBc, the second is –53 dBc, and the third is –60 dBc. I measured the frequency to be 10.000034 MHz (I did not vex myself by trying to tune for exactly 10.000000 MHz) using an HP 5328A universal counter, which itself contains an OXCO. The frequency stability appears to be very good; although I did not attempt to determine the Allan variance, after an hour warm-up (note that the frequency will drop as the oven heats, as predicted by Figure 15), the counter readout changed no more than ±1 Hz during the subsequent 24-hour period.

In the third part of this series, some advanced control modes will be examined, and the second construction project, an RF power meter with digital and analog outputs, will be presented.

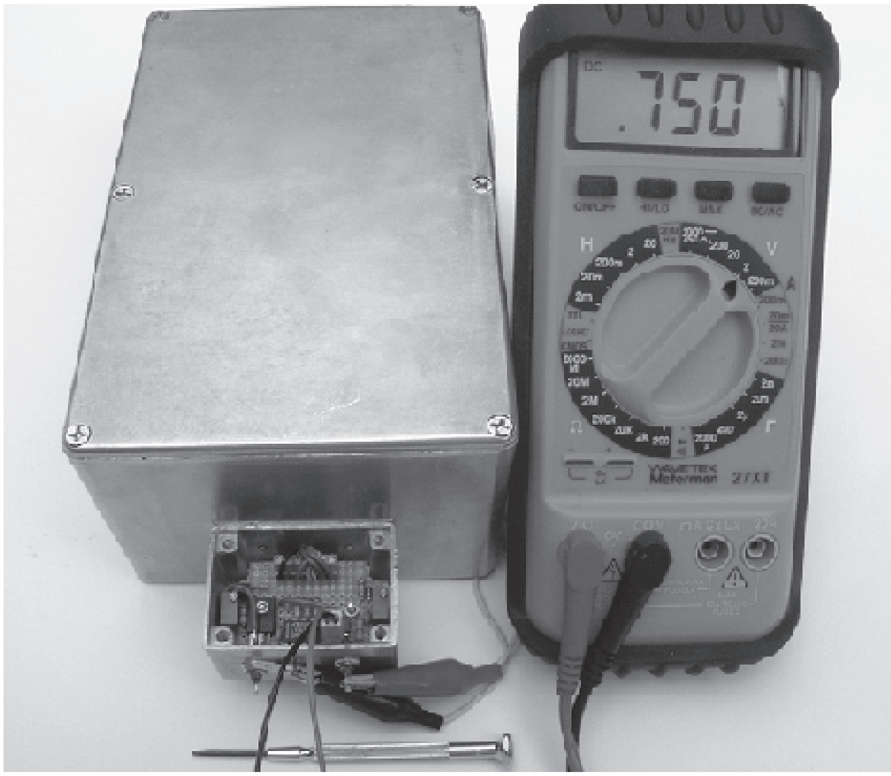


Figure 20 — Here the oven controller is being calibrated by connecting a digital multimeter to the two-pin header on the temperature controller enclosure. The crystal oscillator circuit is inside the larger Hammond 1590E cast aluminum case.

Figure 21 — To increase the crystal temperature to 75°C, the circuit board, crystal and heating element are sandwiched between layers of 1½ inch thick expanded polystyrene insulation, readily available at home improvement stores. This foam is easily cut with a fine-tooth razor saw, and with care it can be shaped and smoothed with sandpaper.

Notes

[9]In control engineering, frequencies are usually calculated using radians per second instead of hertz (remember that radians are dimensionless units, and that $\omega = 2\pi f$, and $f = \omega / 2\pi$).

[10]Pulsar, 21 Carriage Drive, Crawfordville, FL 32327-2496, (850) 926-2009, **www.pcbfx.com**.

[11]Handy shortcuts for 50 Ω impedance calculations:

$$V_{RMS} = \sqrt{\left(\left(\log^{-1}\left(\text{dBm}/10\right)\right)/20\right)}$$

$$V_{P-P} = \sqrt{\left(\left(\log^{-1}\left(\text{dBm}/10\right)\right)/2.5\right)}$$

$$\text{dBm} = 10\log\left(20\left(\text{V}_{\text{RMS}}{}^{2}\right)\right) = 10\log\left(2.5\left(\text{V}_{\text{P-P}}{}^{2}\right)\right)$$

[12]International Crystal Manufacturing, Inc., P.O. Box 1768, Oklahoma City, OK 73101-1768, (405) 236-3741, **www.icmfg.com**.

Continuously licensed since 1964, Gary Steinbaugh, AF8L, is an ARRL Life Member. Holding a BSEE from Case Institute of Technology and several patents, he is a licensed Professional Engineer, and the author of many technical articles. He is Senior Electronic Engineer for AtriCure, Inc., a manufacturer of RF electrosurgical instruments used for soft tissue ablation. Gary is also a Certified Flight Instructor and a semi-pro musician.

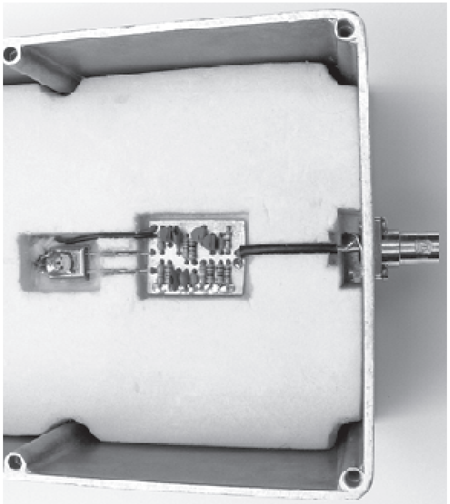


Figure 22 — I cut a rectangular cavity in the center of the bottom piece of polystyrene insulaton with a hobby knife, then I drilled a hole from the side into the cavity, and inserted a plastic drinking straw as a conduit for wiring to the oven controller. I also surrounded the crystal / sensor / heater with some scraps of fiberglass insulation, both to reduce convection currents and for some operator-error insurance for the polystyrene, which begins to melt at about 120°C.

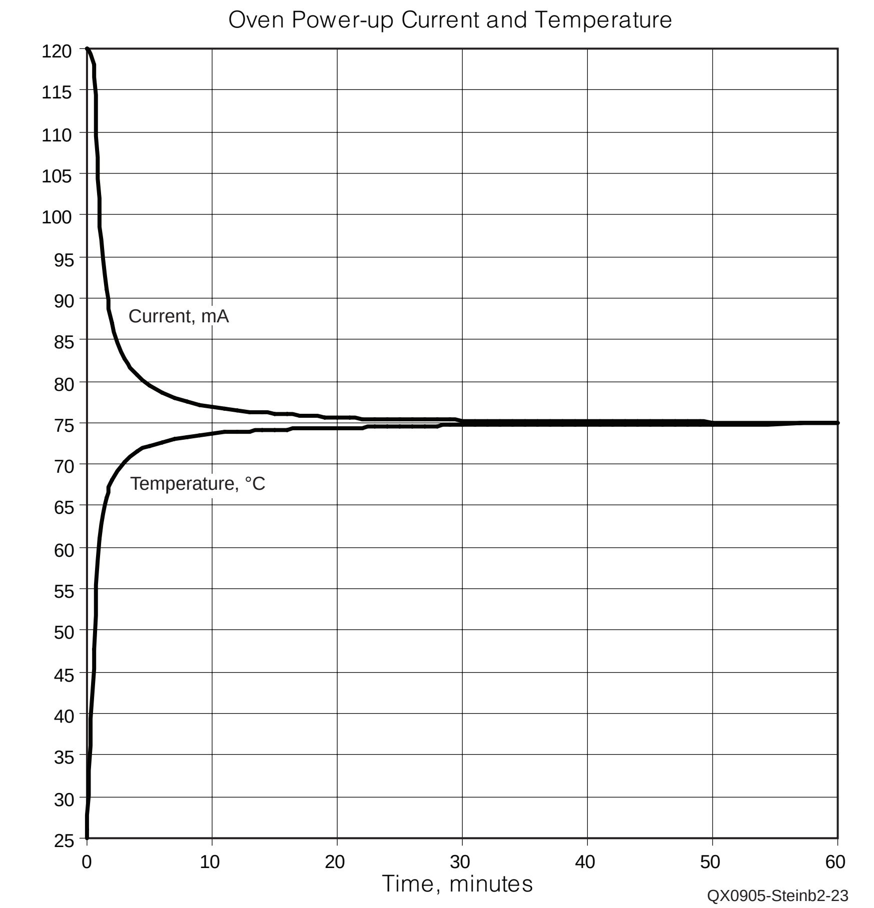


Figure 23 — This graph shows the oven power-up current and temperature versus time. Note that the supply current quickly drops to 75 mA for a steady-state power requirement of just 900 mW. (The power is divided between the heater and the power transistor.)

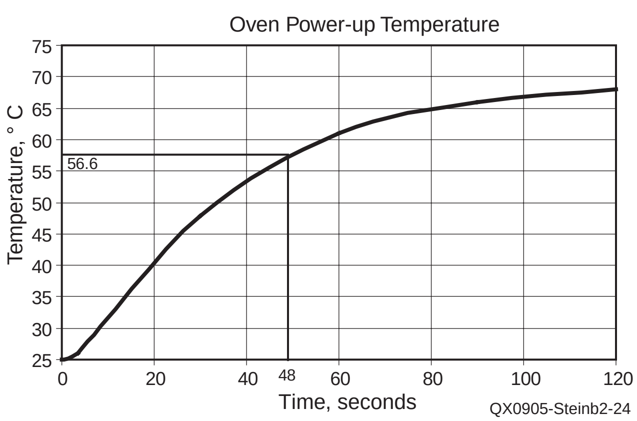


Figure 24 — This graph is an enlarged view of the first two minutes of the power-up graph, clearly showing the bathtub-like 1 s time delay and 48 s time constant that make the oven easy to control.

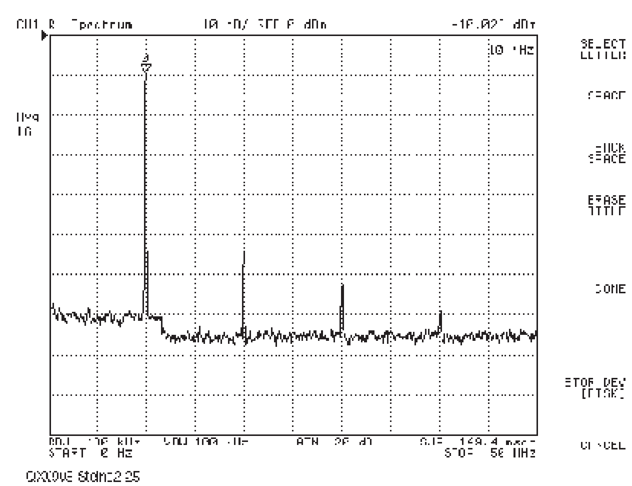


Figure 25 — I was able to use an Agilent 4395A network / spectrum / impedance analyzer to measure the OXCO output spectrum. This graph is the output from that measurement. The first harmonic is –45 dBc, the second is –53 dBc, and the third is –60 dBc. I measured the fundamental frequency to be 10.000034 MHz (I did not vex myself by trying to tune for exactly 10.000000 MHz) using an HP 5328A universal counter, which itself contains an OXCO. QEX

Prof. Dr. Thomas C. Baier, DG8SAQ

University of Applied Sciences, Prittwitzstrasse 10, 89075 Ulm, Germany; baier@hs-ulm.de

A Simple S-Parameter Test Set for the VNWA2 Vector Network Analyzer

This test set eliminates the need to swap input and output connections to the vector network analyzer when measuring S-parameters.

While beta-testing his VNWA2.1 vector network analyzer, Andreas Zimmermann, DH7AZ, came up with a brilliant idea: He took two RF relays and built a simple switch, which could commutate the VNWA TX and RX ports.[1] Thus, a two port device wouldn't have to be disconnected and turned manually any longer in order to measure its two port S-parameters. Surprising to me, his first veroboard prototype already showed good isolation values. Even more surprising to me, the relay data sheet indicated that at 1.5 GHz, isolations of 60 dB could still be expected.[2] That's when I decided to build such a switching unit myself, and modify my VNWA software so it would automatically control the relays. This way, the VNWA2 is turned into a full-featured two port network analyzer. The most charming aspect is the simplicity of the circuit, which only requires two relays.

The General Task

Figure 1 shows the general setup of the VNWA2 without S-parameter test set with a two port device under test (DUT). The signal generator's RF passes through a directional coupler into Port 1 of the DUT. The DUT will generally reflect some fraction, b_1, of the incident wave amplitude, a_1. The directional coupler or a similar device serves to separate and measure the incident wave amplitude, a_1, and the reflected wave amplitude, b_1. The DUT input reflection coefficient S_{11} can be calculated from these measured wave amplitudes, provided that the load impedance, Z_{Load} of the Through Signal detector (RX Port) is identical to the reference impedance (generally 50 Ω) and thus the signal reflected from the detector $a_2 = 0$.

$$S_{11} = \frac{b_1}{a_1} \quad \text{[Eq 1]}$$

The DUT forward transmission coefficient S_{21} can be calculated from the Through Signal, provided that the TX Port impedance is also identical with the reference impedance.

$$S_{21} = \frac{b_2}{a_1} \quad \text{[Eq 2]}$$

Thus, only two of the four DUT S-parameters can be measured with this setup. In order to measure S_{12} and S_{22}, the DUT ports 1 and 2 have to be interchanged manually by disconnecting and reconnecting the DUT. The same effect can be achieved with a simple setup of switching relays, as shown in Figure 2, without manual effort.

S-Parameter Test Set

Figure 2 shows the schematic of the VNWA S-parameter test set. Two relay toggle switches either connect the TX with Port 1 and the RX with Port 2 or alternatively the TX with Port 2 and the RX with Port 1. One interesting aspect of this circuit is the coil driver. The 5 V type relays draw 40 mA each, thus adding up to 80 mA if both relays are to be switched on. Since I wanted to power the test set from the very same USB port as the VNWA, I needed to conserve current. That's why I have connected the coils in series. Once activated, 2 V per relay would be enough to keep them switched on. But a total of 8 V (2 × 4 V) are needed to activate both relays. This is achieved by connecting the charged capacitor C4 in series with the 5 V supply.

In the off state (control input low) Q2 is conducting, but Q1 and Q3 are not. C4 is

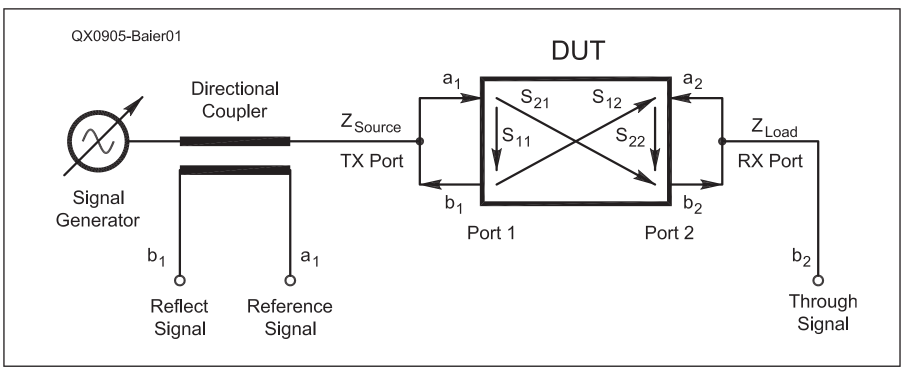


Figure 1 — General setup of the VNWA2 measuring a two port DUT without use of an S-parameter test set.

