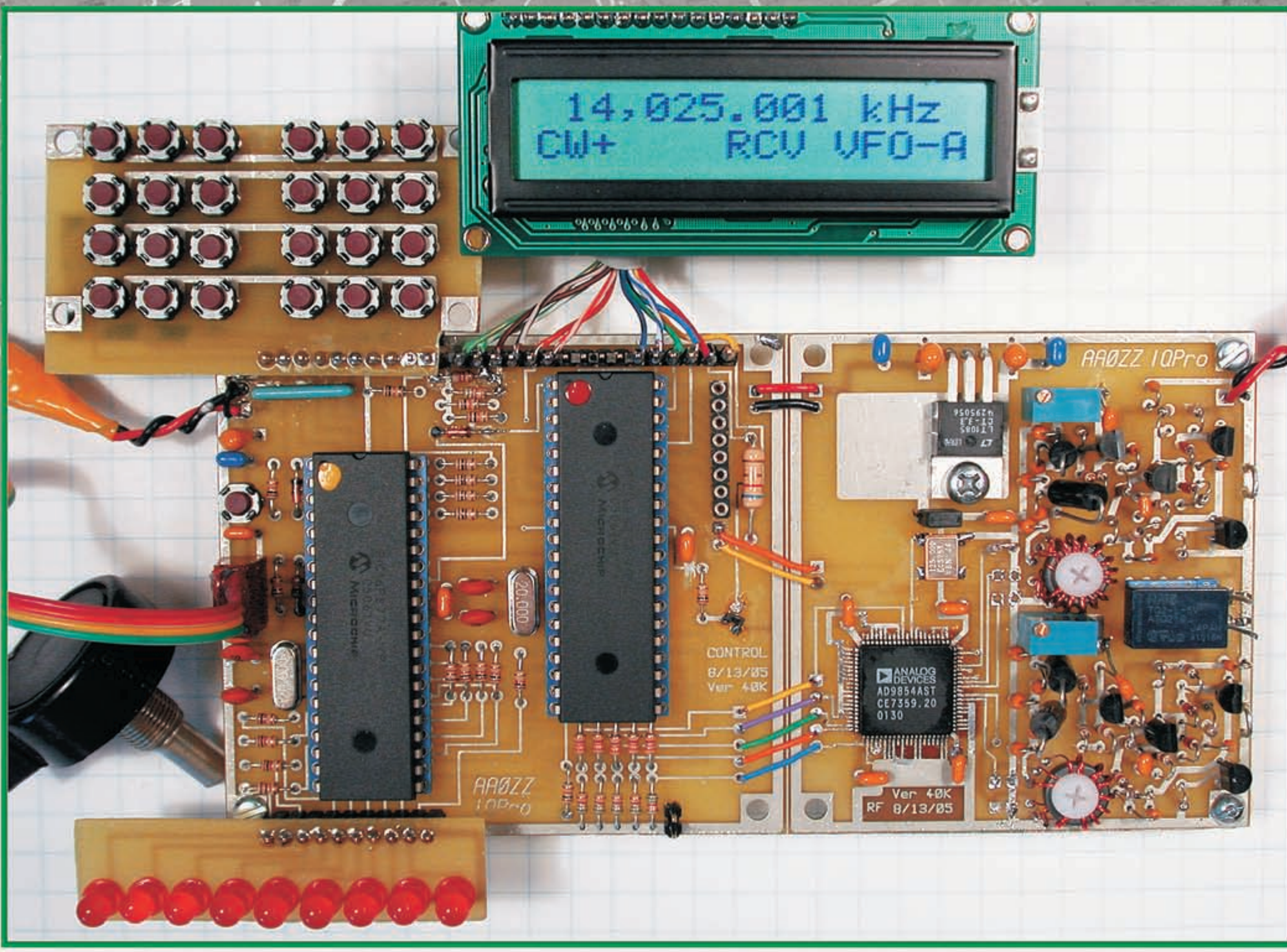
QEX

May/June 2006

A Forum for Communications Experimenters

Issue No. 236



AAØZZ describes a high-performance, quadrature, direct digital synthesized VFO that uses dual peripheral interface controllers (PICs) and a two-line LCD.

ARRL The national association for AMATEUR RADIO
225 Main Street
Newington, CT USA 06111-1494

QEX

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 236

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;

Elsewhere by Surface Mail (4-8 week delivery):
ARRL member $31, nonmember $43;

Canada by Airmail: ARRL member $40, nonmember $52;

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About the Cover

Craig Johnson, AAØZZ, describes a high-performance direct digital synthesized VFO. His project uses dual PICs and a two-line LCD. Dual VFO operation supports split-frequency operation.

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Empirical Outlook

Doug Smith, KF6DX
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In Search of New Receiver Performance Paradigms

Many well-established methods of modeling and measuring receiver performance fail to keep pace with the achievements of modern designs. At least one of them is in danger of immediate extinction.

We say that the primary goal of receiver testing is to discover a unit's limitations. Underlying that is the necessity to relate those limitations to real conditions the unit is likely to encounter. Thus we strive to measure how well the goals of the designer and the user converge. Does the unit do what the designer intended? Was it designed to the realistic expectations of the user?

Third-order intercept point (IP3) has been used for ages as a figure-of-merit for receivers. That figure applies to any circuit or system that exhibits cube-law nonlinearity and remains well-defined. Noise-floor power in a receiver is used to compute IP3. Since it is bandwidth-dependent, that makes IP3 bandwidth-dependent. That's undesirable and we have better ways of stating a receiver's noise performance.

IP3 is not a measurement, but an extrapolation based on the equation:

$$\text{IP3} = 1.5\ (\text{IMD3DR}) + \text{NOISE FLOOR}$$

where IMD3DR is the third-order intermodulation distortion (IMD) dynamic range in dB, and NOISE FLOOR is the noise-floor power in dBm. In turn, IMD3DR is the ratio of the equal power of each of two off-channel tones input to a receiver to the IMD3 product they produce, whose power is equal to the noise-floor power. In a receiver obeying perfect cube-law nonlinearity, it doesn't matter at what level you measure the IMD, since you can still relate that to the noise floor.

What we're finding out is that much of the data published in *QST* and elsewhere don't obey the defining equation. Take the product-review data from the Ten-Tec Pegasus, for example.[1] 14-MHz noise-floor power in a 500-Hz bandwidth is listed as –132 dBm. IMD3DR is listed as 77 dB. The equation yields IP3 = –16.5 dBm. Yet IP3 is listed as +7 dBm in the review. Why the 23-dB discrepancy?

The ARRL Lab's statement on the subject says that receivers don't necessarily follow perfect cube laws, making "…the real intercept point of a receiver subject to the judgment of the person looking at the real response curves…" It's stated "…there really is no true number…."[2] Judgement allegedly comes into play when deciding at what level to measure IMD3. But then we have to ask, "If things aren't cube-law, how can we justify a cube-law extrapolation of the data?" We add that were the "true" IP3 of a receiver measured, the input power of the test signal would fry the front end and maybe a few other things.

We must ask ourselves some very tough questions: Are we ready to declare IP3 meaningless? At a single stroke, do we refute the *Handbook* and all other references to classic third-order behavior? If we change our procedures, how do we compare newly obtained data with old? How do we correct the old data?

The *Handbook* has at least one good avenue to improvement.[3] It describes how to relate dynamic range using a 1-Hz bandwidth in a process called "normalization." That at least removes the bandwidth dependence of the data and suggests a statement of receiver noise in bandwidth-independent terms (noise figure). The next logical step is to "bite the bullet" and commit to measuring IMD3 at more than one point for all units.

In most receivers, automatic gain control (AGC) comes into play over part of that range. Front-end AGCs can fool the heck out of current procedures, since IMD isn't measured under the same conditions as is the noise-floor power. Additionally, direct-conversion receivers with analog-to-digital converters at or near the antenna don't exhibit IMD3 in the same way as conventional units.

All those things and more are forcing changes in our test models. How and when we make changes and correct previous data will be a measure of our collective character and serve as part of our legacy.

[1] Product Review, *QST*, Feb 2000.

[2] E. Hare, W1RFI, "What is the 'Real' Intercept Point," in D Smith, KF6DX, "Improved Dynamic Range Testing," *QEX*, Jul/Aug 2002, **www.arrl.org/tis/info/pdf/020708qex046.pdf**.

[3] W. Sabin, W0IYH, "Receivers and Transmitters," in *The ARRL Handbook*, 2006, R. D. Straw, N6BV, ed, p 14.8.

CW Shaping in DSP Software

Getting into shape, via DSP filters.

Alex Shovkoplyas, VE3NEA

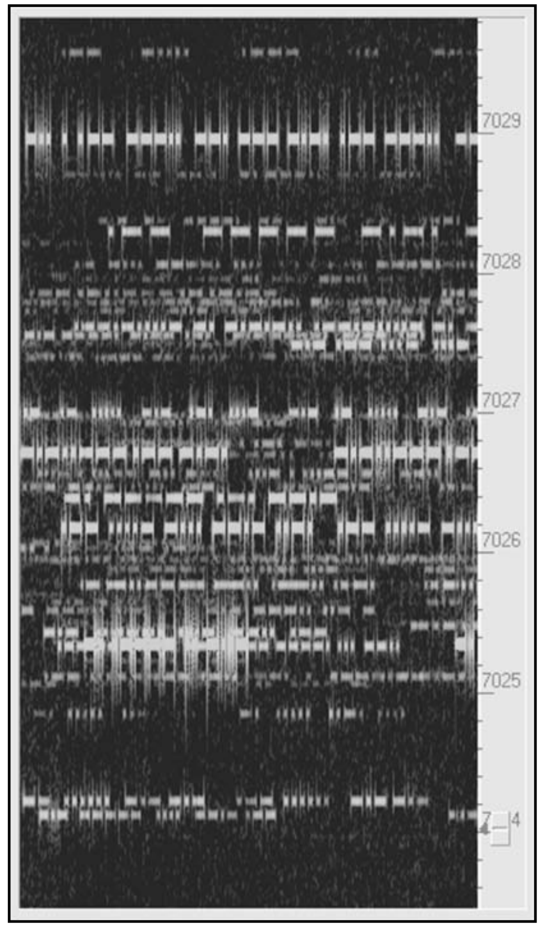

Figure 1 — K7C pileup on 40 m.

CW portions of amateur HF bands are very crowded at times. It is not uncommon to hear several stations within 100 Hz of each other, especially in the big pileups. Figure 1 shows a waterfall display of the K7C pileup on 40 m. The vertical axis is frequency, and the horizontal axis is time. You can see many stations calling the DX; the density of the pileup reaches 10 stations per kilohertz. Under such conditions, it is very important that each signal occupy as little bandwidth as possible.

Traditionally, the bandwidth of a CW signal is reduced by shaping its dots and dashes. In the classical, all-hardware transceivers, CW is shaped with RLC elements, and the choice of shapes that can be achieved is very limited. Now that firmware- and software-defined radios are becoming popular, the developer has much greater control over the CW shape: virtually any shape can be implemented with DSP methods. Among the popular shapes are sine, raised cosine, Gaussian, and the integral of SIN^4.[1-4]

[1]Notes appear on page 7.

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Usually the developer selects a smooth shape, in a hope that it will result in a narrow bandwidth. There is a better approach, though. CW shaping is in fact a filtering problem. When you replace rectangular edges of Morse elements, this is equivalent to low-pass-filtering the envelope, and there is a one-to-one relationship between the shape applied and the equivalent filter. Thus, instead of selecting a shape, we can select a prototype filter with known characteristics, and use its step response as a waveform to shape CW.

We will demonstrate the method by designing a keying shape that has characteristics similar to those commonly used in amateur transceivers. We will design a shape with the rise time of 5 ms, a value typically used in commercial radios, but with a higher level of spurious emission rejection.

The Blackman-Harris FIR filter will be used in this exercise because it has good characteristics both in the time and frequency domain. This filter has no overshooting, its stopband rejection is over 90 dB, it has a pretty good shape factor and no ripple in the passband. Also, the values of the Blackman-Harris kernel can be calculated with just a few lines of Delphi code:

```
function BlackmanHarrisKernel(x: Single): Single;
const
  a0 = 0.35875; a1 = 0.48829; a2 = 0.14128; a3 = 0.01168;
begin
  Result := a0 - a1*Cos(2*Pi*x) + a2*Cos(4*Pi*x) - a3*Cos(6*Pi*x);
end;
```

Code snippet 1 — The function that calculates a single point of the prototype filter's impulse response.

The step response of the filter is an integral of its kernel. We calculate it as follows:

```
type
 TSingleArray = array of Single;


function BlackmanHarrisStepResponse(Len: integer): TSingleArray;
var
 i: integer;
 Scale: Single;
begin
 SetLength(Result, Len);
 //generate kernel
 for i:=0 to High(Result) do Result[i] := BlackmanHarrisKernel(i/Len);
 //integrate
 for i:=1 to High(Result) do Result[i] := Result[i-1] + Result[i];
 //normalize
 Scale := 1 / Result[High(Result)];
 for i:=0 to High(Result) do Result[i] := Result[i] * Scale;
end;
```

Code snippet 2 — The function that calculates the step response of the prototype filter.

The Len parameter is the kernel length, in samples. Given the desired rise time, the required kernel length can be calculated as 2.7 × RiseTime × SamplingRate.

Now that we have a filter, we can apply it to the CW envelope and see how it works. Normally the filter is applied by convolving its kernel with the input signal, but in the case of CW shaping we do not need to calculate the convolution. A sequence of rectangular pulses with unit amplitude can be viewed as a sum of positive and negative step functions, one function per edge. Since our filter is linear, we can filter each step function separately and then add up the results. Moreover, we already have a filtered version of the step function; it is the step response of the filter that we have just calculated. In practice, assuming that the dot duration is greater than the kernel length, we do not even have to do additions; we just copy the step response to the output buffer at the locations where the pulse edges are:

```
var
 Rise, Fall: TSingleArray;


function ShapeCW(Data: TSingleArray): TSingleArray;
var
 i: integer;
begin
 Result := Copy(Data);
 SetLength(Result, Length(Result) + High(Rise));


 for i:=High(Data) downto 1 do
  if (Data[i-1] = 0) and (Data[i] = 1) then
   Move(Rise[0], Result[i], Length(Rise) * SizeOf(Single))


  else if (Data[i-1] = 1) and (Data[i] = 0) then
   Move(Fall[0], Result[i], Length(Fall) * SizeOf(Single));
end;
```

Code snippet 3 — The function that applies shape to the CW envelope.

The code above assumes that we have pre-calculated the step response of the filter with the BlackmanHarrisStepResponse function and stored it in the Rise array. The Fall array is a copy of Rise, mirrored left-to-right.

Note that filtering increases the data length by the number of points in the kernel minus 1. Use the overlap-add method to process multiple blocks of data.

The suggested method of CW shaping was tested on a sequence of dots sent at 40 WPM (30 ms dot length). The 5 ms rise time was selected.

The dot waveform after shaping, and the spectrum of the shaped signal are shown on the chart below (Figure 2). The grid steps on the spectrum display are 100 Hz horizontally and 10 dB vertically.

Figures 3 and 4 show the spectrums of the signals keyed at 20 WPM and 80 WPM respectively, shaped with the proposed method. The fine structure of the spectrum depends on the keying speed, while the spectrum envelope is a function of the shaping filter.

The waveforms and spectrums of several other CW shapes are shown on Figures 5-8 for comparison. All shaping filters are tuned to produce a 5-ms rise time, and the keying speed is 40 WPM.

The chart on Figure 9 shows all waveforms and spectra combined. All spectra have virtually the same bandwidth at the levels down to –40 dB, but beyond that point they differ significantly. At –100 dB, all shapes except Blackman-Harris have a very wide bandwidth, well over 1 kHz. The bandwidth of the Blackman-Harris shape at this level is less than 300 Hz. It is interesting to note that visually all tested shapes look almost the same, the differences are so small that they can be barely seen on the chart; but spurious emissions that we want to reject are also of low levels, between –40 to –100 dB, and small differences in the shape play a significant role in achieving a good rejection factor. As you can see from Table 1, these tiny differences can improve rejection by 30-50 dB!

Table 1 shows the level of spurious emissions at a 300 Hz offset for the keying speed of 40 WPM and the rise time of 5 ms for each of the discussed keying shapes.

Blackman-Harris has much better characteristics than all other shapes in the table.

Design Parameters

Now that we know how to design a keying shape that meets our specifications, we will see what parameters we should set as a design goal.

A good shape must meet several conflicting requirements. The rise time (and the corresponding bandwidth) is only one of those. In fact, it is the easiest parameter to achieve.

Table 1
Spurious Emission Level at 300 Hz Offset

Shape	*Spurious Emissions at 300 Hz Offset*
Blackman-Harris	< –100 dB
Rectangular	–29 dB
Sine	–51 dB
Raised Cosine	–70 dB
Truncated Gaussian	–83 dB

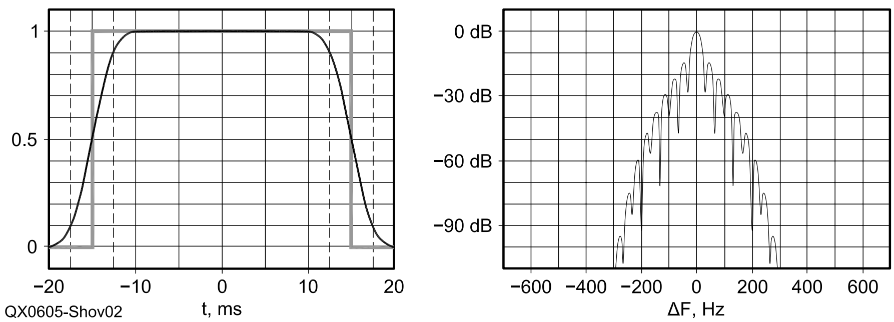


Figure 2 — Blackman-Harris CW shaping.

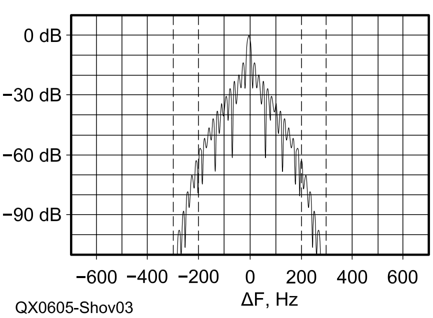


Figure 3 — Blackman-Harris shaping, 20 WPM

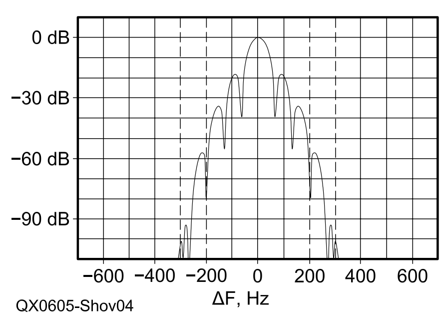


Figure 4 — Blackman-Harris shaping, 80 WPM

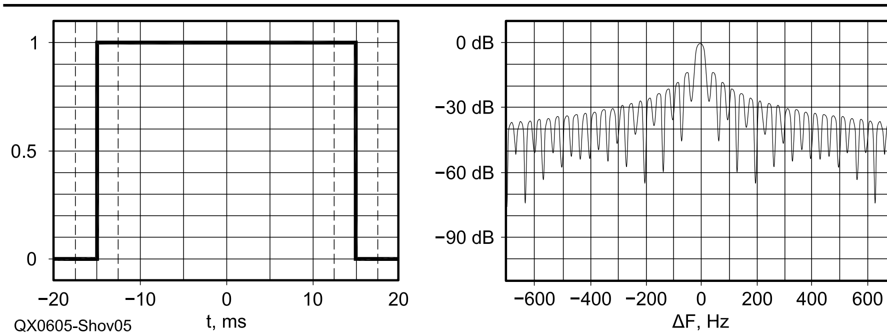


Figure 5 — No shaping.

Basically, we want the following from the signal shape:

1) the signal must be easy to copy, the distortion of the original shape should be minimal;

2) the power density of spurious emissions (key clicks) away from the operating frequency must be as low possible; and

3) the length of the keying shape (which is often many times longer than the rise time)

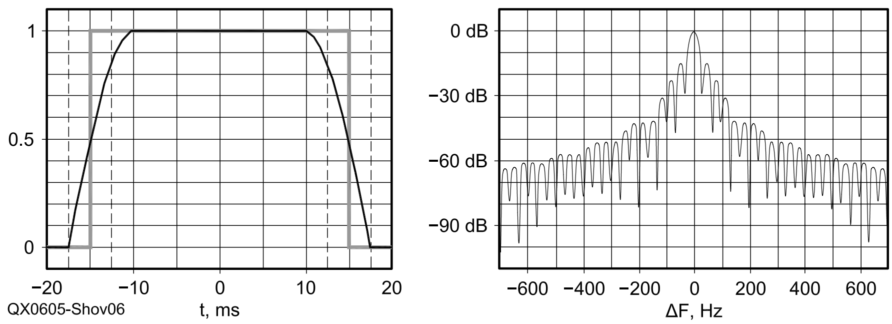


Figure 6 — Sine shaping.