[1]Notes appear on page 32.

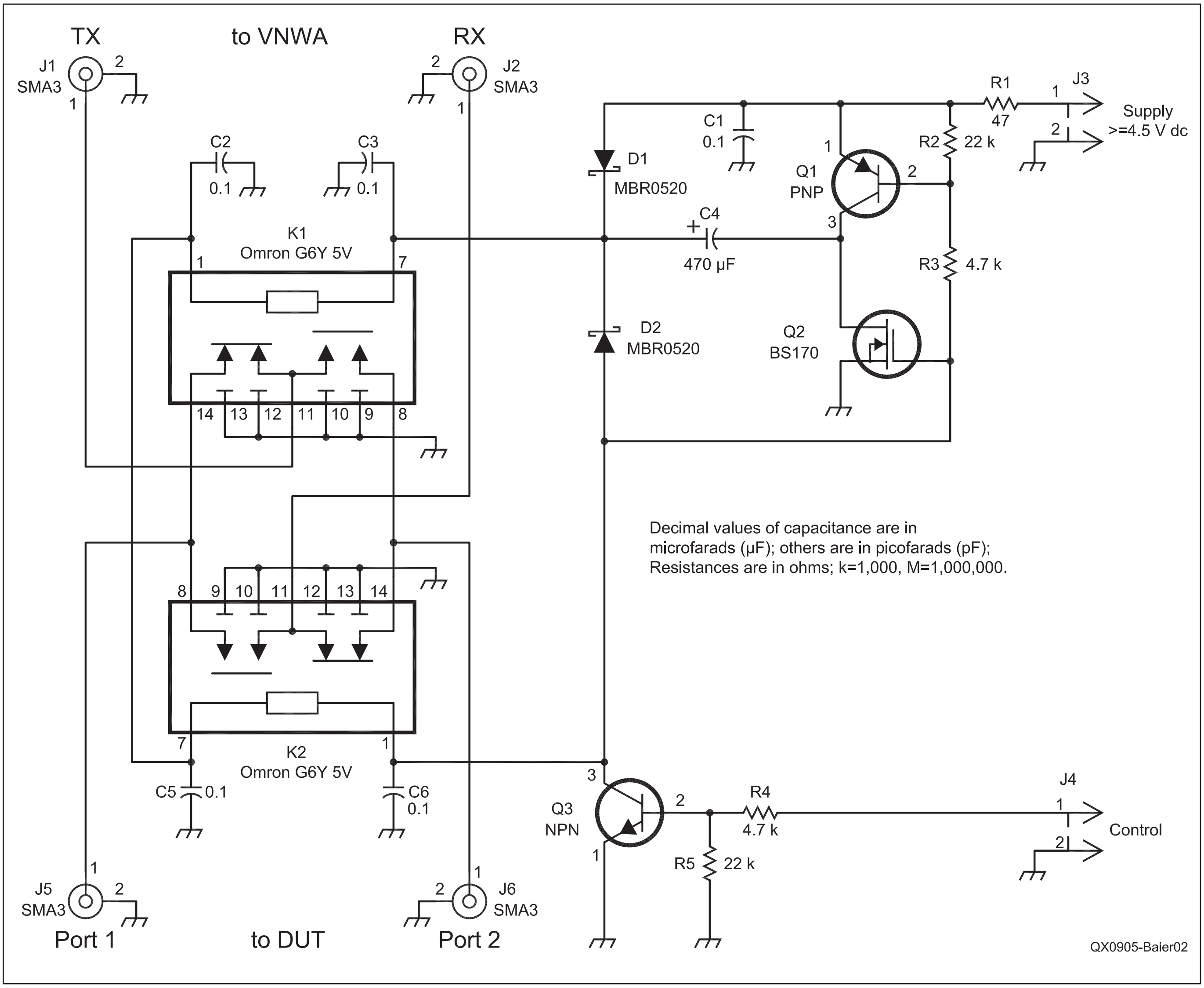


Figure 2 — Schematic of the VNWA S-Parameter Test Set.

charged through D1 and Q2. When the control line goes high, Q1 and Q3 conduct, while Q2 goes into the off state. Q1 connects the minus side of the now charged C4 to the supply voltage, which is thus added to the charge voltage of C4. The sum voltage is enough to activate both relays. After C4 is discharged through the relays, D1 takes over the reduced coil current. A current of 20 mA is enough to hold both relays in the on state. D2 is necessary to suppress high voltage induction voltage spikes, which could damage the driving transistors. I added R1 to further reduce the holding current and as protection: For a very short moment during the activation, Q1 and Q2 are conductive at the same time. R1 serves to limit the current at this moment.

Figure 3 shows the test set from the top. The two black relays can be seen with a shielding wall in between them. The control signal and the power supply are provided through the 3.5 mm audio connector on the left side. The whole circuit is built up on veroboard, which is covered with adhesive copper foil on the relay side.

Figure 4 shows the test set from the bottom, hooked up to the VNWA2. The driver circuitry is assembled onto the veroboard from the bottom, in surface mount style. In order to achieve the highest possible isolation, the shielding wall protrudes through slots in the veroboard to the bottom, and RF wiring is done with semi rigid coax lines. Also, there is a copper sheet shielding around the relay middle contact pins.

Note, that I have tapped the power supply and the control line at the VNWA control connector, thus avoiding modifications of the VNWA itself. Also note the funny homebrewed biangular male-to-male SMA adapters, which allow to freely fit different connector spacings by changing the tilt angles.

Figure 5 shows a wideband isolation measurement of the test set from TX to RX port with ports 1 and 2 disconnected. The isolation is only marginally worse than the dynamic range of the VNWA2 up into the GHz range, and is thus sufficient for most two port measurements. Note that 60 dB of isolation is achieved at 1.2 GHz with a single switch, without shunt switches!

Figure 6 shows a wideband transmission measurement of the test set from TX to RX port with ports 1 and 2 connected to each other. The <1 dB maximum insertion loss up to 1.3 GHz can be neglected for the test set application.

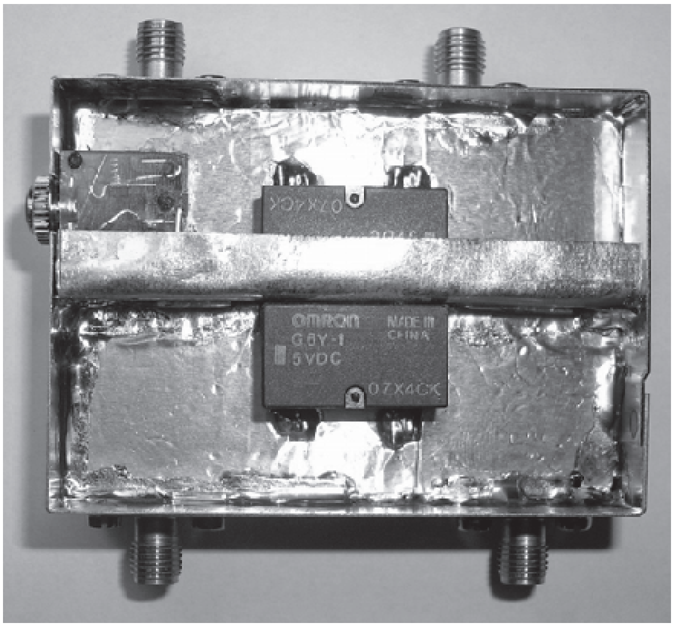


Figure 3 — Test set as seen from top. The shielding wall between the relays is most crucial for good isolation. The sheet metal box size is 65 × 45 mm².

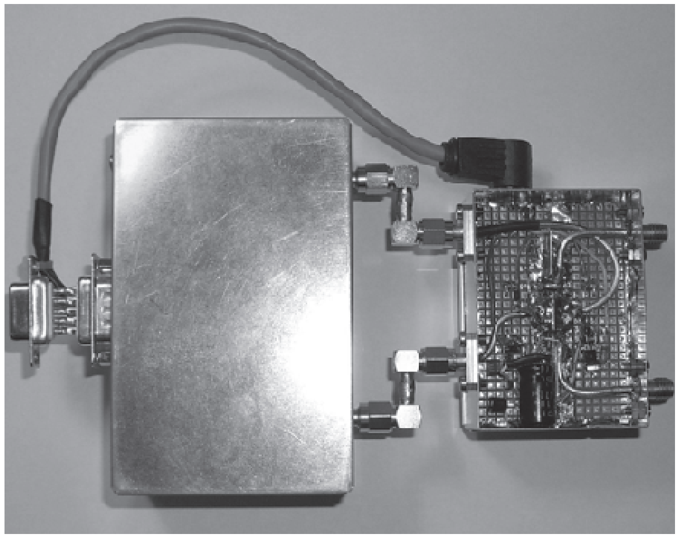


Figure 4 — The test set as seen from the bottom, hooked up to the VNWA2. Note that the connection cable taps the 5 V dc power supply and the control line at the VNWA digital Sub D9 interface.

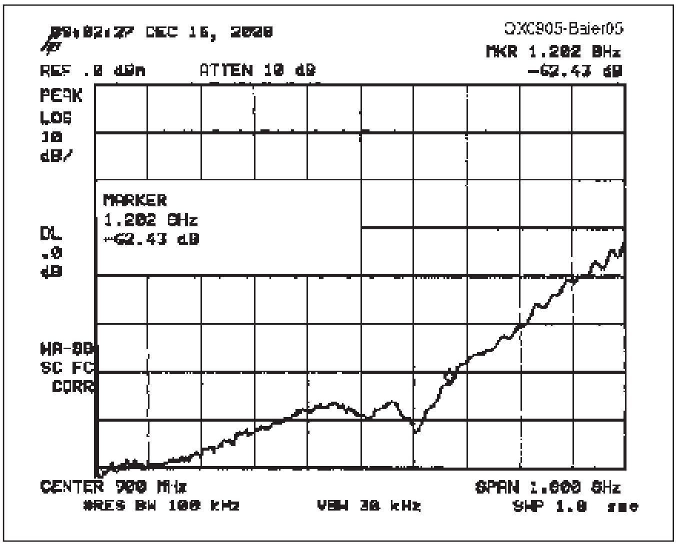


Figure 5 — The isolation measurement of the test set from TX to RX port over 0 to 1.8 GHz with ports 1 and 2 isolated.

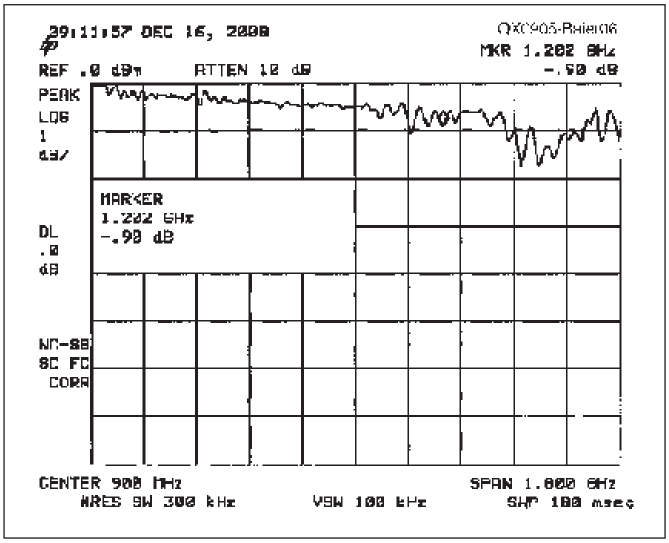


Figure 6 — The transmission measurement of the test set from TX to RX port over 0 to 1.8 GHz with ports 1 and 2 connected to each other.

Advantages of Using a Test Set, Error Correction

The most obvious advantage of the test set is the fact that two port devices can be fully characterized without manually turning them during the measurement. Thus, all four S-parameters (S_{11}, S_{21}, S_{12}, S_{22}) can continuously be measured. This fact yields another, less obvious, advantage: As discussed earlier in this article, S_{11} and S_{21} can be deduced from a forward direction measurement only if the TX port impedance, Z_{Source}, and the RX port impedance Z_{Load}, are both identical to the reference impedance (usually 50 Ω). If this is not the case, some of the transmitted signal, b_2, is reflected at the RX port, is retransmitted through the DUT and adds to the reflected input signal, b_1, which in turn gets reflected at the TX port *again*. Thus, in this case the results of the forward measurement are influenced by all four DUT S-parameters as signals travel through the DUT in both directions. The same holds true for the measurement in the reverse direction. If both directions are continuously measured, then the effects of non ideal source and load impedances can be mathematically corrected in an exact manner. This technique is known as 12-term error correction.[3]

Figure 7 shows the measured S_{11} and S_{22} of the through calibration standard. By definition, both reflection coefficients of the through standard are zero. Traces S_{11} and S_{22} show simply corrected results, which are identical with the detector reflection coefficient caused by $Z_{Load} \neq 50$ Ω. Note that the detector match is better than 25 dB over the displayed span, which is quite good. Mem1 and Mem2 show the same data 12-term corrected. The effect of imperfect load match has completely vanished and the measured data is dominated by noise. The rising noise

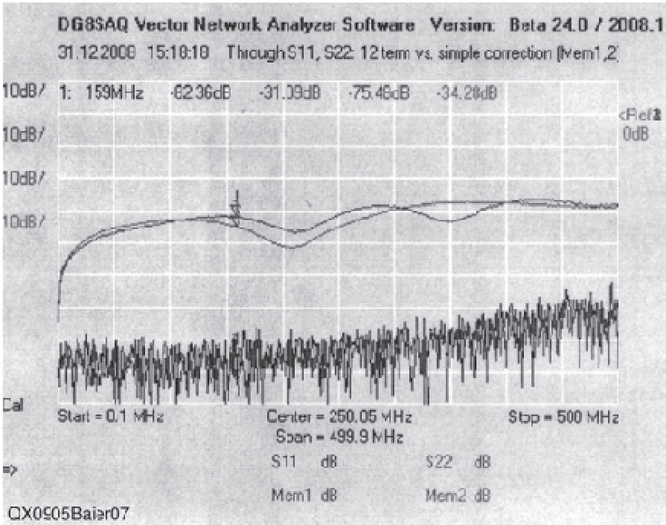


Figure 7 — Impact of 12-term error correction on measured S_{11} and S_{22} of the through calibration standard. Traces S_{11} and S_{22} show simply corrected results, while Mem1 and Mem2 show the same data 12-term corrected.