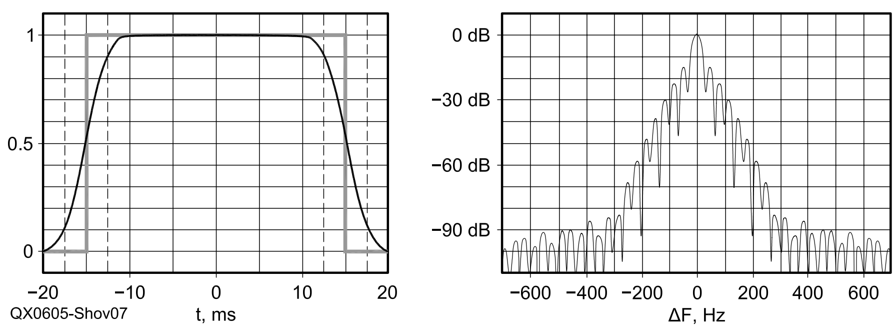


Figure 7 — Raised Cosine shaping.

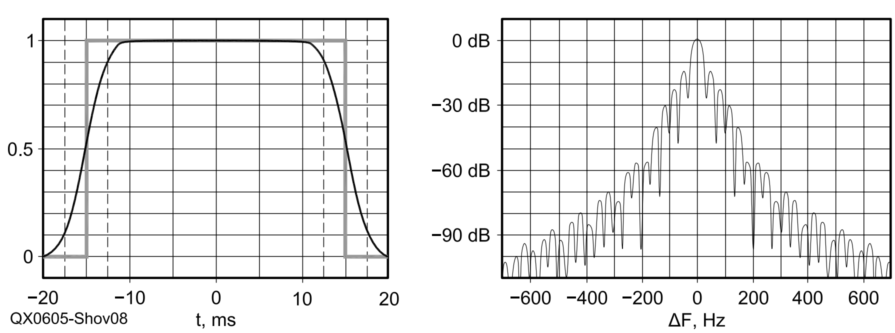


Figure 8 — Gaussian shaping, kernel truncated at 3 × sigma.

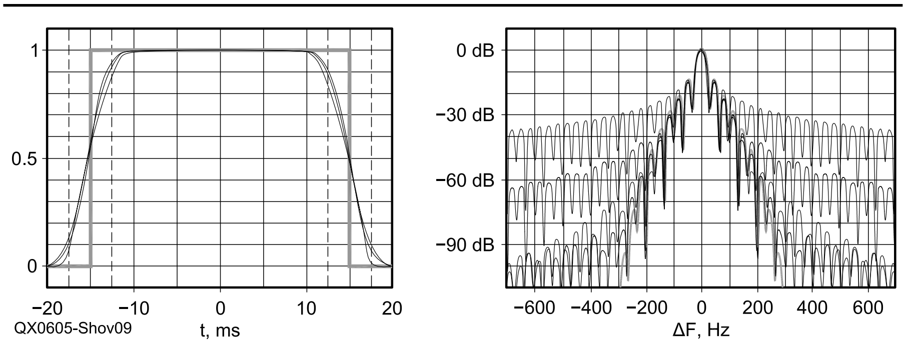


Figure 9 — All charts combined. The Blackman-Harris shaping chart is the bold line.

must be small since it affects the signal delay introduced by the shaper.

It follows from the above that the choice of the CW shape is not just a single parameter minimization problem. The developer must decide what level of spurious emissions he wants to achieve, at what frequency offset that level should be measured, and to what degree (1) and (3) can be relaxed to meet (2).

One possible strategy is to design the filter and resulting keying shape to achieve (1) and to use the remaining degrees of freedom to optimize if for (2) and (3). The design will then be reduced to choosing the three main parameters of the prototype filter: bandwidth, shape factor, and stopband rejection.

Obviously, we want the stopband rejection to be as high as possible, but what about the other two parameters? How do they affect the operator's ability to copy weak signals? To answer this question, I wrote a program called *CwBwTest*, and performed a few tests.

The program can be downloaded from **www.dxatlas.com/CwShaping/CwBwTest.zip**. It has three tabbed pages. On the first page, there are controls that play CW audio with specified keying speed, signal-to-noise ratio, bandwidth and shape factor.

Page 2 presents a weak signal reception test. The program plays random call signs in the noise at 20 WPM, prompts you to copy those call signs, and plots the copy accuracy on a chart as a function of the SNR for different shaping parameters. Filtering is applied to the CW signal only, the noise bandwidth is 500 Hz in all cases. This is equivalent to shaping the signal in the transmitter and receiving it with a typical CW receiver.

Page 3 in the program shows the envelope oscillogram of the signal being played. The results of the tests are shown on Figure 10. The tests were performed by three operators independently; 1500 call signs were copied in total. The results were within 0.5 dB of each other, and the curves had the same shape. The chart presents the combined results of all tests.

As can be seen from the chart, the optimal bandwidth of the 20 WPM signal is 30 Hz. The copy accuracy does not increase as the bandwidth increases beyond this point, but the accuracy drops at lower bandwidths. At the optimal bandwidth, the exact shape of the envelope does not affect the accuracy.

Based on the test results, we can conclude that the keying shape at 20 WPM should be selected to produce the 30 Hz bandwidth at –6 dB, or 1.5 × WPM. The shape factor of 2.7 is optimal — not because of the copy

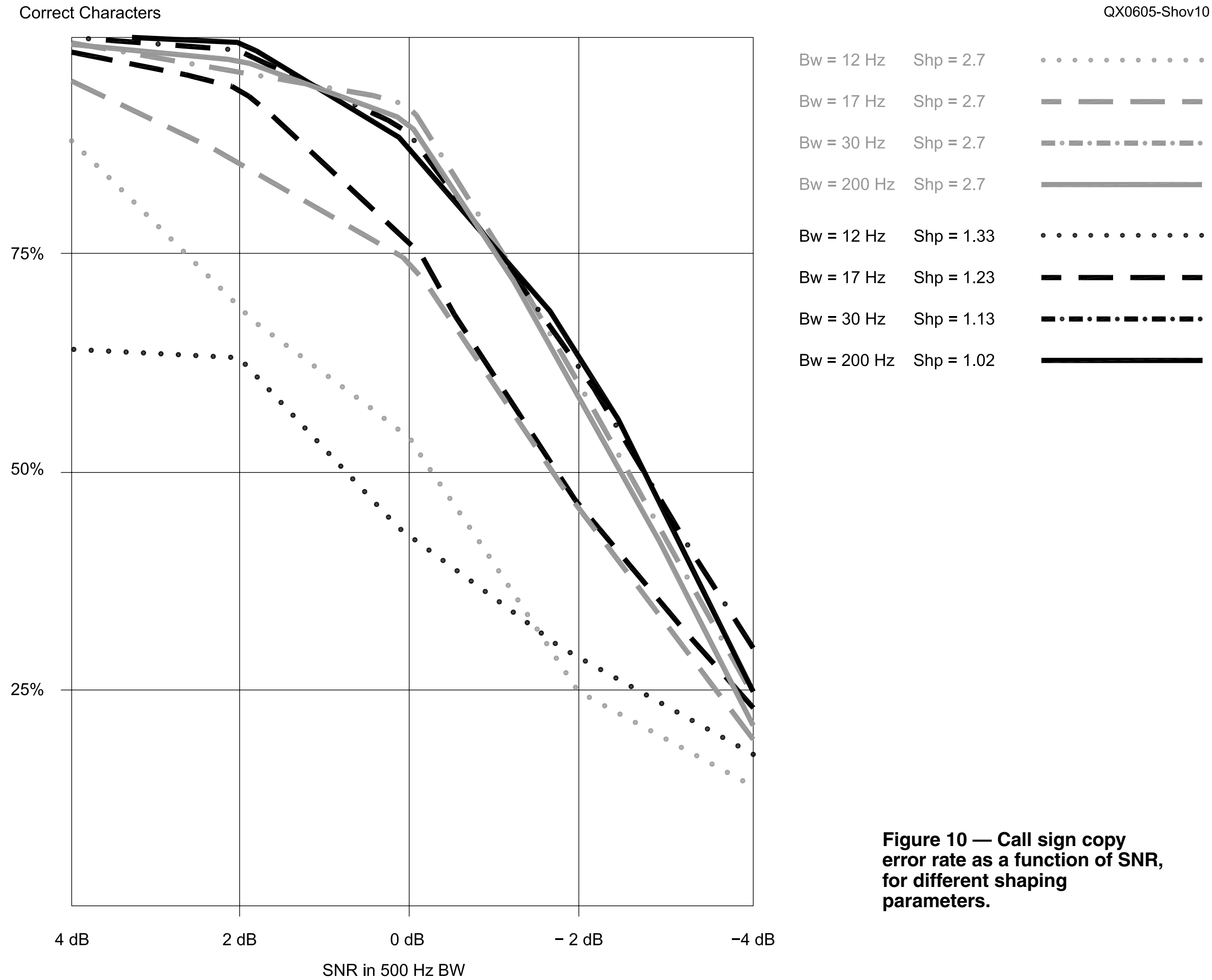


Figure 10 — Call sign copy error rate as a function of SNR, for different shaping parameters.

accuracy, but because it ensures no Gibbs effect that makes CW sound unpleasantly. With these parameters and the Blackman-Harris keying shape, the total bandwidth at the –100 dB level is about 80 Hz.

Notes

[1] *On the Occupied Bandwidth of CW Emissions*. Douglas T. Smith Editorial Services, 2004. Online publication: **www.doug-smith.net/cwbandwidth1.htm**.

[2] PowerSDR 1.4.4 Source Code: Flex-Radio Downloads. Online download: **www.flex-radio.com/download_files/**.

[3] *Spectral Analysis of a CW keying pulse.* Kevin Schmidt, W9CF. Online publication: **fermi.la.asu.edu/w9cf/articles/click/index.html**.

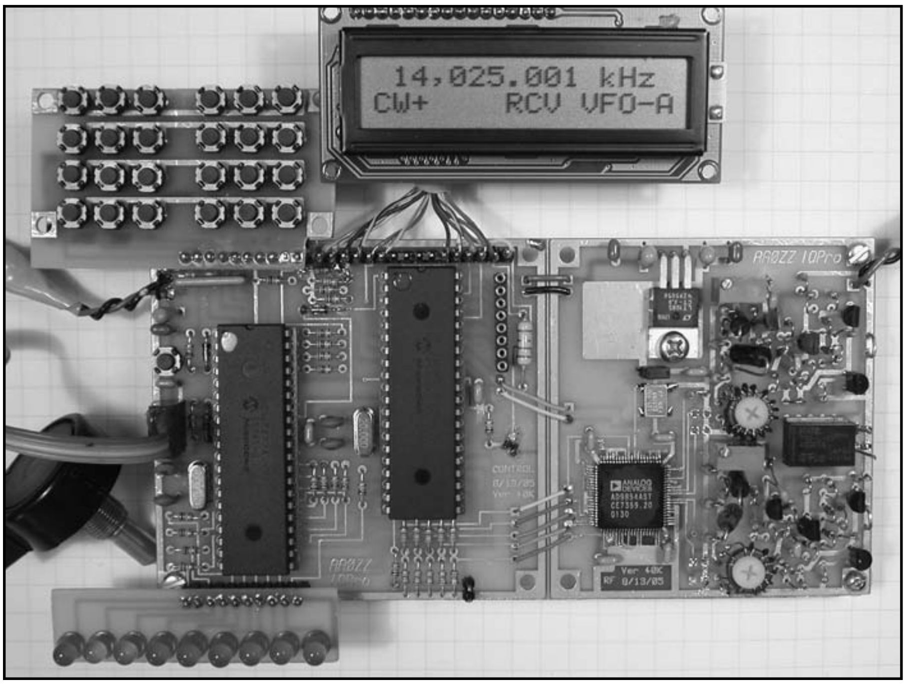
[4] *The T03DSP High Performance Transceiver with DSP IF processing.* Oleg Skydan, UR3IQO. Online publication: **skydan.in.ua/T03DSP/CWExciter.htm**.

Alex Shovkoplyas, VE3NEA (ex-UB5EMI and UR5EMI), has been licensed since 1980. He attended Dnipropetrovsk State University in the Ukraine, and majored in Radio-Physics and Electronics. Founder and owner of Afreet Software Inc, he is employed as a software developer with Resume Mirror Inc. He is the author of Morse Runner, HamCAP, DX Atlas *and other software. His Web sites are* **www.dxatlas.com** *and* **www.netvampire.com/Ham**. □□

The IQPro: A High-Performance Quadrature DDS VFO

This update of a 1997 signal generator project uses the AD9854 digital synthesizer.

Craig Johnson, AAØZZ



Introduction

In the July 1997 issue of *QEX*, Curtis Preuss, WB2V, published an article featuring a signal generator that used an Analog Devices AD9850 DDS device controlled by a PIC16F84 microcontroller. Since that time, a newer DDS device has been introduced by Analog Devices: the AD9854. This device has some significant advantages over the AD9850.

First, the AD9854 produces two outputs with a 90° phase difference. This phase relationship remains constant regardless of the frequency. Second, the AD9854 has a 12-bit DAC (digital-to-analog converter) while the AD9850 has a 10-bit DAC. This is important in reducing undesirable byproducts (spurs).

Third, the version of the AD9854 used in this project has a 200 MHz maximum clock speed, compared to 125 MHz for the AD9850. The faster clock speed raises the Nyquist limit, allowing the VFO to operate cleanly through 30 MHz.

This project is an extension of the concepts introduced in that original AD9850 project, but there have been many major changes. It still has an optical encoder for frequency selection and it still uses an LCD for displaying the frequency. However, the IQPro VFO has been changed in these ways:

- The IQPro VFO now uses two PIC microcontrollers: one controlling the user interface (optical encoder, push buttons and LEDs) and the other controlling the AD9854, the LCD and the band-switching relay interface.
- The IQPro VFO now uses a two-line LCD so more information is displayed.
- The IQPro VFO now operates as two VFOs with two frequencies. The active VFO is selectable via a push button.
- The two-part VFO supports split-frequency operation. This means you can transmit on one frequency and receive on another.
- The two AD9854 outputs are amplified to produce signal levels that are compatible with common receiver requirements (such as the R2, MiniR2, Binaural or R2Pro by Rick Campbell, KK7B).

Theory of Operation

Early versions of this VFO project (WB2V follow-on) used a single PIC microcontroller for all VFO functions. However, the PIC's main code execution loop was very large and it varied a great deal in length, depending on if the LCD and DDS needed to be updated and if push buttons were being pressed. Since the single PIC needed to monitor the input signals from the optical encoder, detecting the number of changes in the "gray code" to determine how fast the encoder was being turned, the variable size main execution loop made accurate timing and smooth operation very difficult to accomplish.

In the next version of project, which I developed with Bruce Stough, AAØED, we decided to use a dedicated PIC16F628 microcontroller to handle the optical encoder. Why? They are inexpensive and provide a rather elegant solution to the problem of accurate timing for smooth tuning operation.

It is changed even more in the IQPro. Now there are two 16F877 PICs. One is defined as the Interface PIC and the other is the Driver PIC. The encoder logic has been moved into the Interface PIC. The Interface PIC monitors the gray code output of the encoder and

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periodically sends information to the Driver PIC (by way of an encoder message) to indicate how far the encoder shaft has been turned since the last update, as well as the direction of turn. The Interface PIC sends this data to the Driver PIC by setting up the tick-count (up to 6 bits, indicating the number of encoder transitions detected since the last update) and the direction bit in a message byte and then sending this encoder message to the Driver PIC. When the Driver PIC gets the encoder message, it extracts the tick-count and direction from the message, updates the working registers with the new frequency, and sets a flag that indicates the DDS and LCD need updating. A very short time later, code in the main execution loop of the Driver PIC will detect the "changed" flag and call the appropriate routines to update the frequency in the DDS and the LCD.

The Interface PIC also handles the push buttons. When a push button is pressed, the Interface PIC detects it and sends a message to the Driver PIC, causing the Driver PIC to take the appropriate action. Similarly, whenever the Driver PIC needs to turn an LED on or off, it sends a message to the Interface PIC and the Interface PIC performs that action. In addition, the Driver PIC sets up output signals on a header when the VFO frequency changes to a different band. These signals are intended to set and reset relays to activate band-specific filters that remove harmonics from a transmitter output before sending the signal to the antenna.

The IQPro takes another step forward in improving the "user interface." The Interface PIC handles all operator inputs and outputs. In addition to handling the optical encoder, the push buttons and the LEDs are controlled by the Interface PIC. The Driver PIC handles the AD9854, the LCDs, and the filter relays on the Filter/Relay Board.

The push button board is a small board on which 24 push buttons can be mounted. The push button board interface header plugs into the main (control) board's header HDR3. For testing purposes, the builder can install the push buttons directly on the push button board; however, the builder will eventually want to move the push buttons to the front panel of the project's chassis. It is still convenient to use the push button board since it provides an easy way of connecting push buttons to the four rows and six columns. Even so, there are several options. The push button board can remain plugged into the main (control) board header with individual wire-pairs to bring the push buttons to the front panel or the push button board can be mounted directly behind the front

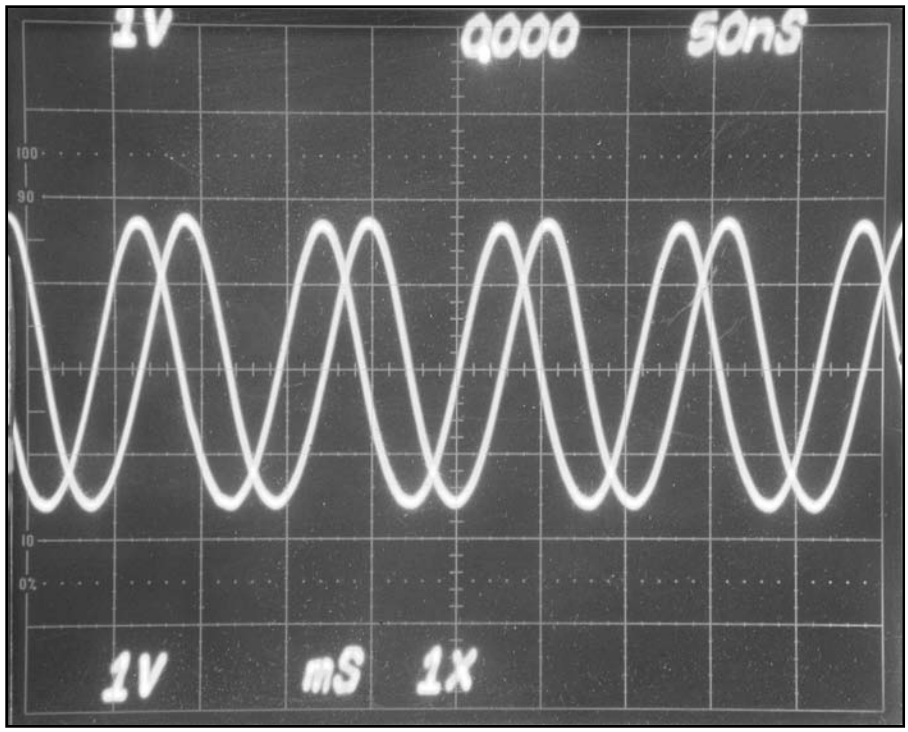


Figure 1 — A snapshot of I and Q signals generated by the AD9854.