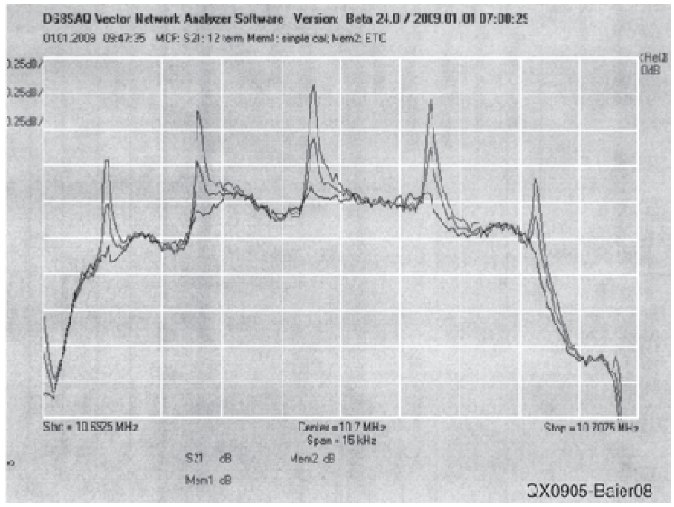


Figure 8 — Transmission data of a monolithic crystal filter measured at a reference impedance of 50 Ω with various error correction schemes and then recalculated to source and load impedances of 3000 Ω.

floor to the high frequency side reflects the decreasing available signal strength of the instrument.

One should think that the effects of 12-term correction are generally small, but they can become large if the measured S-parameters are used in a system simulation. Figure 8 shows transmission data of a monolithic crystal filter measured with the VNWA at a reference impedance of 50 Ω with various error correction schemes and then recalculated to source and load impedances of 3000 Ω, which are the optimum filter matching conditions for a flat pass band.

Note that the simple error correction scheme (trace S_{21}) shows distinct spikes in the filter pass band, while the 12-term corrected result (Mem4) is as smooth as can be. Also shown are two simple enhanced through correction schemes (Traces Mem1: ETC1; Mem2: ETC2), which only require measurements in one direction. They partly take into account non perfect matching conditions, but they do neglect multiple reflections. ETC2 can be obtained from the 12-term equations by setting $S_{22} = 0$ for the forward terms and $S_{11} = 0$ for the backward terms. ETC1 is obtained from ETC2 by setting the detector impedance equal to the reference impedance (usually 50 Ω). The ETC corrections are better than the simple one, but worse than the 12-term correction.

Summary and outlook

I have described an easy to build automatic 2 port S-parameter test for the VNWA2. In spite of its simplicity, it offers sufficiently high isolation and low insertion loss. The test set also allows application of the very accurate 12-term error correction scheme in a very simple manner. With a little imagination, the test set can easily be upgraded to include bias tee networks for measuring active components. The most recent VNWA software is capable of automatically controlling the test set. It can be found at my Web site.[4] The version of the VNWA software current as of the publication date of this article is also available for download from the ARRL *QEX* Web site.[5]

Thanks to Andreas Zimmermann, DH7AZ, for sharing his great test set idea and to Paul Kiciak for many fruitful discussions on error correction.

Professor Dr. Thomas Baier MA teaches physics, mathematics and electronics at the University of Applied Sciences in Ulm, Germany. Before his teaching assignment, he spent 10 years of work on research and development of surface acoustic wave filters for mobile communication with Siemens and EPCOS. He holds 10 patents.

Tom, DG8SAQ, has been a licensed radio amateur since 1980. He prefers the soldering iron to the microphone, though. His interests span from microwave technology to microcontrollers. Lately, he has started Windows programming with Delphi. Tom spent one year in Oregon USA rock climbing and working on his master's degree.

Notes

[1]Professor Dr. Thomas C. Baier, DG8SAQ, "A Small, Simple, USB-Powered Vector Network Analyzer Covering 1 kHz to 1.3 GHz," *QEX*, Jan/Feb 2009, ARRL, pp 32 – 36. See also **www.arrl.org/qex/2009/01/Baier.pdf**

[2]Omron G6Y Relay, see **www.omron.com/ecb/products/pdf/en-g6y.pdf**

[3]See, for example, Agilent AN 1287-3, "Applying Error Correction to Network Analyzer Measurements," Application Note, **http://cp.literature.agilent.com/litweb/pdf/5965-7709E.pdf**

[4]See **www.mydarc.de/DG8SAQ/NWA.html**

[5]The VNWA software current as of the publication date of this article is available for download from the ARRL *QEX* Web site. Go to **www.arrl.org/qexfiles** and look for the file **5x09_Baier.zip**.

QEX

Bob Hillard, WA6UFQ

10312 Cyrus Cove, Austin, TX 78739; rhillard@austin.rr.com

Universal VFO Controller

This controller works with the AMQRP Club DDS-30 and DDS-60 synthesizers, as well as boards using Silicon Labs Si570 XO ICs and other VFOs.

Two new technologies for generating an RF signal have caught the eye of the Amateur Radio homebrewer. The first technology to arrive a number of years ago was direct digital synthesis (DDS), followed more recently by Silicon Laboratories' patented DSPLL technology.

A few years ago, the New Jersey QRP Club (NJQRP) made a signal generator kit that used the Analog Devices AD9850 DDS chip. Named the DDS-30, it was built on a 1 × 2 inch plug-in circuit board, and was capable of generating frequencies up to 30 MHz.

Three years later, the American QRP Club's (AMQRP) DDS-60 kit became available. With the same form factor as the DDS-30, the DDS-60 uses the Analog Devices AD9851 DDS chip, and is capable of generating signals up to 60 MHz. The DDS-60 kit is currently available from the AMQRP club at **www.amqrp.org/kits/dds60/**.

Boards for the Analog Devices AD995x family of DDS chips, and their AD9912 DDS chip are available from David Brainerd, WB6DHW, on his Web site at **wb6dhw.com/index.html**. His AD995x board also includes a quadrature sampling detector (QSD), and can be used as a SDR communications receiver covering 1.5 to 30 MHz.

Figure 1 — This VFO controller was built by James Elledge, KB2XJ. Jim used a PacTec LH64-200 enclosure, and a white on blue LCD.

Si570: The New Kid on the Block

The Silicon Labs Si570 XO chip has become very popular among homebrewers. It uses one internal fixed-frequency crystal and DSPLL clock synthesis to provide "any rate" frequency operation with exceptional frequency stability.

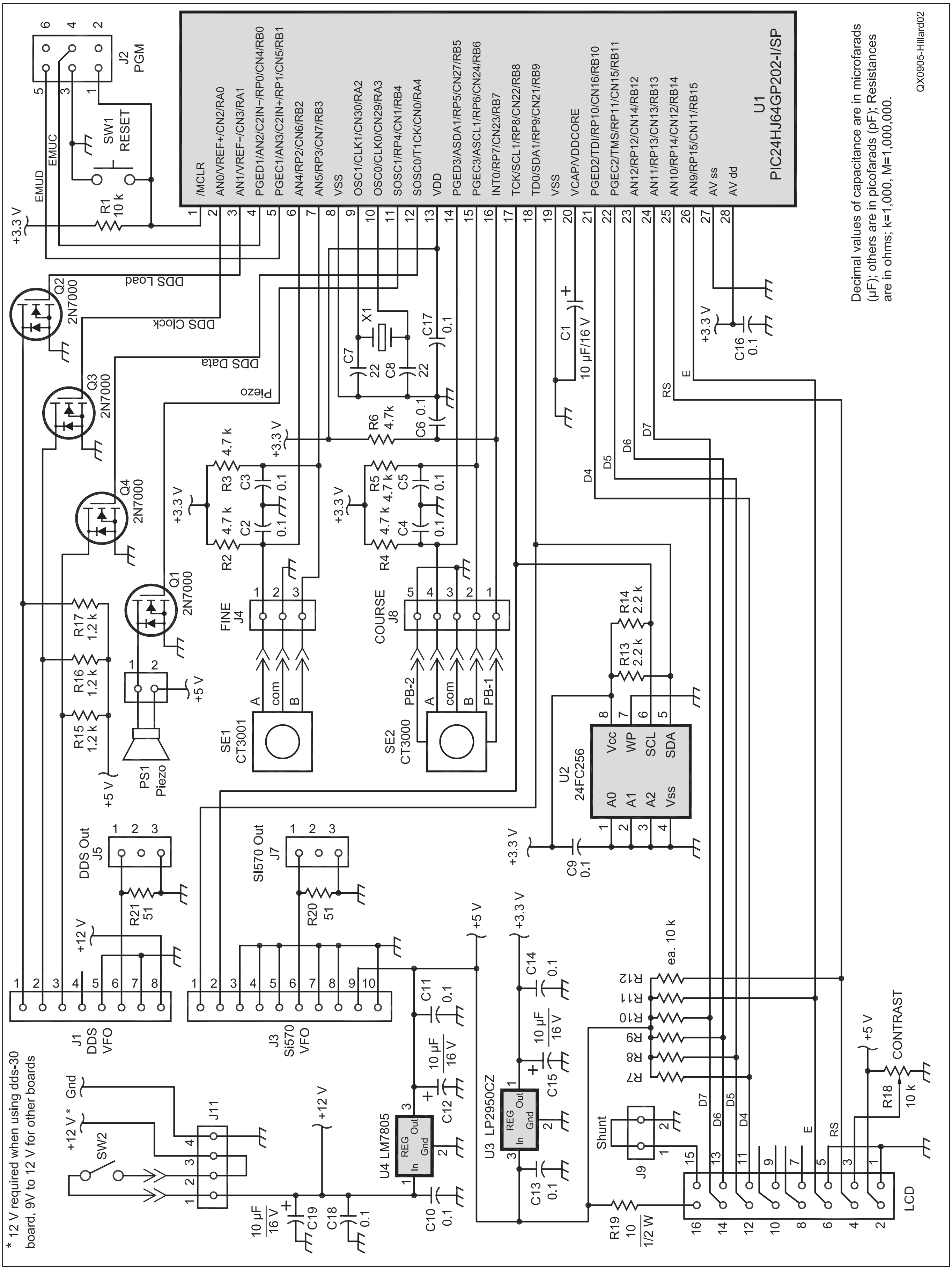
The Si570 chip is available in a number of configurations and frequency grades that must be selected at the time of order. Options such as desired frequency range, start-up frequency, temperature stability, chip address, output format, supply voltage and other options are reflected in the chip's part number. Fortunately, Silicon Labs provides a Web browser-based part number utility to simplify the ordering process at **www.silabs.com/products/clocksoscillators/Pages/BuildPartNumber.aspx**.

The part number configuration most popular with home brewers is 570CAC000141DG. Ordering this part number gets you a CMOS Si570 chip that is in the C frequency grade (10 to 160 MHz for the CMOS version), has a start-up frequency of 56.32 MHz, and a chip address of 55 hex. This start-up frequency was made popular by the software defined radio group. The universal controller can use chips with other start-up frequencies, however, since the start-up frequency is entered and saved in the controller memory at the time of set-up.

While the C frequency grade CMOS chip specification is 10 to 160 MHz, in true Amateur Radio fashion, the C grade chip may be found to operate above 160 MHz. Purchasing the more expensive B grade chip will give you contiguous output frequencies from 10 to 810 MHz; and the even more expensive A grade chip will get you to 945 MHz, and it also tunes two other groups of frequencies (970 MHz to 1.134 GHz, and 1.2125 to 1.4175 GHz). The universal controller has the capability to tune to these higher frequencies with a higher grade chip installed on the oscillator card.

Time to Experiment

Having purchased a number of the DDS-30 and DDS-60 kits, I needed to find a way to control them from my computer. So, a few years ago I wrote a Visual BASIC program called DDS Controller, and made the program available on my Web site at **home.roadrunner.com/~WA6UFQ/ddscontroller.html**. The desire for a stand-alone controller finally drove me to design and construct a number of prototype controllers, however, and to write code for a universal VFO controller. The controller was to be universal since it would be capable of controlling the new Silicon Labs Si570 programmable XO chip, as well as most of the Analog Device DDS chips.



U1
PIC24HJ64GP202-I/SP
/MCLR
AN0/VREF+/CN2/RA0
AN1/VREF-/CN3/RA1
PGED1/AN2/C2IN-/RP0/CN4/RB0
PGEC1/AN3/C2IN+/RP1/CN5/RB1
AN4/RP2/CN6/RB2
AN5/RP3/CN7/RB3
VSS
OSC1/CLK1/CN30/RA2
OSC0/CLK0/CN29/RA3
SOSC1/RP4/CN1/RB4
SOSC0/T1CK/CN0/RA4
VDD
PGED3/ASDA1/RP5/CN27/RB5
PGEC3/ASCL1/RP6/CN24/RB6
INT0/RP7/CN23/RB7
TCK/SCL1/RP8/CN22/RB8
TD0/SDA1/RP9/CN21/RB9
VSS
VCAP/VDDCORE
PGED2/TDI/RP10/CN16/RB10
PGEC2/TMS/RP11/CN15/RB11
AN12/RP12/CN14/RB12
AN11/RP13/CN13/RB13
AN10/RP14/CN12/RB14
AN9/RP15/CN11/RB15
AV ss
AV dd
J2
PGM
SW1
RESET
R1
10 k
EMUD
EMUC
+3.3 V
Q2
2N7000
Q3
2N7000
Q4
2N7000
Q1
2N7000
DDS Load
DDS Clock
DDS Data
Piezo
X1
C7
22
C8
22
C17
0.1
C1
10 µF/16 V
C16
0.1
R6
4.7k
C6 0.1
R2
4.7 k
R3
4.7 k
C2
0.1
C3
0.1
R4
4.7 k
R5
4.7 k
C4
0.1
C5
0.1
FINE
J4
COURSE
J8
SE1
CT3001
SE2
CT3000
A
com
B
PB-2
PB-1
R17
1.2 k
R16
1.2 k
R15
1.2 k
PS1
Piezo
+5 V
R14
2.2 k
R13
2.2 k
U2
24FC256
A0
A1
A2
Vss
Vcc
WP
SCL
SDA
C9
0.1
DDS Out
J5
Si570 Out
J7
R21
51
R20
51
+12 V
J1
DDS
VFO
J3
Si570
VFO
C11
0.1
C14
0.1
C15
10 µF
16 V
C12
10 µF
16 V
U4 LM7805
REG
In
Gnd
Out
U3 LP2950CZ
C10
0.1
C13
0.1
C18
0.1
C19
10 µF
16 V
SW2
J11
+12 V *
Gnd
* 12 V required when using dds-30 board, 9V to 12 V for other boards
R12
R11
R10
R9
R8
R7
ea. 10 k
Shunt
J9
R19
10
1/2 W
R18
10 k
CONTRAST
LCD
D4
D5
D6
D7
RS
E
Decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); Resistances are in ohms; k=1,000, M=1,000,000.
QX0905-Hillard02

Figure 2 — The circuit diagram of the VFO controller. (Parts are available from Mouser Electronics: tel 800-346-6873, www.mouser.com; Digi-Key Electronics: tel 800-344-4539, www.digikey.com; BG Micro: tel 800-276-2206, www.bgmicro.com/index.asp)