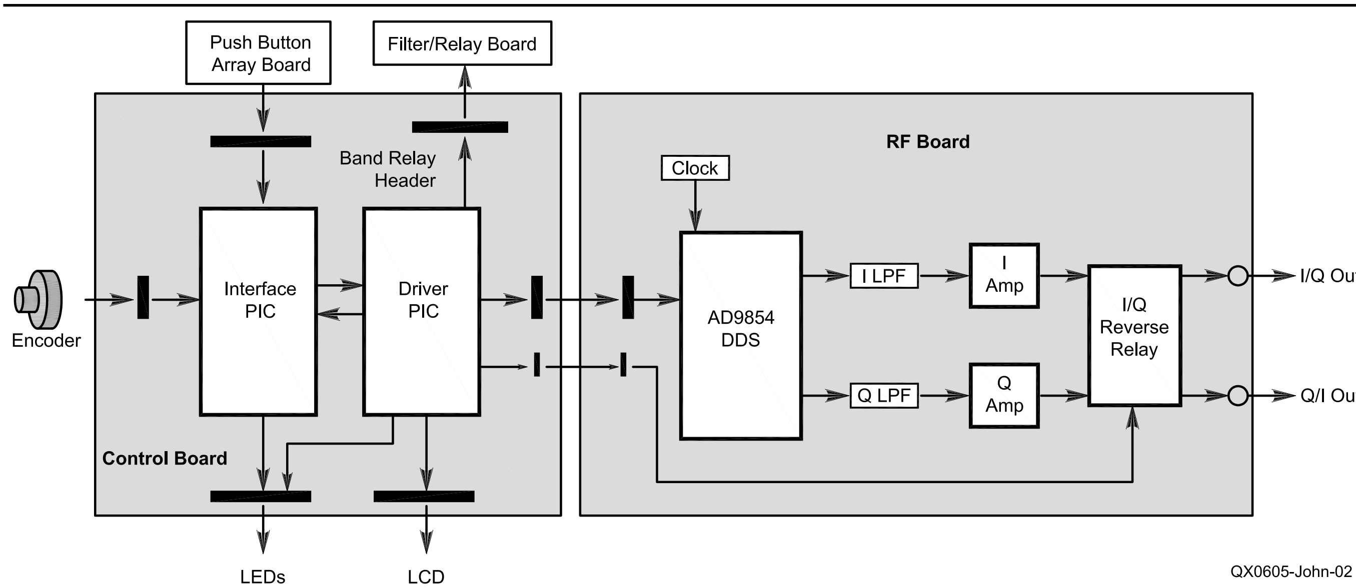
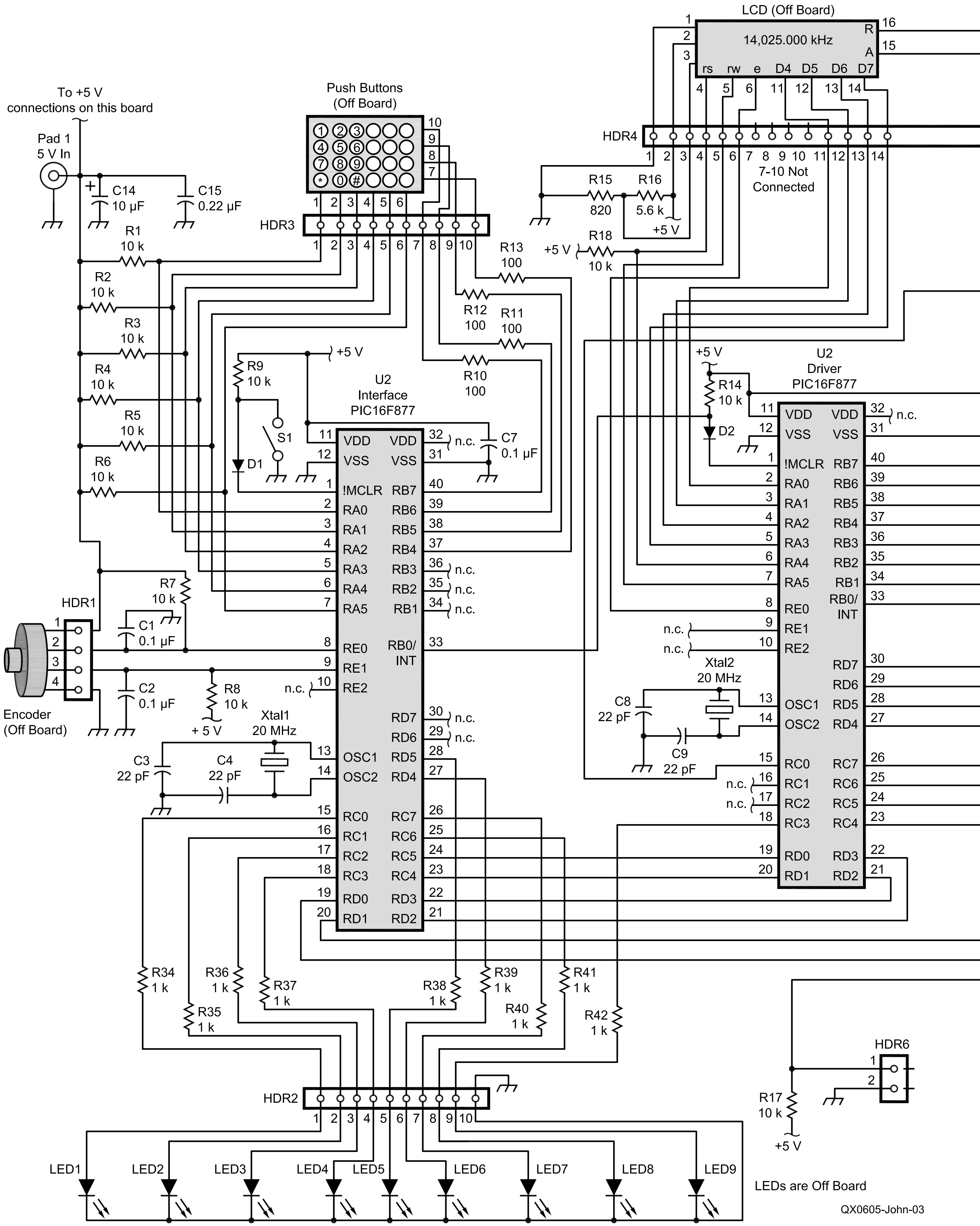


Figure 2 — IQPro block diagram.


LCD (Off Board)
14,025.000 kHz
rs rw e D4 D5 D6 D7
R A
HDR4
7-10 Not Connected
R15 820
R16 5.6 k
R18 10 k
+5 V
Push Buttons (Off Board)
HDR3
To +5 V connections on this board
Pad 1 5 V In
C14 10 µF
C15 0.22 µF
R1 10 k
R2 10 k
R3 10 k
R4 10 k
R5 10 k
R6 10 k
R7 10 k
R8 10 k
R9 10 k
R10 100
R11 100
R12 100
R13 100
S1
D1
C7 0.1 µF
U2 Interface PIC16F877
VDD VSS !MCLR RA0 RA1 RA2 RA3 RA4 RA5 RE0 RE1 RE2 OSC1 OSC2 RC0 RC1 RC2 RC3 RD0 RD1
VDD VSS RB7 RB6 RB5 RB4 RB3 RB2 RB1 RB0/INT RD7 RD6 RD5 RD4 RC7 RC6 RC5 RC4 RD3 RD2
n.c.
HDR1
C1 0.1 µF
C2 0.1 µF
Encoder (Off Board)
Xtal1 20 MHz
C3 22 pF
C4 22 pF
U2 Driver PIC16F877
R14 10 k
D2
Xtal2 20 MHz
C8 22 pF
C9 22 pF
R34 1 k
R35 1 k
R36 1 k
R37 1 k
R38 1 k
R39 1 k
R40 1 k
R41 1 k
R42 1 k
HDR2
HDR6
R17 10 k
LED1 LED2 LED3 LED4 LED5 LED6 LED7 LED8 LED9
LEDs are Off Board
QX0605-John-03

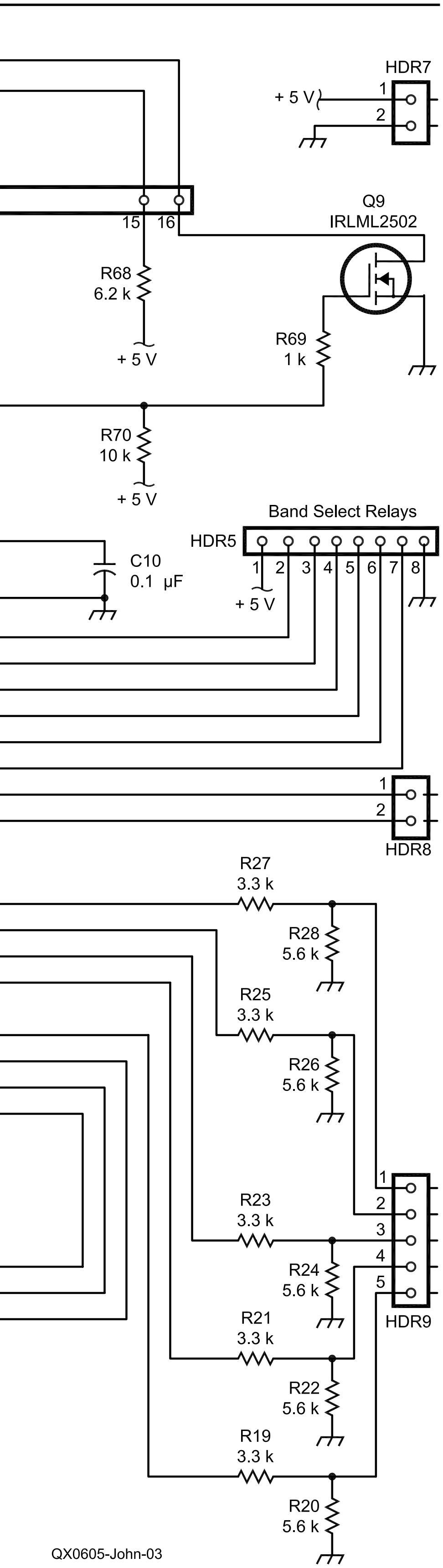


Figure 3 — Schematic diagram of the IQPro control board.

panel, connected to header HDR3 by way of a 10-wire harness.

The LED board has positions for the nine LEDs to be mounted on it. For testing purposes it can be plugged directly into HDR2 of the main (control) board. Then it can be moved to the chassis front panel or individual wires (one wire per LED, plus a common ground) can be moved to the front panel.

The band-switch test LED board has six LEDs and is for testing purposes only. It plugs into main (control) board header HDR6. This header provides a way to attach an 8-wire harness to bring the six relay signals (plus the ground and the optional +5 V power line) to an external filter/relay board (not supplied).

The I and Q signals generated by the AD9854 DDS pass through low-pass filters and are amplified before going off-board to the receiver or transmitter. The I/Q reverse relay, activated by pressing a push button, is used to switch the I and Q signals before going off-board. This allows the transmitter or receiver to toggle between upper and lower sideband. For additional details see the "I-Q Reversal" section.

Figure 1 shows a snapshot of the I and Q signals generated by the AD9854. Note that this 90° (quadrature) phase relationship is consistent for all generated frequencies, from 0 to 30 MHz. Figure 2 shows a block diagram of the IQPro.