C1, C12, C15, C19 — 10 µF, 16 V tantalum capacitor, SMT, Mouser 74-293D106X9016B2TE3.
C2, C3, C4, C5, C6, C9, C10, C11, C13, C14 — 0.1 µF ceramic capacitor, SMT, Mouser 80-C1206C104J5R.
C7, C8 — 22 pF ceramic capacitor, SMT, Mouser 81-GRM426C220J50D.
HS1 — Heat Sink, Mouser 532-507302B00.
J1 — 8 pin SIP connector, Digi-Key 929974E-01-08-ND.
J2 — 10 pin connector, Mouser 571-5-146254-5.
J3 — 10 pin SIP connector, Digi-Key 929974E-01-10-ND.
J4, J8, J9, J10, J11 — 40 pin header stock, Digi-Key A32701-40-ND ORDER ONE , cut with razor blade knife.
J5, J7 — 3 pins of SIP Output connector, Mouser 538-22-23-2031.
J6 — 16 pin connector for LCD, Digi-Key A33163-ND.
LCD — 16×2 display, BG Micro LCD1031.
PS1 — Piezo Sounder, BG Micro AUD1082.
P4, P8, P11, P12 — Female Header w/wire, pack of four, BG Micro CAB2154 (cut to size with razor blade knife).
P5, P7 — 3 pin connector housing, Mouser 538-22-01-2037.
Crimp terminals for above (6 ea), Mouser 538-08-50-0114.
P6 — 16 pin ribbon connector for LCD, Digi-Key AKC16H-ND.
Q1-Q4 — 2N7000 transistor, NFET, Digi-Key 2N7000FS-ND.
R1 — 10 kΩ resistor, SMT, Mouser 263-10K-RC.
R2-R6 — 4.7 kΩ resistor, SMT, Mouser 263-4.7K-RC.
R7-R12 — 10 kΩ resistor array, Mouser 652-4607X-1LF-10K.
R13, R14 — 2.2 kΩ resistor, SMT, Mouser 263-2.2K-RC
R15-R17 — 1.2 kΩ resistor, SMT, Mouser 263-1.2K-RC.
R18 — 10 kΩ potentiometer, Digi-Key 3386P-103LF-ND.
R19 — 10 Ω ½ W resistor, Mouser 660-CF1/2CT52A100J.
R20, R21 — 51 Ω resistor, SMT, Mouser 263-51.
SE1 — Shaft encoder CTS, Digi-Key CT3001-ND (DO NOT SUBSTITUTE).
SE2 — Shaft encoder w/switch CTS, Digi-Key CT3000-ND (DO NOT SUBSTITUTE).
SK1 — 8 pin DIP socket, Digi-Key ED3108-ND.
SK2 — 28 pin DIP socket, Digi-Key ED3128-ND.
SW1 — Reset switch, Mouser 101-0161-EV.
SW2 — SPST On/Off switch, BG Micro SWT1106.
X1 — 10 MHz crystal, Mouser 559-FOXS100-20-LF or CITIZEN crystal: 695-HC49US-10-U.
U1 — PIC24 microcontroller, Mouser PIC24HJ64GP202-I/SP (pre-programmed PIC24 available at home.roadrunner.com/~wa6ufq//universal_vfo_controller.html).
U2 — EEPROM 24FC256, Digi-Key 24FC256-I/P-ND.
U3 — 3.3 V regulator LP2950, Digi-Key LP2950CZ-3.3-ND.
U4 — 5 V regulator MC7805, Digi-Key LM7805CT-ND.
PCB1 — Silk screened and solder masked printed circuit board (available at home.roadrunner.com/~wa6ufq//universal_vfo_controller.html).

The universal controller would have these features:

- Control cards using the Si570 XO grades A through C chips.
- Control AMQRP DDS-30 and DDS-60 cards.
- Control WB6DHW's AD995x and AD9912 boards.
- Support a 2 × 16 liquid crystal display (LCD).
- Have Coarse and Fine Tune controls using two shaft encoders:
  - Coarse:
    100 MHz-10 MHz-1 MHz-
    100 KHz-10 KHz-1 KHz-100 Hz.
  - Fine: 10 Hz.
- Start up at the last frequency used.
- Use frequency wrap-around.
- Have 99 memory channels in EEPROM.
- Have Sweep and Scan functions.
- Use a PIC24 28 pin DIP microcontroller running at 40 MIPS.
- Use a piezo sounder for audible feedback.
- Be powered by a 9 to 12 V dc supply at 250 mA.

All controller functions would be selected from the controller menu:

- Select Device (AD9850/AD9851/AD9912/AD995x/Si570).
- Set Start Frequency.
- Set Stop Frequency.
- Set Offset Frequency for superhet applications (+ or – offset).
- Set Output Multiplier to 1×, 2×, or 4× (for software defined radio (SDR) applications)
- Set Sweep Step Size.
- Set Sweep Dwell Time.
- Set Sweep On.
- Set Memory Scan Dwell Time.
- Set Memory Scan On.
- Set Master Clock Frequency.

The first controller I built was a wire wrapped version. Once the controller prototype was completed, I designed and etched an Si570 oscillator card using the same form factor as the AMQRP DDS cards. The prototype Si570 oscillator card, WB6DHW boards (**wb6dhw.com/index.html)**, and all of my DDS cards were found to perform quite well with my wire wrapped controller.

The original concept was to have only one socket that either type oscillator card could be plugged into. A number of 0.1 inch shunt jumpers would be used to configure the socket for the type of card to be used.

I decided that nothing would be gained by doing this, though, as the shunts would use almost as much real estate on the board as a second card socket would. So the design evolved into providing a socket for each type of card. Interface cables for the WB6DHW boards would plug into the DDS card socket.

The next step was to design and etch a prototype printed circuit board using the free software available from ExpressPCB (**www.expresspcb.com**). The double-sided printed circuit board that resulted shrank the size of the prototype to 4.2 × 2.7 inches.

My design used SMD components for the resistors and capacitors, but used through-hole components for all active devices, connectors, headers and so on. I chose to use a microcontroller in a DIP package for ease of construction by the average builder, and settled on the MicroChip PIC24, which is available in a 28 pin skinny DIP package. The MicroChip PIC24HJ is a high performance 16 bit microcontroller capable of 40 million instructions per second (MIPS).

Code for the Universal VFO Controller was written in *C*. The Si570 routines were developed for a Freescale microcontroller by a fellow ham and good friend, John Fisher, K5JHF. With a little "tweaking," I was able to make his Si570 routines work in the PIC24 environment.

Building It

Solder all SMD components to the printed circuit board prior to soldering any through-hole components. Using this sequence ensures that the printed circuit board will remain stable on your workbench while the SMD components are being soldered to the board. For best results, use a very small-tipped iron and 0.015 inch diameter solder for the SMD components (my printed circuit boards are shipped with a hank of 0.015 inch solder).

The controller printed circuit board, the Si570 printed circuit board, and a pre-programmed PIC24 are available from my Web site at **home.roadrunner.com/~wa6ufq//universal_vfo_controller.html**. Sockets are recommended for the PIC (U1) and the EEPROM (U2). The two regulators, the crystal, and the 2N7000 FETs may be soldered directly to the printed circuit board, however. I used 0.1 inch header stock for connecting to the two shaft encoders (J4 and J8), RF out (J5 and J7), piezo sounder (J12), LED backlight on /off (J9), and the power connector (J11), but wires from these components could be soldered directly to the pads on the board, eliminating the headers. Do not substitute other shaft encoders for SE1 and SE2; the CTS encoders are the only ones that will work correctly with the controller. Construction Guidelines for the controller and oscillator board can also be downloaded from my Web site.

The Si570 oscillator card uses a ten-pin right angle connector (J1) to mate with its controller board socket. This connector is the only through-hole component on the oscillator card, and should be installed after all SMD components have been soldered to the board. The Si570 chip is a CMOS device, and should be handled accordingly.

Putting the Controller to Work

The controller can be configured for any of the DDS cards or the Si570 card by plugging the desired card or interface cable into the appropriate socket on the controller board, turning the unit on, and choosing Select Device from the menu. In fact at start up, if you have the wrong type card installed, the display will ask you to "Set Device Type." Once configured for one of the card types, the controller retains that configuration until the card is changed.

Selecting the Set Clock function from the menu configures the master clock frequency for the type of card in use. The default values of 100 MHz for the DDS-30; 180 MHz for the DDS-60; 400 MHz for the AD995x board; 1000 MHz for the AD9912 board, or the start-up frequency for the Si570 can be used. For a more accurate frequency display, however, the DDS card clock frequency can be measured with a frequency counter and entered into the controller memory. The entry for the Si570 start-up frequency can be adjusted ± a few Hz to correct for any possible slight frequency error of the Si570.

Two rotary shaft encoders, FINE and SELECT, are used to perform all controller functions, and to select all features of the controller. The FINE encoder is a vernier tuning control, providing frequency control in 10 Hz increments. The SELECT encoder performs a number of functions including coarse frequency control. It has detents that divide the shaft rotation into sixteen positions, and a momentary switch that is activated by pressing on its shaft.

The controller wakes up with an underline cursor at the 1 kHz digit of the LCD. This indicates that rotation of the SELECT control will change the VFO frequency in 1 kHz increments. By pressing on the encoder's shaft and rotating it clockwise or counterclockwise, the underline cursor can be moved to any of the other digits on the display, and will cause the VFO's frequency to change at whatever rate is indicated by the position of the cursor on the display.

Pressing on the SELECT control's shaft and rotating it clockwise until the cursor is located under the ">" character on the far right position of the display's top line puts the controller in Menu mode. Then by rotating the SELECT control, the display increments through each menu item. When the desired menu item is found, pressing on the encoder's shaft selects that menu item.

In a similar manner, pressing on the SELECT control and rotating it counterclockwise moves the cursor to the "<" character on the left side of the display and puts the con-

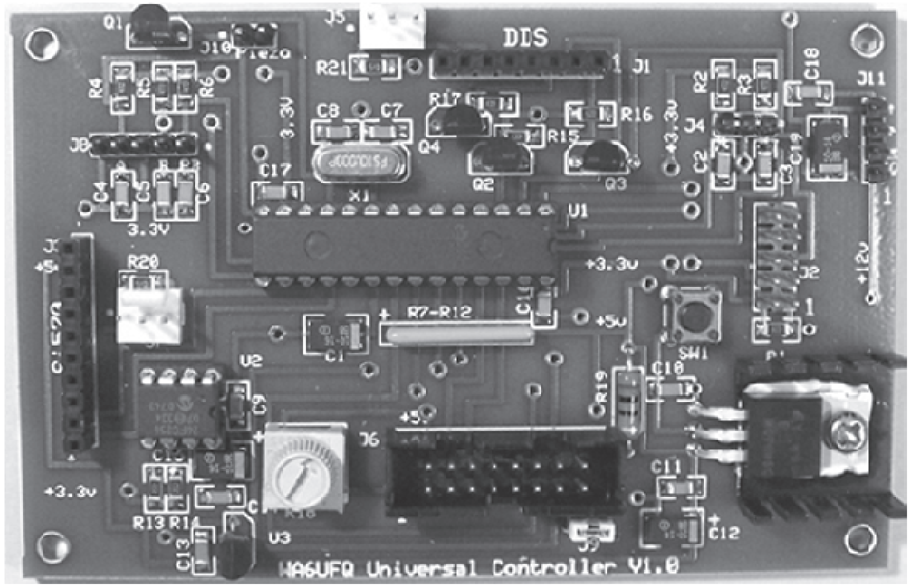


Figure 3 — This photo shows the controller board. The socket for the Si570 oscillator card is located on the left side of the board; the socket for the DDS oscillator card or the interface cable for the WB6DHW boards is located at the top of the board. The PIC24 is in the center of the board.

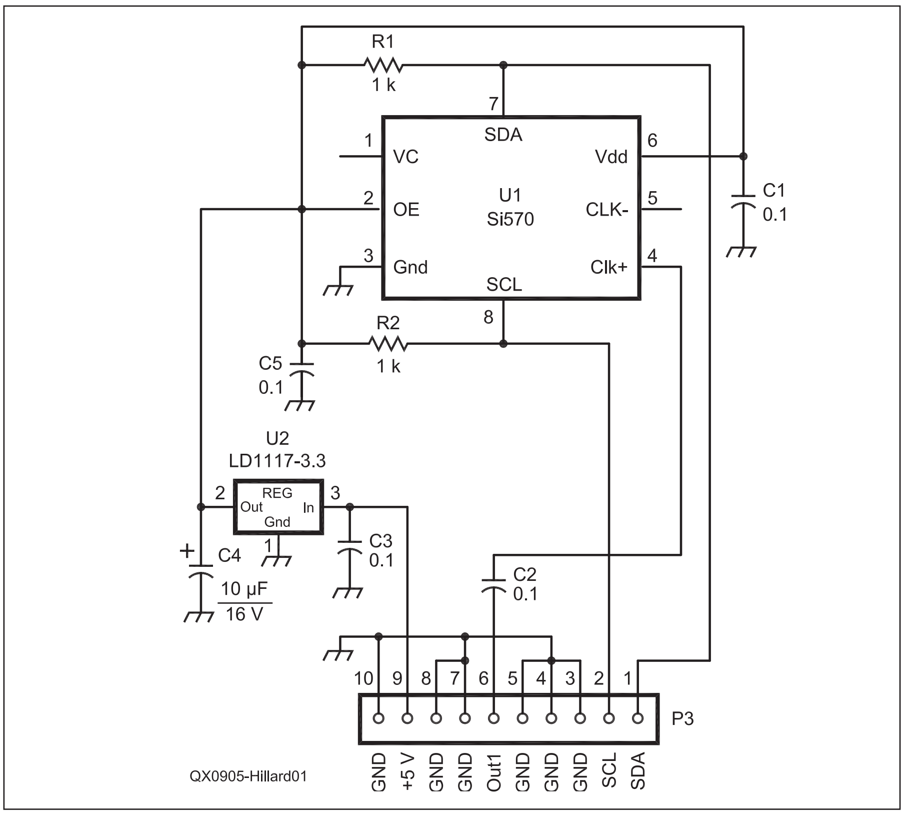


Figure 4 — Circuit diagram of the Si570 oscillator card. (Mouser Electronics: tel 800-346-6873, www.mouser.com).

C1, C2, C3, C5 — 0.1 µF ceramic capacitor, SMT, Mouser 80-C1206C104J5R.
C4 — 10 µF, 16 V tantalum capacitor, SMT, Mouser 74-293D106X9016B2TE3.
J1 — 10 pin right angle header, Digi-Key WM6110-ND.
R1, R2 — 1 kΩ resistor, SMT, Mouser 263-1.0K-RC.
U1 — Si570 oscillator chip, 570CAC000141DG, Available from Tom Hoflich, KM5H; www.softrockradio.org/si570.
U2 — 3.3 V regulator, LD1117-3.3, Mouser 511-LD1117S33.
PCB2 — Silk screened and solder masked printed circuit board (available at home.roadrunner.com/~wa6ufq//universal_vfo_controller.html.

troller in Memory mode, allowing the Mem Read, Mem Write, or Mem Erase functions to be selected.

When the Scan function is selected from the menu, the frequency of the VFO continuously sweeps from the Start frequency to the Stop frequency. There are menu items that allow you to set the rate that the VFO sweeps, as well as the amount of dwell time between each frequency step. A three-millisecond pulse at the start of the sweep is available for triggering other devices.

Selecting Memory Step from the menu, steps the VFO through each of the memory channels that have been saved in the controller EEPROM. A menu item is also available to set the amount of dwell time between each step.

If the controller is to be used as a VFO in a superhet radio, it can be configured from the menu using the Set Offset Frequency function to offset the frequency displayed by the amount of the radio's intermediate frequency (IF), allowing the controller to display the actual operating frequency of the radio. Likewise, the Set Multiplier function can be used to configure the controller if it is to be used in an SDR application, when the SDR clock is required to be twice or four times the actual operating frequency.

Powering the Controller

The controller can be powered from a 9 to 12 V dc source capable of supplying 250 mA of current. In fact the only time that 12 V is needed is if a DDS-30 card is used that has the Rev 1.4 update kit installed. This update added a second regulator for the output amplifier on the card.

Summary

The Universal VFO Controller can be used as a signal source on your workbench, or it can be imbedded in your SDR or Superhet radio project. If used with the Si570 oscillator card, it gives you access to a new and exciting technology developed by Silicon Labs. The controller can also be used with the popular AMQRP Club DDS oscillator cards available through the AMQRP club, and boards using Analog Devices AD9912 and AD9951 — AD9954 DDS chips are available from WB6DHW (**wb6dhw.com/index.html**).

Silk screened and solder masked printed circuit boards for the controller and Si570 oscillator, a pre-programmed PIC24 microcontroller, and the object code for the controller are all available at **home.roadrunner.com/~wa6ufq//universal_vfo_controller.html**. The C grade Si570 chip is available from Tom Hoflich, KM5H, at **www.softrockradio.org/si570**. Tom does group buys of the Si570 to keep the price of the chip affordable. The AD995x and AD9912 boards are available from David Brainerd, WB6DHW, at **wb6dhw.com/**.

There is a Yahoo! Groups site named Universal VFO Controller for builders and users of the VFO controller to come to discuss construction of the unit, and its applications and features.

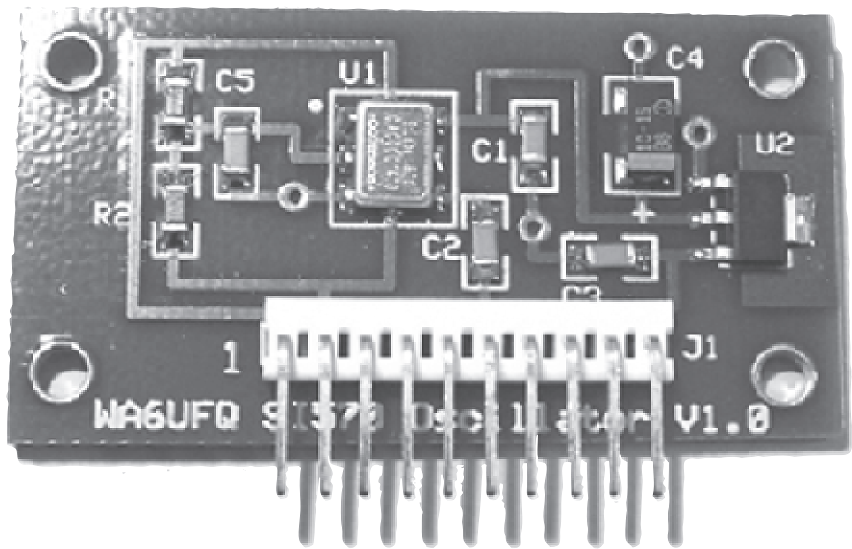


Figure 5 — This photo shows the Si570 oscillator card. The Si570 chip (U1) can be seen in the center of the plug-in card.

Bob Hillard, WA6UFQ, was first licensed in 1965 as WAØJBH. Bob is an ARRL member, and a member of the Austin QRP Club. He is also an avid homebrewer. In 1999, he retired from AT&T Broadband as the Director of Engineering for the Salt Lake City area, and then in 2002 relocated to Austin, Texas.

Rudy Severns, N6LF

PO Box 589, Cottage Grove, OR 97424; n6lf@arrl.net

Experimental Determination of Ground System Performance for HF Verticals Part 4 How Many Radials Does My Vertical Really Need?

Experimental results to answer an often-asked question.

A frequently asked question is "How much of a ground system do I really need to make my vertical antenna work?" Usually, what's wanted is an answer in the form of "This much ground system will improve your signal by X dB." Another common question is "Does it matter if I lay the radials on the ground surface instead of burying them?" This is a practical consideration because it's often much easier to lay out the radials on the surface and let them vanish into the grass.

These questions can be addressed analytically and with modeling, but for most of us that's not very convincing. It's much more satisfying to see actual field measurements on real antennas. In the past there has been professional work at MF broadcast frequencies and also the excellent work by Jerry Sevick, W2FMI, at HF.[1, 2] The problem with an experimental approach is the practical limit on the number of test examples: you can't do all the possible variations! What's needed are reliable field measurements that can be compared to calculations and/or modeling to see if there is reasonable correlation. If there is, we can use calculations or modeling for the wide variety of antennas and soil characteristics we which we couldn't test.

Some of the material that follows represents a redo of Sevick's work with better instrumentation, but the material in this section, along with the other five parts of the series, goes well beyond Sevick's work. The details of the test equipment and experimental setup were given in Part 1 of this series.[3]

Efficiency Limitations

The purpose of the ground system is to improve antenna efficiency so that less power is lost in the soil and more is radiated. Efficiency is the ratio of the power radiated to the total input power at the feed point. Of course what we want is to radiate all the input power (100% efficiency) and maximize our signal, but there are practical limits. We can represent the resistive part of the feed point impedance (Rs) by three series resistors as shown in Figure 1.

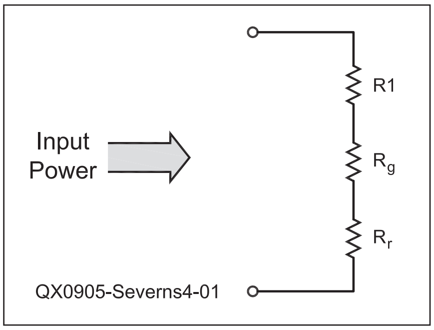


Figure 1 — An antenna input equivalent circuit. Rl represents the ohmic loss due to conductors, loading inductor series resistance, and so on. Rg represents the power dissipated in the soil by the near-field of the antenna. Rr is the radiation resistance, which accounts for the radiated power.

The input resistance at the feed point is Rs = Rr + Rg + Rl. We have to be a bit careful what we mean by "radiation resistance." Rr is usually defined as the value of the resistance at a current maximum attributable to radiation. In a vertical antenna with a height of ¼ λ or less over perfect ground, this point is at the base of the antenna, which is the usual feed point. In real antennas with various numbers of ground surface radials, however, the height of the antenna may have to be modified to maintain resonance and the current maximum may actually be out on the radials or possibly even back up into the vertical. What this means in practice is that the fraction of the feed point impedance we attribute to Rr may not be converging to the ideal value from theory as we add radials or change radial lengths. For example,

[1]Notes appear on page 42.

a resonant, very thin ¼ λ vertical over perfect ground will have Rr = 36.2 Ω but a real antenna may converge to a somewhat different value as we add radials and reduce ground loss.

With a ¼ λ vertical it is often assumed that if Rl is small, then Rg is simply Rs – 36.2 Ω. This is not the case and should not be assumed. The radiation resistance varies as the ground system changes, and does not approach 36 Ω until the ground system is relatively large. In a broadcast antenna with 120 radials 0.4 λ long, this approximation is very good, but in the limited ground system typical of amateur antennas at HF, it is not. A detailed discussion of this point can be found in an article available on my Web site, "Radiation Resistance Variation with Radial System Design." [4] (This may become a *QEX* article in the future.)

Because we are interested in the effect of efficiency on signal strength, it is handy to express efficiency (η) in terms of dB:

$$\eta = 10\text{Log}\left(\frac{1}{1+\dfrac{Rg}{Rr}+\dfrac{Rl}{Rr}}\right)$$

[Eq 1]

For 100% efficiency, Rl = Rg = 0 and η = 0 dB. If we increase Rl and/or Rg, η will decrease. For example 80% efficiency would be about –1 dB.

Experimental Tests

All of the measurements were made on 40 m, 7.2 MHz in most cases. I chose 40 m verticals for their manageable size. Even at that size, the ground system that had to be laid down and taken up numerous times, required over 2000 feet of wire.

I used five different antennas:

- A ¼ λ, 1 inch aluminum tubing vertical, adjusted to resonate at 7.2 MHz.
- An 1/8 λ, 1 inch aluminum tubing vertical with three top loading wires sloping at roughly 45°, again, resonated at 7.2 MHz.
- An 1/8 λ, 1 inch aluminum tubing vertical with no top loading, but resonated to 7.2 MHz with a base inductor.
- A 40 m Hamstick mobile whip (about 7.5 feet high), the top section adjusted for resonance at 7.2 MHz.
- A Cushcraft R7000 vertical.

The minimum conceivable ground system for a vertical would be a single ground stake with a coaxial feed line back to the shack. In this case, the feed line acts as a single random length radial. For these measurements I adopted this as the "zero radial" system, where the stake was a 4 foot copper-clad steel rod with ½ inch Andrews Heliax, buried 6 inches below the ground surface, back to the shack. The ground system was improved progressively by adding 33 foot (no. 18 AWG) radials in the progression: 0, 4, 8, 16, 32 and 64. This was repeated for each antenna. A ¼ λ in free space is close to 33 feet at 7.2 MHz. As was shown in Part 2, however, the *electrical* length of the radials changes when the radials are placed close to the soil.[5]

The soil characteristics under the radial system were measured using the technique given in *QEX*.[6] The average soil constants in the test field were: conductivity, $\sigma = 0.02$ S/m and relative dielectric constant, $\varepsilon_r = 30$. I will refer to this as "N6LF soil."

For each number of radials and each antenna, two measurements were made: the input impedance and the relative signal strength at a point 1.8 wavelengths away from the test antenna, at an elevation angle of about 8 degrees. Because the number of radials affected the resonant frequency, each antenna was re-resonated by adjusting its height as the number of radials was changed.

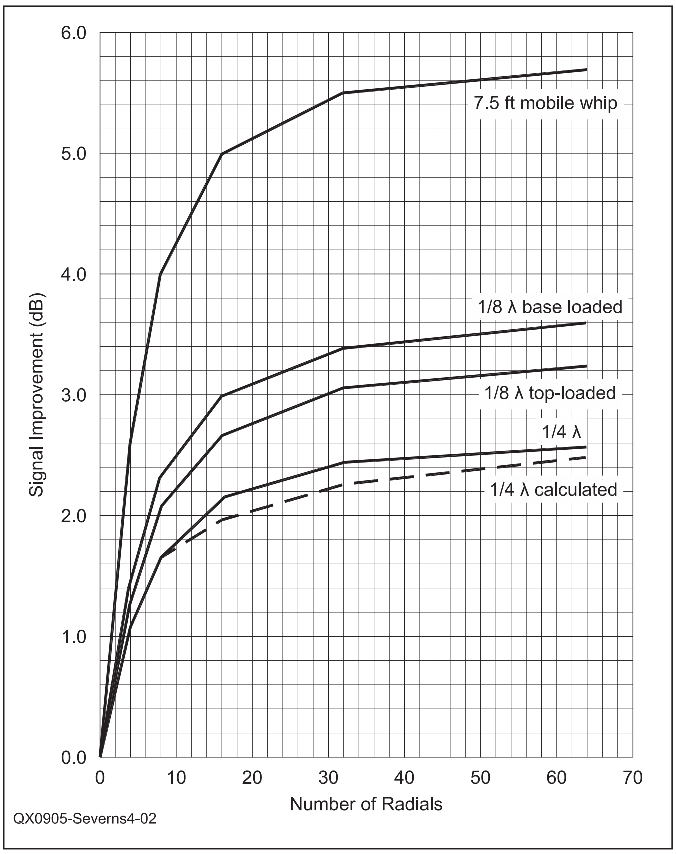


Figure 2 — Typical improvement in signal as ¼ λ radials are added to the basic ground system (a single ground stake).

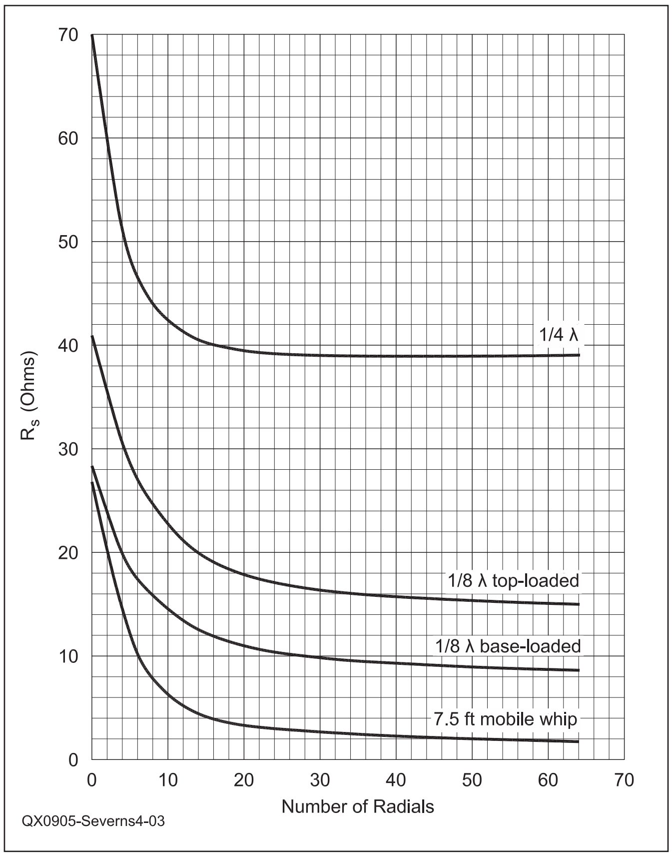


Figure 3— Measured input resistance (Rs) at resonance as a function of the number of radials.

Experimental Results

When we compare the results for different numbers of radials on a given antenna, the change in relative signal strength directly answers the question of how much signal improvement we get by adding radials. Typical test results are shown in Figure 2.

Note that the graph is in terms of the *improvement* in signal over the single ground stake with no radials for *each* antenna. The graph does *not* compare the relative worth between each antenna. Obviously a short, lossy mobile whip will yield much less signal (–10 dB or worse!) than the ¼ λ vertical.

The effect of radial number on input resistance (Rs) is shown in Figure 3.

In the case of the Hamstick mobile whip, I have subtracted Rl from the measured input resistance because it has a fixed value independent of radial number. Rl is determined by the loading coil Q. We can see that as we add larger numbers of radials the values for Rs begin to level out and approximate, but not equal, values for ideal lossless antennas.