Control and RF Board Descriptions

Figures 3 and 4 show the two-part schematic for the IQPro. The IQPro is divided into two boards: a control board (Figure 3) and an RF board (Figure 4). The reason for the division into two boards is to allow the builder the option of completely shielding the RF board, if desired. If they are separated, it is suggested that the RF board be mounted in an RF-shielded box (eg, Hammond 1590BB) and feedthrough capacitors used to bring the power and control lines from the control board into the RF board in the shielded box. If the control and RF boards are not separated, jumpers need to be installed between the matching pins of headers HDR7 and HDR10, HDR8 and HDR11, and HDR9 and HDR12.

The two socketed PIC microcontrollers control the IQPro functions. Both are type 16F877 and are 40-pin parts. Each PIC pin that is used as an input has a 10 kΩ pull-up resistor attached to +5 V. The optical encoder feeds its two gray code signals to the Interface PIC.

The RF board uses a 125 MHz SMT clock oscillator for the AD9854. This is a trade-off. We could have used a lower frequency clock oscillator and then used the AD9854's internal clock multiplier but that introduces additional spurious products as well.

The output current (I_{out}) from the AD9854 is set by the size of the R_{set} resistor. We are using an R_{set} of 3.9 kΩ, which translates to 10 mA of output current. (20 mA is maximum, but 10 mA is selected because it provides the best Spurious Free Dynamic Range — SDFR.)

The two boards together measure 3 × 6 inches. There is a "partial-cut" line between the two boards so the builder can choose to break them apart or to keep them together. If they are kept together as a single board, simple jumpers can be installed between the appropriate pins in HDR7 and HDR11, HDR8 and HDR11, and HDR9 and HDR12. If you choose to mount the RF board in an RF shielded box, it fits nicely into a Hammond aluminum diecast box (Hammond 1590BB, DigiKey part number HM152-ND), which is 3.7 × 4.7 × 1.2 inches.

External power to the control board (passed to the RF board) is 5 V. If the transistor amplifier option is selected, +12 V is also needed. There is no 12 V to 5 V converter on the IQPro boards. However, a small converter board is supplied with the IQPro kit.[1] It can be mounted at any convenient location in the case.

The AD9854 must run at 3.3 V, so its reference clock (125 MHz) is also a 3.3-V part. However, since the PIC16F877 runs at 5 V, a method of converting the voltages of the signal lines that connect these PIC16F877 ICs to the AD9854 was needed. It is done with simple voltage-dividing resistors on each signal line.

Five-volt power consumption for the entire board is about 325 mA with the LCD backlight off and about 400 mA with the backlight on. About 265 mA of that current goes to the AD9854 and the 125 MHz clock. The 12 V transistor amplifiers draw a total of 40 mA if this option is implemented.

I-Q Reversal

The AD9854 has two output signals, I and Q, with a 90° phase difference. The I and Q outputs are reversed by the activation of the IQ Reverse Relay. The signals to set and reset this latching relay are generated by the Driver PIC when the MODE CYCLE push button is pressed.

Band Relay Driver

The IQPro is set up to cover the frequency range of 0-30 MHz. As seen in the block diagrams, the IQPro is intended to drive a receiver directly with the I and Q signals.

[1]The kit is available from KangaUS — **www.kangaus.com** (Bill Kelsey, N8ET, **kanga@bright.net**). Three options are being offered:

IQPro kit including PC boards, manual, parts on the board, including pre-programed PICs, premounted AD9854 chip and a high-quality optical (not mechanical!) Clarostat encoder — $195 plus $5 s/h.

Bare board set (PC boards only; no parts, no AD9854, no printed manual) — $25 plus $2.50 s/h ($5 via Priority Mail).

Pre-programmed PICs — $25 plus $2.50 s/h ($5 via Priority Mail).

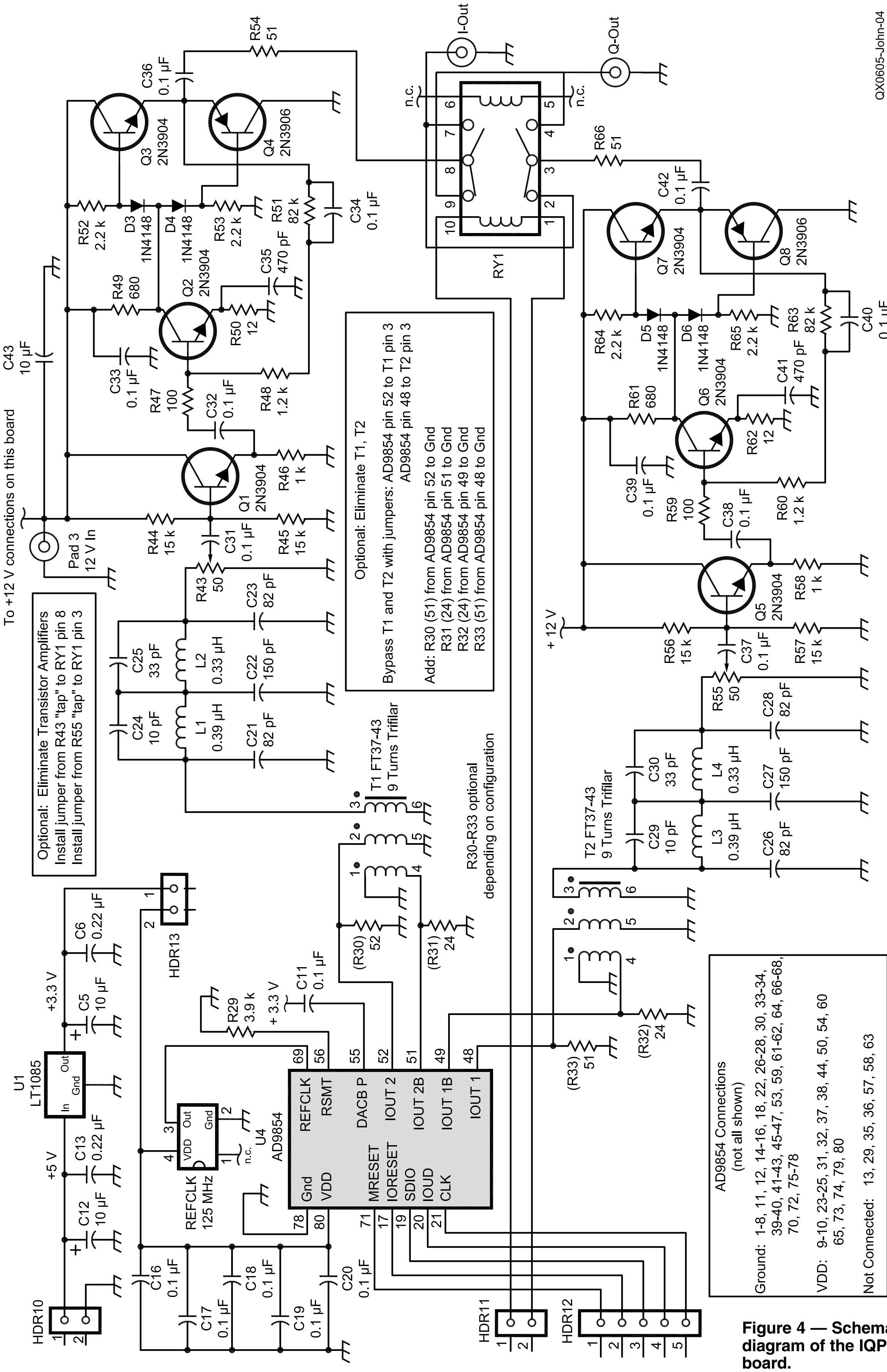


Figure 4 — Schematic diagram of the IQPro RF board.

However, transmitters need low-pass filters at their outputs, just before going to the antenna, to remove harmonics. Obviously a single low-pass filter (such as 30 MHz) cannot be used to eliminate the harmonics for the entire range. The IQPro has a scheme in which signals are generated (HDR5) which can directly trigger latching relays on another board (the filter/relay board) that can engage the proper low pass filter.

Within the IQPro frequency range there are currently nine US amateur bands (ignor-

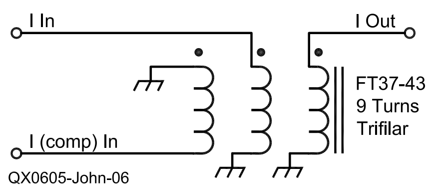


Figure 6 — The toroid transformer.

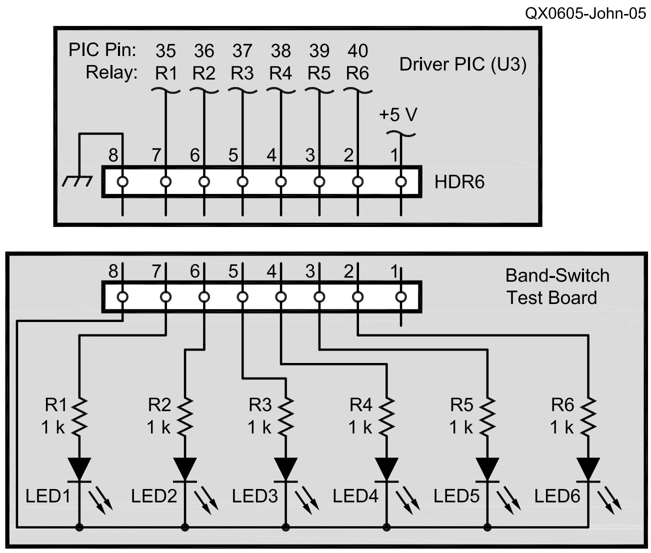


Figure 5 — Schematic diagram of the band-switch test board.