Interpreting the Data

One of the interesting things about Figure 2 is that it shows that *the shorter and more heavily loaded the antenna, the more you have to "gain" from an aggressive ground system*. For example, the improvement for the ¼ λ vertical, going from 0 to 64 radials, is about 2.6 dB, but for the 1/8 λ base loaded vertical it's more like 3.4 dB, and for the mobile whip, nearly 6 dB.

What's going on here? As I pointed out in my July 2000 *QST* article on ground systems, when we shorten an antenna but keep the input power the same, both the magnetic and electric field intensities in the immediate vicinity of the antenna increase dramatically.[7] This translates to much higher ground losses. What we see in Figure 2 is that adding the radial system reduces these losses, but since the losses are higher to start with for the shorter antennas, the improvement is greater. No mystery!

From Figure 2 we can see that for all the test antennas, most of the improvement comes with the first 16 radials. As we add more radials beyond 16, there is still improvement but it is proportionately smaller. You gain perhaps another fraction of a dB going to 32 radials but by the time you reach 64 radials there isn't much change. The broadcast standard of 120 radials 0.4 λ long is hard to justify for amateur use, particularly given the present price of copper wire!

Figure 2 also has a dashed line very close to the curve for the ¼ λ vertical. This is a prediction using Abbott's calculation method.[8] I could have also added calculated lines for the other antennas and would have seen the same reasonable correlation, but that would have really cluttered the graph so I left them off.

We do have to be a little careful in using these graphs as general guides. They represent experimental results over my particular soil, at one frequency. Can we really draw any general conclusions? In lieu of running tests on all possible soils, we can get a feeling for this by calculating the signal improvement for different soils using Abbott's calculation method. (See Note 8.) Typical calculated results for different soils, at 7.2 MHz, are shown in Figure 4. This graph starts at 8 radials and goes to 64 radials. Smaller numbers of radials are omitted because the underlying calculation becomes inaccurate as the angle between the radials increases beyond 45°, the 8 radial case. From a practical point of view this is not a serious limitation. As I pointed out in Part 2 in the Jan/Feb 2009 issue of *QEX* (see Note 5), and as the data in Figure 2 shows, a four-radial ground system has very minimal performance; 8, or better yet 16 radials, should be the minimum, except perhaps in an emergency.

For the soil over which these tests were done (N6LF), the calculated 8 to 64 radial change is about 0.8 dB. Going back to Figure 2 we see that the measured change for the ¼ λ vertical is 0.9 dB (8 to 64 radials). The calculation agrees quite well with the measurements. Figure 4 tells us that when the soil is better, a given number of radials gives somewhat less improvement and with poorer soils there is more improvement. Again, no surprise. If you have better soil, you have lower losses to start with, so the improvement will be less. But even with very good soil it's still worthwhile to use at least 16 radials.

What about frequencies other than 40 m? There are a couple of complications to extending the 40 m work to another band. First, the graph in Figure 4 does not scale directly with frequency because the field intensity at a given distance (feet or meters), for a given base current, does *not* scale linearly with frequency. Second, at a given site the ground characteristics will vary with frequency. (See Note 6) The result is that the ground loss is not the same for the scaled antennas at other frequencies, even though the input power may be similar.

As we go down in frequency, soil conductivity typically decreases, which tends to increase ground loss but the relative dielectric constant goes up, which tends to decrease ground loss. For N6LF soil at 7.2 MHz, σ = 0.020 S/m and ε_r = 30, but at 1.8 MHz, σ = 0.013 S/m and ε_r = 68. The net effect on signal improvement (8 to 64 radials) is shown in Figure 5.

If you examine Figures 2 and 3 closely and compare the curves for the ¼ λ vertical, you may see something funny going on. In Figure 2, even when we go from 32 to 64 radials, there is still some improvement in signal. But if you look at Figure 3, there appears to be no change in Rs, so how can the antenna be more efficient? This same paradox shows up in the Brown, Lewis and Epstein data (see Note 1) taken 70 years ago, and has been the subject of comment ever since. What's going on? Several things are going on simultaneously. First, the number of radials is increasing, which reduces Rg. Second, we are steadily increasing the height

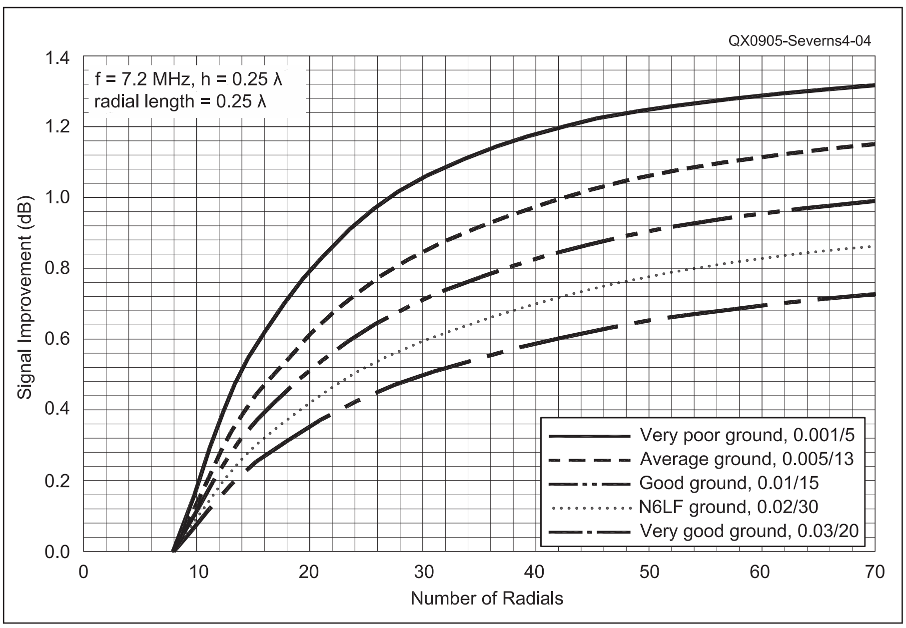


Figure 4 — Calculated signal improvement as we vary the number of radials over different soils with a ¼ λ vertical with ¼ λ radials at 7.2 MHz. Note: 0 dB is for the 8 radial case.

of the antenna to re-resonate it due to the effect of the radials on the ground, which we will look at shortly. This tends to raise Rs. In the case of the measurements for the ¼ λ antenna, the two effects cancel to some extent. Notice that for the other antennas, Rs is still trending down as signal strength goes up with number of radials. Altering the height as we add radials is not the full story, however, Rr is also affected by the radial system. (See Note 4.)

Additional Tests

In addition to the tests where antenna height and number of ¼ λ radials were the variables, I ran a few others. In one, I compared the performance of the ⅛ λ top-loaded vertical with 64 radials, with and without, an ⅛ λ circular ground screen (diameter = 36 feet) added over the radial fan. The addition of the ground screen made no detectable difference, which is in line with previous work. See Note 1. Obviously, if you have only a few radials, then a ground screen would help.

Modeling of gain versus radial number and radial length indicates that a larger number of shorter radials may be just as good or better than fewer longer radials, assuming both radial systems use the same amount of wire.[9] To check this out I ran a test using the top-loaded ⅛ λ vertical, comparing sixteen ¼ λ (33 ft) radials versus thirty two ⅛ λ (17 ft) radials. In line with the modeling and also calculations, the signal strengths were almost the same. The feed point impedances were substantially different however. I had to lengthen the vertical to re-resonate it with the 32 short radials. This is a good example of the interaction between the feed point impedance and the radial system. If space is restricted, then more short radials in place of fewer long radials may work just fine, but to properly evaluate that option it would be best to do the modeling or calculation for a particular vertical and soil characteristics.

I made measurements on the R7000, with and without an external ground system, which showed that adding a 64 radial ground system had almost no effect on signal strength (+0.1 dB). This surprised me until I had an e-mail conversation with Joe Reisert, W1JR, the original designer. The antenna was designed to work without a ground system and although the antenna is physically less than ¼ λ on 40 m (25 ft), the loading is arranged so that it behaves more like a ⅜ λ. There are a set of 48 inch radials at the base, which are isolated from ground. The current maximum is well up into the antenna and the base is a high impedance point. The conventional wisdom, to which I have been a subscriber, is that even with a ½ λ vertical, adding an extensive ground system

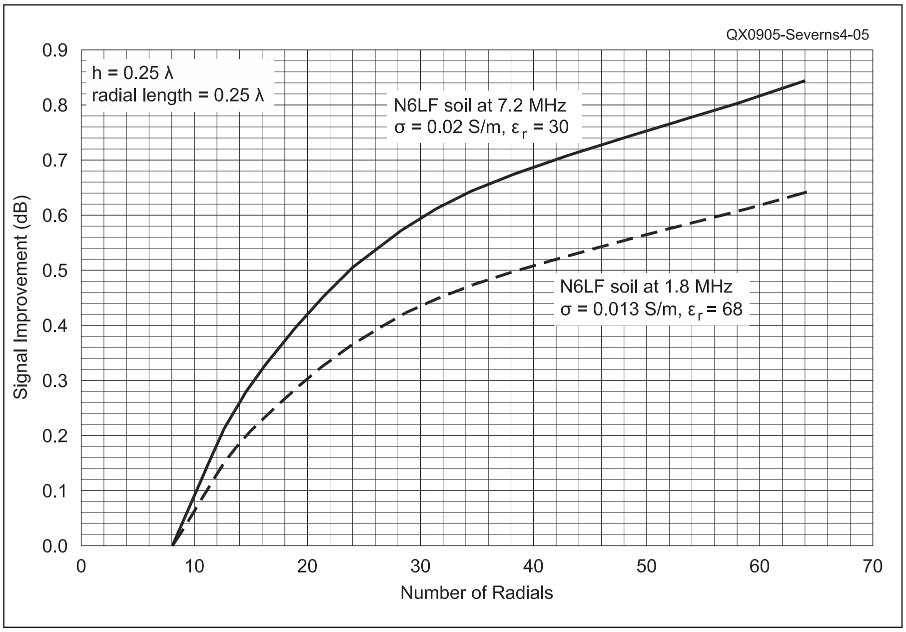


Figure 5 — Difference in signal improvement between 1.8 and 7.2 MHz over N6LF soil using the same vertical height and radial length in wavelengths (scaled with frequency). 0 dB is for the 8 radial case.

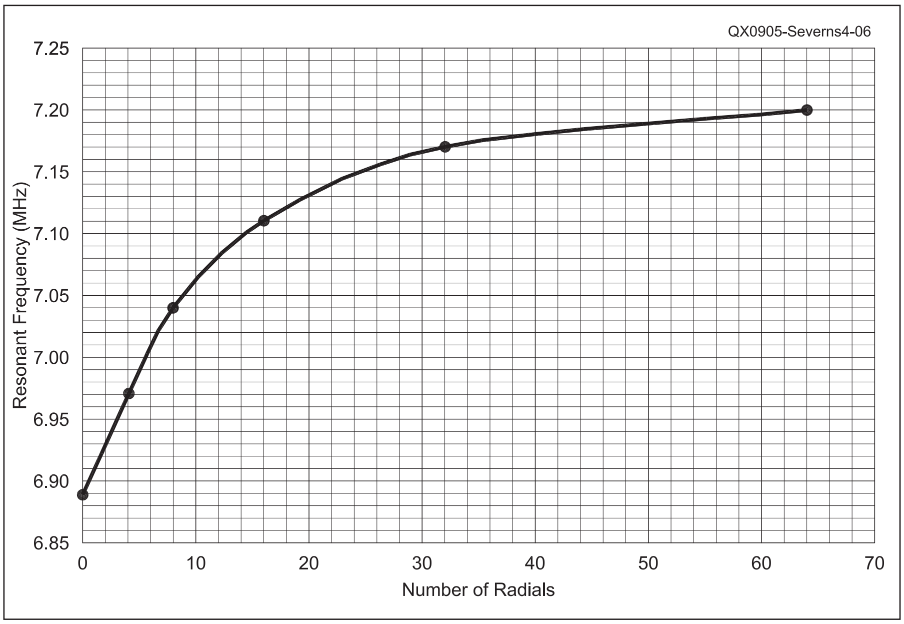


Figure 6 — Resonant frequency of a vertical antenna resonated at 7.2 MHz with sixty four 33 foot radials, as a function of the number of radials.

will improve performance. I did not see that here. This is a subject for more experiments, perhaps.

Measured Resonant Frequency

During the experiments, I found that changing the number of radials changed the resonant frequencies of all the antennas except the R7000. For example, using the ¼ λ vertical, I laid down 64 radials and adjusted the height of the vertical so that it was resonant at 7.2 MHz. I then started removing radials (but not changing the height), measuring the resonant frequency as I went down to zero radials. The results are shown in Figure 6.

Obviously the resonant frequency is affected by the radials. You can of course re-

resonate the antenna by changing its height or loading. During the experiments for signal strength and input impedance, I adjusted the height to restore resonance at 7.2 MHz. With 64 radials resonance at 7.2 MHz was obtained with h = 33 feet 7 inches. With no radials, the 7.2 MHz resonant height was 32 feet 11 inches, 8 inches shorter.

What's going on? When there are no radials, only the ground stake and the random length of feed line, the resonant frequency is low primarily because the upper portion of the stake effectively adds to the antenna height. Even though the stake is driven into the soil, the top layer of soil, at least in summer when these measurements were made, is quite dry. The effective ground surface is actually somewhat below the physical surface. There was also some inductance in the lead connecting to the ground stake. As we add radials this effect is reduced but only slowly because, as shown in Part 2 (see Note 5), the radials are heavily loaded by their close proximity to the soil. They are resonant below 7.2 MHz so they are inductive at 7.2 MHz. This shunt inductance is across the base of the antenna. As we add more radials we are adding more inductors in parallel, which reduces the effective reactance and increases the resonant frequency.

Conclusions

The answer to our original question, "*Does laying the radials on the surface matter?*" is a little clearer now. *For the same number of radials of the same length*, the efficiency will be pretty much the same whether buried or on the surface, but the effect on feed point impedance may be somewhat different. This can become a practical problem if the antenna tuning varies with the season (wet or dry or frozen ground). Radials lying on the ground surface really behave more like elevated radials even though they may be lying right in the dirt.

We can summarize all this with the following advice:

- Try to use at least sixteen ¼ λ radials.
- If you don't have the space for ¼ λ radials, lay down a larger number of shorter ones.
- More than 16 radials will help but give only a fraction of a dB over average or better soils.
- The shorter your antenna, the more you need a good ground system.
- The poorer your soil the more you need a good ground system.
- A surface-radial ground system will affect the resonant frequency and you may have to adjust the vertical height for that.
- Work hard at making the antenna itself more efficient. In other words,. use high-Q loading coils, use top loading to minimize the size of loading coils, minimize conductor loss, and so on.
- Modeling and calculations seem to be in reasonable agreement with measurements and, with some caution, can usefully be used to estimate the magnitude of improvement when adding to a ground system.

Acknowledgments

This work was inspired by the classical articles by Jerry Sevick, W2FMI, which have served us so well.[2, 10, 11, 12, 13] In many ways my experiments are just an update and reconfirmation of Sevick's work.

I want to thank Mark Perrin, N7MQ, for his help in making many of the measurements. Especially helping to drag the monstrously unwieldy chicken wire ground screen into position and out again.

In addition to the references already cited in this article, I have included several more related references, which the reader may find useful. See Notes 14 through 21.

Notes

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[5]Rudy Severns, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 2," *QEX* Jan/Feb 2009, pp 48 - 52.

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Rudy Severns, N6LF, was first licensed as WN7WAG in 1954 and has held an Extra class license since 1959. He is a consultant in the design of power electronics, magnetic components and power-conversion equipment. Rudy holds a BSE degree from the University of California at Los Angeles. He is the author of two books and over 80 technical papers. Rudy is an ARRL Member, and also an IEEE Fellow.

QEX

Ray Mack, W5IFS

17060 Conway Springs Ct, Austin, TX 78717; w5ifs@arrl.net

SDR: Simplified

I had fully intended to address setting up the Blackfin Stamp and doing our first real DSP application in this issue. The task of setting up the hardware and software is more than it appeared on first look. So, in this issue, we will only have enough time to finish our journey to the "dark side" of being software people.

I do software development as a big part of my day job, so I am used to the documentation not really matching the software or hardware. I wasn't disappointed here. The Stamp comes with a CD that contains documentation and three sets of software. I'll point out what is useful and what to ignore as we go through the process. This is not a criticism of those doing the work. The project is done completely by folks who volunteer their time to put the software together for us to use for free. We owe them a big vote of thanks.

Development System Hardware Setup

I discovered some more things we will need to be able to work with the Stamp, which were not mentioned in the first issue. You will want a serial port connection on your *Windows* computer. Desktops come with them from the factory, so you just need a straight-through 9 pin male to 9 pin female cable. Modern laptops no longer include floppy drives or serial ports. If you intend to use a laptop without a serial port, you will want to buy (or build) a USB to serial port converter.

You will need some way to attach the network port of the Stamp to your *Windows* system. This can be as simple as a cross over cable, but I recommend using a more capable network. You likely have either a DSL modem or a cable modem to connect to the Internet. I recommend that you add a router between the modem and your computer and Stamp. I did a quick check of **frys.com** and found the cheapest wired or wireless router costs about $30 new. The network will also require a short CAT-5 cable to connect the Stamp to the router.

The router default setup will work for you right out of the box. Most home routers have four ports. This means you have five connections on the back. The one labeled WAN connects to your modem. You will plug the computer and the Stamp into any two of the other connections.

All computer networks assign an identifier called an IP address to each connection. The router allows us to ignore everything that connects to your modem and focus on the computer side. The standard for home routers is to use IP addresses that are 192.168.1.NNN. "NNN" is a number between 0 and 255. The router will be setup so it uses 192.168.1.1. You almost certainly have your computer set up to automatically get its IP address from the Internet Service Provider (ISP) using DHCP. This configuration is also what is required by the router, so you don't have to change anything on your computer to add a router. My router starts assigning automatic addresses at 100, so my main computer usually gets 192.168.1.100 as its address. All those numbers between 1 and 100 are available for you to assign as you need. The networking world calls these static addresses. We will use the IP addresses extensively later.

If you eventually want to do serious work with the Stamp operating system, I suggest getting an old, cheap desktop computer to run *Linux*. We have a Goodwill Industries store in Austin where you can find old computers and routers for significantly cheaper than buying new. Most towns also have used computer stores now. I got lucky when I put out the word among friends that I was looking to rescue an old computer, and got a nice one for $50.

As an alternative to a dedicated *Linux* computer, you can run a version of *Linux* on your *Windows* machine, but it is not without its problems. The Blackfin CD contains a directory called *coLinux*. This is a free version of *Linux* that runs as part of *Windows*. The documentation warns that it is a test version, and that it can slow down your *Windows* system. They aren't kidding! I installed *coLinux* from the CD to see how it works. I run *Windows 2000* on an old Dell laptop with only 384 MB of memory. When you subtract memory for *coLinux*, there isn't much left for *Windows* and it slows down considerably. If you have a newer computer with a lot more memory, you will likely lose less performance. A more pricey option ($189.00) that is also safer is to use *VmWare Workstation* and run *Linux* as a virtual computer on top of *Windows*.

If you run server versions of *Vista, XP* or *2000*, another option is to download the free *VMWare Server* software and install that, and then run *Linux* as a virtual computer on top of *Windows*. Robert Kluck, N4IJS, told us how to do that in the May/June 2008 issue of *QEX*.[1]

I prefer to keep my *Windows* machine safe from the bad things that can happen in software development, so I will use a dedicated *Linux* computer. I anticipate that it will be quite a while until I am ready to tackle that level of *Linux* work, so the *Linux* machine is just sitting in the garage for now.

Overview of the Stamp

The Stamp is a fully functional general purpose computer very similar to your *Windows* computer. It has a main CPU (the Blackfin 537), a reasonable amount of memory, a solid state "disk drive," a network connection, a command prompt through the serial port, and expansion ports (roughly equivalent to the PCI bus in a PC).

The "BIOS" on the Stamp is actually more full featured than the one in a typical PC. The BIOS is called u-boot. There are binary files and source files for u-boot on the CD under the u-boot directory. We will not be changing the boot loader, so you can ignore these files. U-boot allows you to load a new version of the operating system if you need it. U-boot is the first program to run when the Stamp board starts. It uncompresses the RAM disk version of the operating system and then loads the operating system into memory.

Once the operating system is finished initializing, you have a fully functional *Linux* computer. If we ever need to reload the operating system, the files are on the CD under the *uClinux* directory on the CD.

Notice that we have mentioned two different *Linux* systems that are on the CD. The *coLinux* system is designed to be a cross development system. "Cross development" means that we will write and compile programs on a computer (our PC), but those programs will not run on that computer. Instead, they work on a completely different computer (the Stamp in this case). The *uClinux* system is designed to work on microcomputer systems, hence "µC". In general, you would not develop working programs on a *uClinux* system, but you can. You also would not typically run *uClinux* on a PC, but you could.

Target Hardware Documentation and Setup

The root directory of the CD has a Readme.txt file. It describes the four items on the disk. The docs directory holds a set of the Web pages from 2007 from the Blackfin uCLinux Web site and, in a normal world, would be the place to start.

I am really a hardware guy, so the first thing I wanted to look at was the schematic of the board. The label on the CD says that

[1]Robert Kluck, N4IJS, "Linux Under Windows," *QEX*, May/Jun 2008, pp 25 – 32.

it contains the schematics for the Stamp board. I spent a lot of time looking for them on the CD; they don't exist. You will need to go to the Web page: **blackfin.uclinux.org/gf/project/stamp/frs/** and look for the Stamp schematic toward the bottom of the page.

Don't be put off by the title page on the schematic. The schematic you will be looking at is really for the BF537 Ez-Kit Lite. In order to leverage the design process, the Stamp and Ez-Kit use the exact same copper layout. The difference is that the Stamp has fewer devices populated. Sheets 4,5, and 8 cover hardware that is not on the Stamp. These cover the audio ADC, audio DAC, and the ELVIS (External Live Video System) for doing TV work. All other pages are the correct schematic for the Stamp board.

Ignore the document **/docs/bf537_quick_start.html** on the CD. The only correct information is the serial port set up. Instead, look at **/docs/bf537-stamp_board.html**. It has a correct picture and describes the setup better. Ignore the information on setting the MAC address; it should never be necessary.

Connecting the hardware is easy. First, connect your serial cable to the 9 pin connector. Set your terminal emulator to 57600 bits per second, no parity, 1 stop bit, and no flow control. Chances are that you will use *Hyperterminal* that has come with every version of *Windows* since *Windows 98*. You should be aware that the version that is shipped with *XP* (and probably *Vista*) is broken. The only text that is correct is that in the lower boxed area. Scrolling up the screen will almost certainly result in garbage being displayed. If you want more than 25 lines of text, you will want either the full version of *Hyperterminal* or perhaps *Procomm*. You can get by with 25 lines, but it is less convenient than 50 to 60 lines of output. The second step is to connect the Stamp to the router using a normal CAT-5 cable.

Start your terminal emulator and then plug in the power connector to the Stamp. If everything is set up correctly, you will see the output of the boot loader scrolling very quickly across the screen. If you don't see output or cannot see a response to typing <Enter>, you need to troubleshoot normal serial computer communications. In my case, I grabbed a serial cable with an intermittent broken wire and <Enter> did nothing. In the *Linux* world, this terminal is called the console. It comes from the days when computers took entire buildings and the main interface for the guy who started up the computer was a BAUDOT teletype terminal with a paper tape reader (ASCII if they were really lucky) connected to the boot hardware.

Target Software Setup

The next step is to make sure you can talk to the network of the board. Figure 1 shows the terminal screen after the boot is finished. At this point, the network hardware is initialized, but the network is not actually running. Type the command "**ifconfig eth0 192.168.1.99 netmask 255.255.255.0 up"**.

Figure 2 shows the output from running the command with the cable unplugged and then repeating the command. The third command is **ifconfig** without arguments. This shows you how the network is configured.

Here is a good place to talk about the

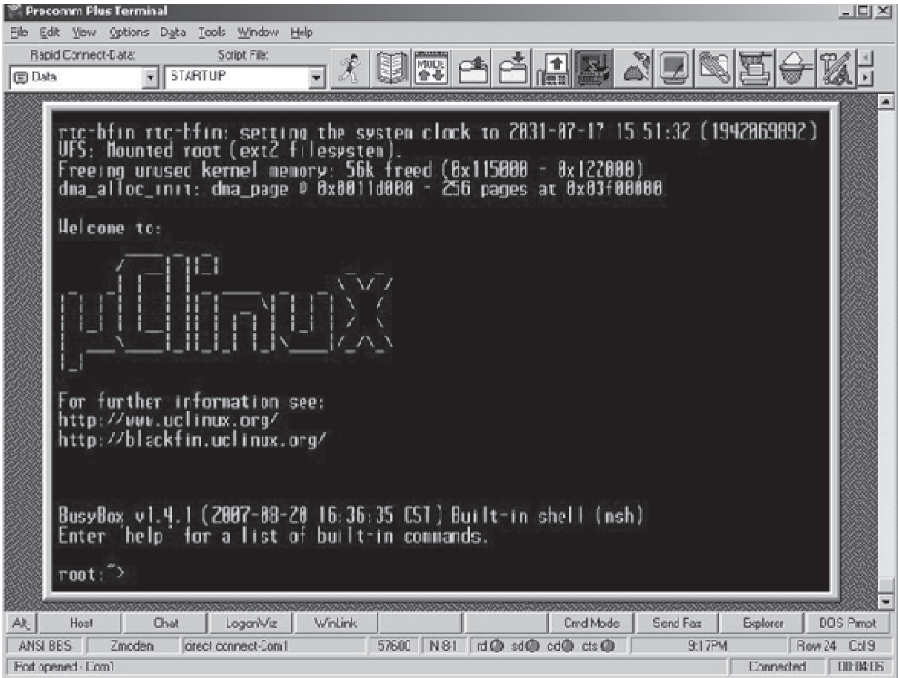

Figure 1 — Stamp Console output immediately after boot. The system is ready for commands to be typed. The prompt "root:~>" shows that you were automatically logged into *Linux* as the user name "root".

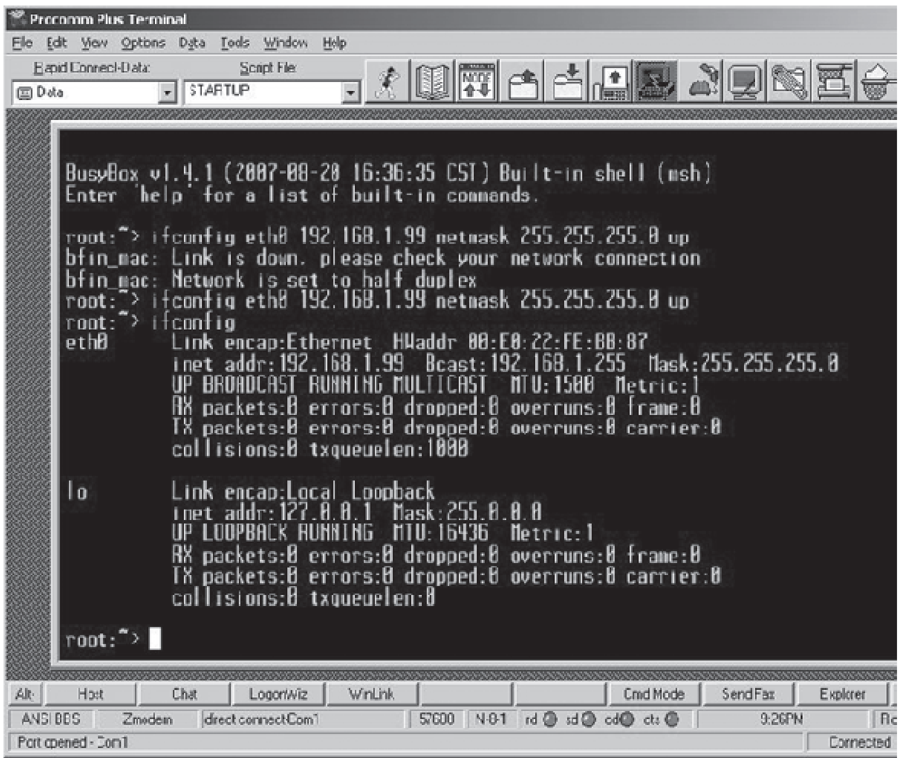

Figure 2 — Stamp console output from three consecutive ifconfig commands. The first command failed because the network cable came unplugged. The second command repeats the first and configures the network with a static IP address. The third ipconfig command without arguments displays the parameters that the system uses for the network.