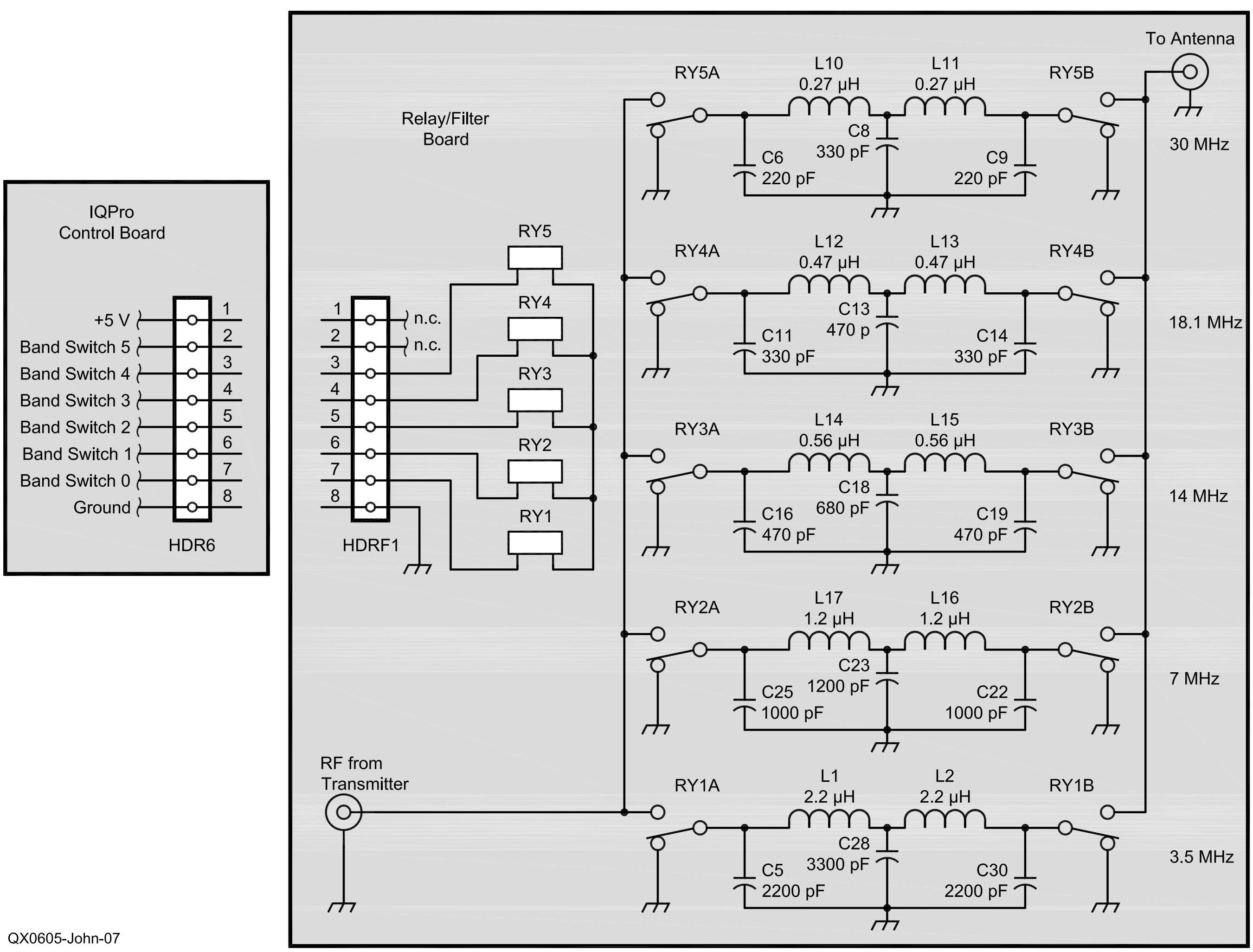


Figure 7 — An example of a 5-band filter board.

ing the 60 meter band). While nine individual filters could be used, that is not really necessary. The IQPro has a scheme in which five low-pass filters are sufficient to remove harmonics for all nine amateur bands. Builders can easily change these ranges:

- Relay 1 enables the LPF with a 3.5-MHz cutoff (for 160 m, 2nd harmonic removed).
- Relay 2 enables a LPF with a 7.0-MHz cut-off (for 80 m, 2nd harmonic removed)
- Relay 3 enables a LPF with a 14-MHz cut-off (for 40 and 30 m, 2nd harmonic removed)
- Relay 4 enables a LPF with a 18.1-MHz cut-off (For 20 m, 2nd harmonic removed)
- Relay 5 enables a LPF with a 30-MHz cut-off. (For 17, 15, 12 and 10 m, 2nd harmonic removed)
- Relay 6 (unused, available for future use.)

Since the IQPro is continuous-coverage from 0 to 30 MHz, the following eleven band numbers and ranges are implemented:

- Band Number 1 Relay 1
  0 - 1.799999 MHz
- Band Number 2 Relay 1
  1.8 - 3.499999 MHz
- Band Number 3 Relay 2
  3.5 - 6.999999 MHz
- Band Number 4 Relay 3
  7.0 - 10.099999 MHz
- Band Number 5 Relay 3
  10.1 - 13.999999 MHz
- Band Number 6 Relay 4
  14.0 - 18.067999 MHz
- Band Number 7 Relay 5
  18.068 - 20.999999 MHz
- Band Number 8 Relay 5
  21.0 - 24.889999 MHz
- Band Number 9 Relay 5
  24.89 - 27.999999 MHz
- Band Number 10 Relay 5
  28.0 - 29.999999 MHz
- Band Number 11 Relay 5
  30 MHz

The five latching band relays are activated by signals on header HDR5. Each relay on the filter board is assumed to have one side connected to a unique header pin (coming from the Driver PIC on the main control board) and the other side of each relay connected to a common connection point. Thus, to set and latch a relay, the Driver PIC simply puts a low level on all pins on the header and then raises the appropriate header pin to set that relay. The pin is kept high for 8 ms and then brought back to a low level. The selected relay is now latched in a SET state. Similarly, to reset a relay, the Driver PIC puts a high level on all header pins (causing no current to flow) and then makes the appropriate relay's header pin a low. Current will then flow through the selected relay's coil in the reverse direction and the relay will go to the RESET state. After 8 ms, all pins are brought back to a low level again. The relay is now latched in a RESET state.

This method of setting and resetting the relays is used because it minimizes the number of PIC pins (and header pins) required. Note that during a reset operation, forward current is flowing through the coils of all of the non-selected relays. However, the forward current through the non-selected relay coils is not enough to set them inadvertently.

Band-Switch Test Board

A band-switch test board is included with the IQPro kit. The schematic is shown in Figure 5. This board is designed to plug into HDR5 of the main (control) board to allow

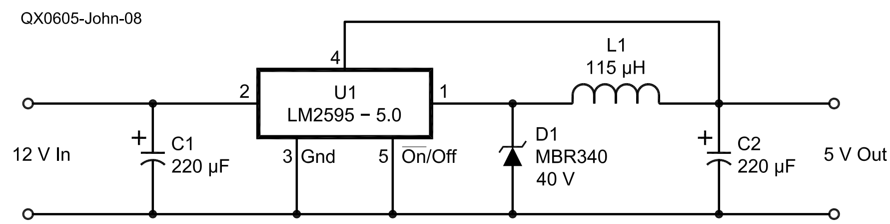


Figure 8 — Schematic diagram of the switching power supply.

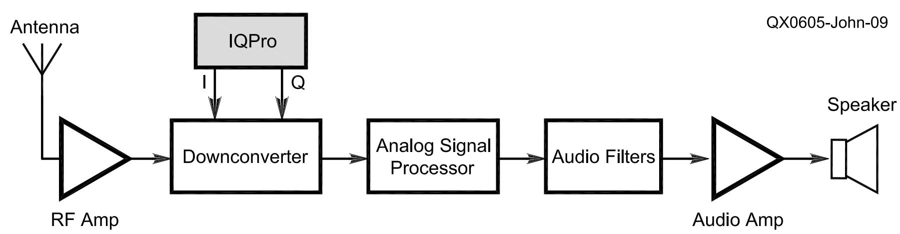


Figure 9 — Block diagram of the IQPro in R2 receiver.

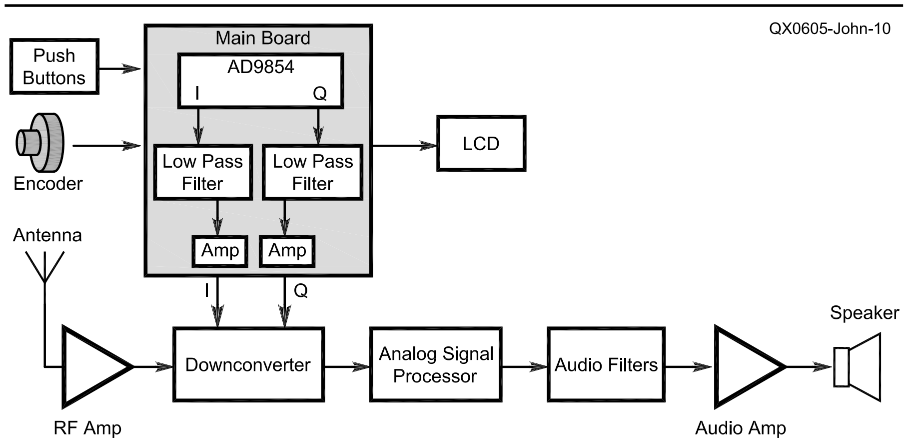


Figure 10 — Detailed block diagram of the IQPro in R2 receiver.

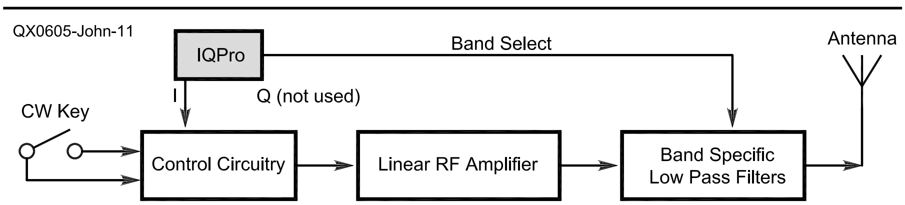


Figure 11 — Block diagram of the IQPro in a CW transmitter.

the builder to see the relay signals pulse as the frequency ranges change. Surface mount (size 0805) 1 kΩ resistors are used on this test board. See the Relay/Filter Board section for more information about implementing a relay/filter board.

I and Q Amplifiers

The I and Q signals generated by the AD9854 are not sufficient in amplitude to drive most mixers. As mentioned in the RF board section, the output current (I_{out}) from the AD9854 is set by the size of the R_{set} resistor. The default value for R_{set} of 3.9 kΩ, which translates to 10 mA of output current. 20 mA is maximum, but 10 mA provides the best Spurious Free Dynamic Range (SDFR).

The output current of 10 mA into a 50 Ω load translates into 500 mV p-p (177 mV RMS), which is about –2 dBm. However, many common mixers (such as the TUF-3 used by the R2Pro) require +7 dBm signal levels so some amplification is required.

There are three amplifier options available on the IQPro board:

Option 1**:** ***Toroids only (no transistor amplifiers).*** In this option, two FT37-43 toroids with trifilar windings are used. The toroid connections go directly to the AD9854. The transistor amplifiers parts are not installed. The AD9854's two complementary outputs are "added" to the primary outputs by way of the toroid transformers.