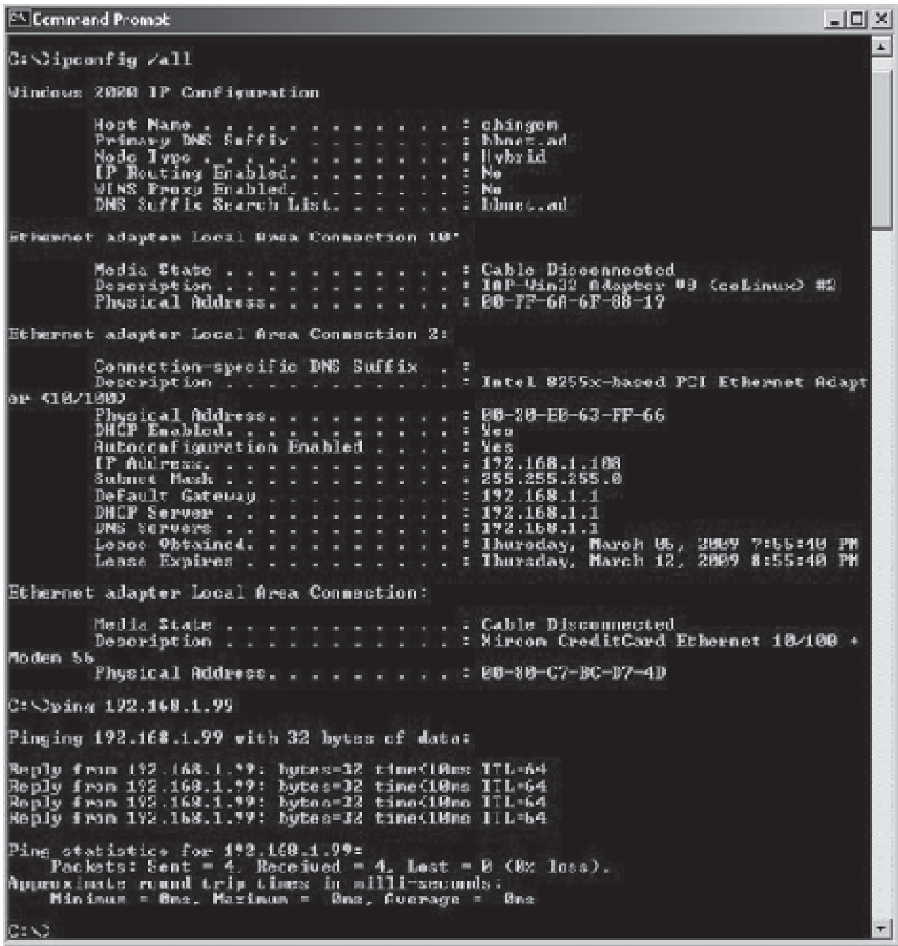

Figure 3 — Representative *Windows* command prompt output for ipconfig command and ping command.

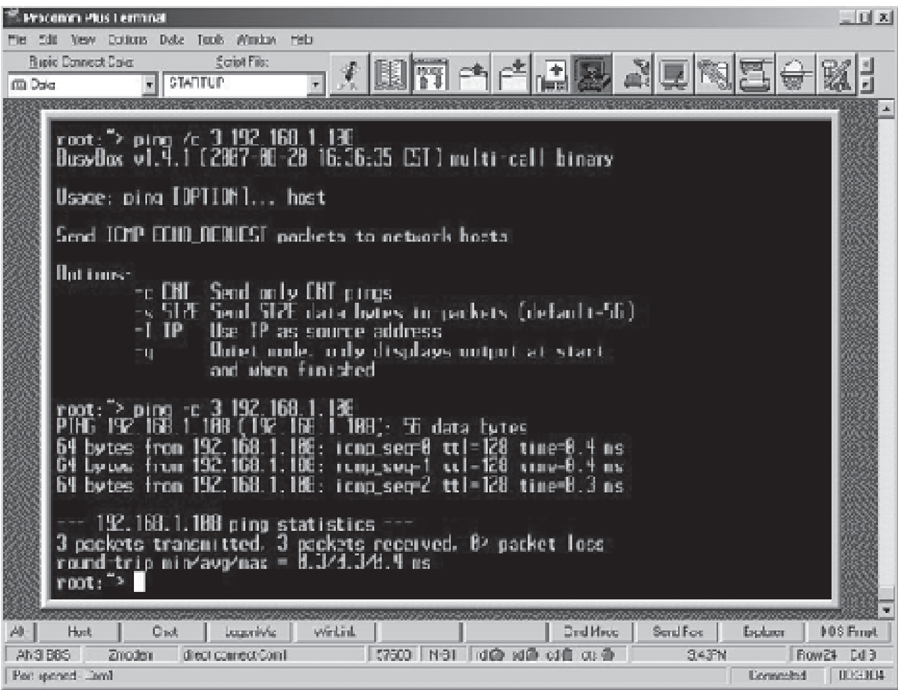

Figure 4 — Representative Stamp shell command output for the ping command. Note that the command does not understand the *Windows* syntax "/c" in the first attempt to run the command.

shell. In *Windows*, you can use a window called a Command Prompt that is just a DOS command system running on *Windows*. Both the *Windows* command prompt and the *Linux* shell allow you to type in commands to the operating system and run programs. It is really tedious to type the same commands over and over, so both systems use some keys on the keyboard to recall old commands so they can be run again or modified and then run again. The up arrow and down arrow keys cycle back and forth between commands. The left and right arrow keys move back and forth along a command so you can edit it.

Open up a *Windows* Command Prompt from the Start menu using **Start, Programs, Accessories**. Now you can type **ipconfig /all**. Ipconfig is the *Windows* equivalent of ifconfig. It gives similar information. The important information is buried in the middle of the screen. The IP address for my computer is 192.168.1.108. The tool you use to tell if the very basic network is working is "ping" (named after the SONAR sound). From the Stamp we type "**ping –c 3 192.168.1.108**" to send three data packets across the network from the Stamp to the laptop. Then in the DOS command prompt we type "**ping 192.168.1.99**". The *Windows* ping does four packets and stops. *Linux* ping will send packets forever unless you specify three packets with "**-c 3**". Since both computers give success for the ping commands, we know that they can communicate across the network.

Notice that the command parameter to ipconfig was **"/all**". The equivalent with ifconfig is "**-a**". Linux uses a "**-**" to indicate a parameter where DOS uses a "**/**". When we need to access directories, we will run across the other main difference in the two systems. Linux uses a path such as "**/bin/usr/bin**". The equivalent in DOS is "**\bin\usr\bin**". You will just have to train your brain to handle which system you are working on.

Development Software Setup

Do not read any of the tool chain information on the CD. It is totally useless for our purposes.

You will go to **blackfin.uclinux.org/gf/project/toolchain/frs/?action=FrsReleaseView&release_id=392** to get the *Windows* hosted toolchain. The *Windows* file (blackfin-toolchain-win32-2008R1.5.exe) is at the very bottom of the Web page. This is the set of software that allows you to write a *C* program and run it on the Stamp. We will use the *Windows* computer as our cross development system for our DSP programs on the Stamp. Download the exe file and run it. It will do all of the set up for you automatically. Select a convenient directory where you will know how to find it. I recommend against using "c:\program files". I used "c:\analog devices".

Once the install is complete, you will have the compiler and linker necessary to build a program to run on the Stamp under the *Linux* operating system. Chances are good that you have not used the "make" program before. This is an automated way to build programs and all real software developers are familiar with make. We won't need it for now, but it is there. We will use make when we have non-trivial programs to build.

The next step is to create a "Hello World" program using your favorite software editor. I use *Codewright* because it highlights features in C code and does some automatic formatting. *Visual Studio* is also a good choice for the same reasons and it is free. All the program does is print "Hello world." on the console terminal for the Stamp. Here is the C code for the program:

```
#include <stdio.h>
int main(void)
{
        printf("Hello world\n\r");
        return(0);
}
```

Note that your program must have a blank line after the final "}" or the compiler will complain. Store the code above in a file named hello.c. I stored my file as "C:\Analog Devices\GNU Toolchain\2008R1.5\examples\hello.c". Next, open your command prompt and change to "C:\Analog Devices\GNU Toolchain\2008R1.5\examples\" directory. You run the compiler with the following command: C:\Analog Devices\GNU Toolchain\2008R1.5\examples>bfin-linux-uclibc-gcc hello.c –o hello

The first part of the command line is the name of the compiler. There are actually several compilers as part of the package. The compiler you use depends on whether you are making a user program or the entire *Linux* system. The name "linux-uclibc" tells the main part of the Gnu compiler (gcc) that it will use a library package called "uclibc" for *Linux*. That library is where the function printf is implemented. The argument "–o hello" tells the compiler to make a file that can run on a *Linux* system called "hello".

The next step will copy the hello program from your *Windows* computer to the Stamp. Type the command **ftp 192.168.1.99** in the command prompt window. The *Windows* computer connects to the Stamp using the "file transport protocol." You will see a console banner from *uClinux* running on the Stamp. You will need to login. The user name is "root" and the password is "*uClinux*". Be sure to type the password correctly. Upper and lower case counts!

You now have an authenticated connection between the computers. Next type "binary<enter>" to change the file transfer mode from ASCII to binary. Now type "put hello<enter>". This transfers the file to the root directory of the Stamp. Type "quit" to close the ftp connection.

Open your terminal emulator window again. Type "**ls –l**" (that is lower "L" and not a "1") to show a directory of the files in the root directory of the system. "ls" is the equivalent of "dir" in DOS, and "–l" tells it to print a long format with all the properties of each file. Depending on your terminal capabilities, you will see a list of color coded files. One of those is the file hello. You cannot run the file hello yet. *Linux* knows the file is there, but it does not have permission to be used as a program. The first thing you see for hello is "-rw-r-----". These letters tell Linux that the file can be read and written, but it cannot be executed. The next step is to change hello from just a file to a program. Type "**chmod 777 hello**" and then "**ls –l**". The chmod command (change mode) changes the properties of the file so that it can run as a program. The permissions are now "-rwxrwxrwx". Also notice that the color for the hello ls listing has changed. Run the program by typing "**./hello**" and you will see "Hello world" printed by itself on a line. I have never figured out why you have to type "**./hello**" rather than just "**hello**", but it is a weird little idiosyncrasy of *Linux* and *Unix*.

DSP Software Operation

So far we have used the Stamp as just a computer. We really want to use it as a signal processor. *Linux* gives us the best of both worlds. It lets us write normal computer programs in the *C* language to do our work. *Linux* has a fairly easy interface to allow us direct control of the hardware on our boards. The version of *uClinux* on our Stamp boards has a diverse array of device drivers that supply the interface between the hardware and our computer programs. You can get an idea of the types of devices by looking in the /dev directory using the command "**ls /dev**". The ones we will use are the ones that mention SPI, PPI, and SPORT. Even the DSP engine is available through a device driver interface.

The AD7476 ADC connects to the Stamp using the SPI interface. This ADC is designed for "on demand" conversion. The ADC starts a conversion each time the chip select pin goes from high to low. When we write an application to use the AD7476, we will need to use a timer to start each conversion to keep the sampling period constant.

The parallel port interface is useful for parallel interface ADC's and DAC's. Most of the high speed devices use synchronous serial interfaces on the SPORT ports. These two interfaces have DMA and interrupt support, so they are useful for designs that use a constant clock for acquisition.

Next Issue

The plan for the next issue is to put together the AD7476 board, the Stamp, and a parallel 8 bit DAC to create a very simple AM broadcast receiver. Think of it as the DSP equivalent to a crystal set.

Upcoming Conferences

43rd Annual Central States VHF Society Conference

July 23-25, 2009
Elk Grove Village, IL

The Central States VHF Society is soliciting papers, presentations, and Poster / table-top displays for the 43rd Annual CSVHFS Conference to be held in Elk Grove Village, IL, near Chicago on July 23-25, 2009. Papers, presentations, and posters on all aspects of weak-signal VHF and above Amateur Radio are requested. You do not need to attend the conference, nor present your paper to have it published in the *Conference Proceedings*. Posters will be displayed during the conference.

Topics for papers and presentations may include:

Antennas, including modeling/design, arrays, and control.

Construction of equipment, such as transmitters, receivers, and transverters.

Digital Modes, such as WSJT, JT65 and other modes.

RF power amps, including single and multi-band vacuum tube and solid-state.

Propagation, including ducting, sporadic E, tropospheric and meteor scatter.

Pre-amplifiers (low noise).

Software-defined Radio (SDR).

Regulatory topics.

EME.

Digital Signal Processing (DSP).

Test Equipment including Homebrew, Using, and making measurements.

Operating, including contesting, roving, and DXpeditions

Non-weak signal topics, such as FM, Repeaters, packet radio, and so on, are generally not considered acceptable. There are always exceptions, however. Please contact K9XA, as listed below, if you have any questions about the suitability of a topic.

Submission Deadlines:

For the Proceedings: Monday, 1 June 2009

For Presentations delivered at the conference: Monday, 29 June 2009

For notifying us you will have a Poster to display at the conference: Monday, 29 June 2009. Bring your poster with you on July 23/24.

Contact: Kermit Carlson, W9XA, 1150 McKee St, Batavia IL 60510; **w9xa@yahoo.com**

Please see the Web site at **www.csvhfs.org** for more information.

The 28th Annual ARRL and TAPR Digital Communications Conference

September 25-27, 2009
Chicago, Illinois

Mark your calendar and start making plans to attend the premier technical conference of the year, the 28th Annual ARRL and TAPR Digital Communications Conference to be held September 25-27, 2009, in Chicago, Illinois. The conference location is the Holiday Inn Elk Grove Village Hotel, Elk Grove Village, IL. This is the same location as last year's DCC.

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

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

This is a three-Day Conference (Friday, Saturday, Sunday). Technical and introductory sessions will be presented all day Friday and Saturday.

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

The ever-popular Sunday Seminar focuses on a topic and provides an in-depth four-hour presentation by an expert in the field. Check the TAPR Web site for more information: **www.tapr.org**.

Call for Papers

Technical papers are solicited for presentation and publication in the *Digital Communications Conference Proceedings*. Annual conference proceedings are published by the ARRL. Presentation at the conference is not required for publication. Submission of papers are due by 31 July 2009 and should be submitted to:

Maty Weinberg, ARRL, 225 Main Street, Newington, CT 06111, or via the Internet to **maty@arrl.org**.

MICROWAVE UPDATE

October 23-25, 2009
Dallas, TX

Call for Papers

Microwave Update 2009 will be held on Friday, October 23 through Saturday, October 24 in Dallas, Texas. Technical papers are currently being solicited for publication in the ARRL *Microwave Update 2009 Conference Proceedings*. You do not need to attend the conference nor present your paper to have it published.

Strong preference will be given to original work and to those papers that are written and formatted specifically for publication rather than as a visual presentation aid. As this is a microwave conference papers must be on topics for frequencies above 900 MHz. Examples of such topics include microwave theory, construction, communication, deployment, propagation, antennas, activity, transmitters, receivers, components, amplifiers, communication modes, LASER, software design tools, and practical experiences.

Authors retain the basic copyright, and by submission, consent to publication in the *Proceedings* and possible publication of the *Proceedings* in CD/DVD format. If you are interested in submitting a paper please contact Kent Britain, WA5VJB at **wa5vjb@flash.net** for additional information. The deadline for papers is Monday, August 31, 2009.

Please refer to the Proceedings Style Guideline on the conference Web site at **www.microwaveupdate.org/index.php**.

If you are interested in presenting at Microwave Update please contact Al Ward, W5LUA, at **w5lua@sbcglobal.net**.

Next Issue in *QEX*

Bob Zavrel, W7SX, uses a theoretical analysis in discussing “Maximizing Radiation Resistance in Vertical Antennas.” In his article, W7SX attempts to answer the question, “What is the optimum design of a vertical antenna to maximize radiation efficiency?” The emphasis is placed on maximizing radiation efficiency for a given height restraint rather than the more often published concern of ground losses. Both ground losses and radiation resistance contribute directly to vertical antenna efficiency, so maximizing radiation resistance is equally beneficial as minimizing ground losses.