If +7 dBm signal outputs are required, the nominal value of 10 mA output on the 4 signals is not sufficient. To increase the I and Q signals to 14 mA, change the R_{set} resistor (R29) from 3.9 k to 2.7 k. Figure 6 shows the basic schematic for the trifilar transformer. (This is duplicated for the Q signal.)

Pros: 1) The two complementary outputs from the AD9854 as well as the two "normal" outputs are used, and 2) no 12 V power is needed on the IQPro board.

Cons: 1) Two toroids with trifilar windings must be constructed and connected properly, 2) if +7 dBm outputs are wanted, the R_{set} resistor (R29) may have to be changed from 3.9 kΩ to a smaller value such that more current is put out by the two AD9854 outputs. This might cause a slightly higher level of spurious output products.

To use this option:

Termination resistors R30-R33 are *not* installed on the board.

Install a wire jumper from the output of each Low-Pass Filter to the relay (RY1) inputs.

Transistor amplifier components are *not* installed.

Option 2**: Transistor amplifiers only. If this option is selected, the transistor amplifier components are used and the two toroids are not installed. This gives +7 dBm signal outputs.**

Pro: The AD9854 should be putting out the cleanest signal (highest Spurious Free Dynamic Range) with the nominal R_{set} value of 3.9 kΩ (corresponding to 10 mA output on I and Q).

Con: +12 V power is required.

To use this option:

Termination resistors R30-R33 are installed.

R_{set} resistor (R29) is 3.9 kΩ.

Toroids T1 and T2 are *not* installed.

Transistor amplifier components are installed.

Install a wire jumper from pin 2 to pin 3 of T1 and from pin 2 to pin 3 of T2.

Option 3: **Toroids and Transistor amplifiers. Use this option if signal outputs larger than +7 dBm are needed. With R_{set} (R29) at its "nominal" value of 3.9 kΩ, producing 10 mA signals, the typical outputs of +17 dBm are produced.**

Pro: This option produces the highest output signal levels of the three options.

Con: +12 V power is required.

To use this option:

Termination resistors R30-R33 are *not* installed.

R_{set} resistor (R29) is 3.9 kΩ.

Toroids T1 and T2 are installed.

Transistor amplifier components are installed.

Relay/Filter Board

A relay/filter board is not included in the IQPro kit because requirements for individual applications will vary widely. The concept is shown in Figures 7, 12 and 14. Note that the two sides (a and b) are implemented with the two poles of DPDT relays rather than with separate relays for the two sides. The band relay latching signals from header HDR5 are pulsed when the frequency changes to a new band range. See the Band Relay Driver section regarding generation of the band relay signals.

The signals generated by the Driver PIC and presented on HDR5 are pulsed for approximately 8 ms. The amount of current that is "sourced" by a PIC pin (to HDR5) is limited to about 20 mA, which is exactly the amount of current that this particular type of relay (RY1) requires to set or reset. However, 20 mA is probably *not* sufficient to drive

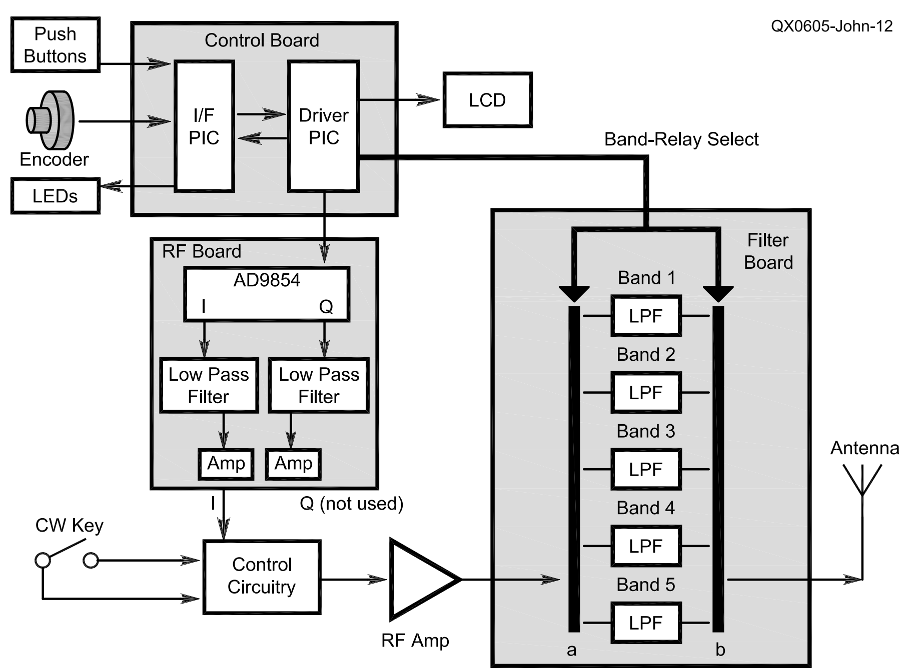


Figure 12 — Detailed block diagram of the IQ Pro in a CW transmitter.

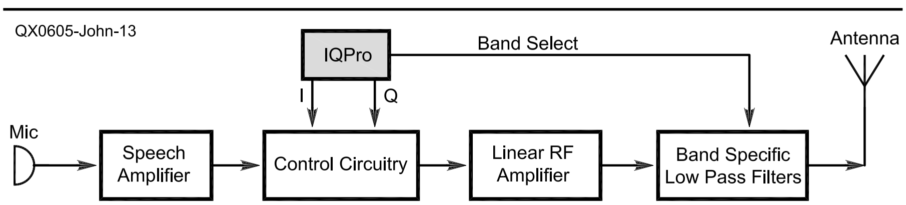


Figure 13 — Block diagram of the IQ Pro in an SSB transmitter.

several relays in parallel. The design assumes 20 mA single coil latching relays are used, each with a common side tied together. To set a relay, the PIC drops all of the band-switch pins to a low level and then raises the single pin for the relay to be set to a high level. Approximately 20 mA of current flows in the "forward" direction through the coil and sets (and latches) the relay. Note that this implementation requires that the other relays in the bank will "share" the return current (5 mA each). To reset a relay, the PIC raises all of the band-switch pins to a high level concurrently and then drops the single pin of the relay to be reset. Approximately 20 mA of current flows in the "reverse" direction through the relay's coil and resets (and latches) the relay. Note that the 20 mA current in the "forward" direction flows through the other four relays in the bank but is low enough such that it does not set any of them.

If you want to drive several relays concurrently from one HDR5 pin you should consider using switching components that in turn enable sufficient current flow to latch the relays. MOSFETs (eg, 2N7000 or IRLML2502 or IRF510) work well in this type of application. For an example, see the LCD backlight enable mechanism using Q9 (an IRLML2502 MOSFET) in the main (control) board schematic (Figure 3). Q9 is driven by a PIC pin that enables the LCD backlight current (75 mA) to flow when the PIC pin is set to a high level.

Note that the PIC software remembers which relay was previously set and resets it before setting the new relay. The exception is on power-on in which case the previous value is unknown. For this reason, the PIC initialization routine resets all the relays, one at a time, before setting the default relay.

Switching Power Supply Board

The main board requires a 5 V input power source and it is customary for most builders to use 12 V power for their equipment. To make this conversion easy, a small switching power supply board is supplied with the IQPro kit. This power supply board converts 12 V to 5 V without dissipating a lot of heat. Figure 8 shows the schematic for this power supply. The conversion from 5 V to 3.3 V (required by the AD9854 DDS and its reference clock) is done on the board.

PIC Software

AD9854 Initialization

The IQPro runs the AD9854 in serial mode. This means it uses one data line, a clock line, and an external update line to send the frequency updates from the PIC to the AD9854. Two reset lines are also controlled by the PIC.

Determining how to get the AD9854 initialized was a real challenge, especially with a 125 MHz reference clock. The problem stems from the fact that the AD9854 powers up in internal update mode while it really needs to be in external update mode to operate in this type of application. See the AD9854 data sheet for details.

The AD9854 initialization sequence is sent from the driver PIC to the AD9854 immediately after the IQPro is powered up and after the reset push button is pressed. However, the AD9854 initialization signals are fairly short in duration and may be difficult to observe with an oscilloscope. To aid in debugging, a special debug mechanism was implemented in the Driver PIC code. This code, when triggered, forces the Driver PIC to go into an endless loop in which the initialization sequence is continually repeated. This debug mode is activated by installing a jumper between the two pins of header HDR6 during power up. The initialization sequence will be continuously sent until the short between the two header pins is removed. Now the signals can easily be seen with an oscilloscope. When the jumper is removed the PIC will continue initializing the IQPro.

IQPro Applications

Figures 9 through 14 illustrate some examples of how the IQPro can be used in a variety of applications.

IQPro in R2 (or R2Pro, etc) Receiver
IQPro in CW Transmitter
IQ Pro in SSB Transmitter

DDS Frequency Calculation Details

When we want the AD9854 to generate a sine wave at a particular frequency, we need to send it a 6-byte number referred to as the Frequency Tuning Word. The AD9854 uses the Frequency Tuning Word and the reference clock to generate the output waveform.

The AD9854 can generate an output frequency with 48-bit accuracy. However, in this application, since we are always going to be generating a frequency that is a whole number of hertz, we keep a 4-byte numeric constant in the PIC's EEPROM that represents a 1 Hz change in output frequency when used with the specified reference clock. We refer to these 4 bytes as:

ref_osc_3 (the most significant byte)
ref_osc_2 (next byte)
ref_osc_1 (next byte), and
ref_osc_0 (the least significant byte)

Note that the AD9854 has a 48-bit frequency tuning word and we are only saving a 32-bit constant in EEPROM. For performance reasons, and since 48-bit accuracy is not required for this application, we will construct the 48-bit tuning word from the 32-bit value. We will use this value to calculate the 48-bit frequency tuning word. We will then multiply the 32-bit value by the desired output frequency, in hertz, to obtain another 32-bit product. This 32-bit product will be used as the most significant 32 bits of the 48-bit frequency tuning word and we'll fill the remaining 16 bits with zeros.

This numeric constant is interpreted as a fixed-point real number in the format:

<refosc3> . <refosc2> <refosc1> <refosc0>

where:

refosc3 = (2^{32}/ osc_freq_in_Hz)

refosc2, **refosc1**, and **refosc0** are the fractional part of: (2^{32} / osc_freq_in_Hz) times 2^{24}.

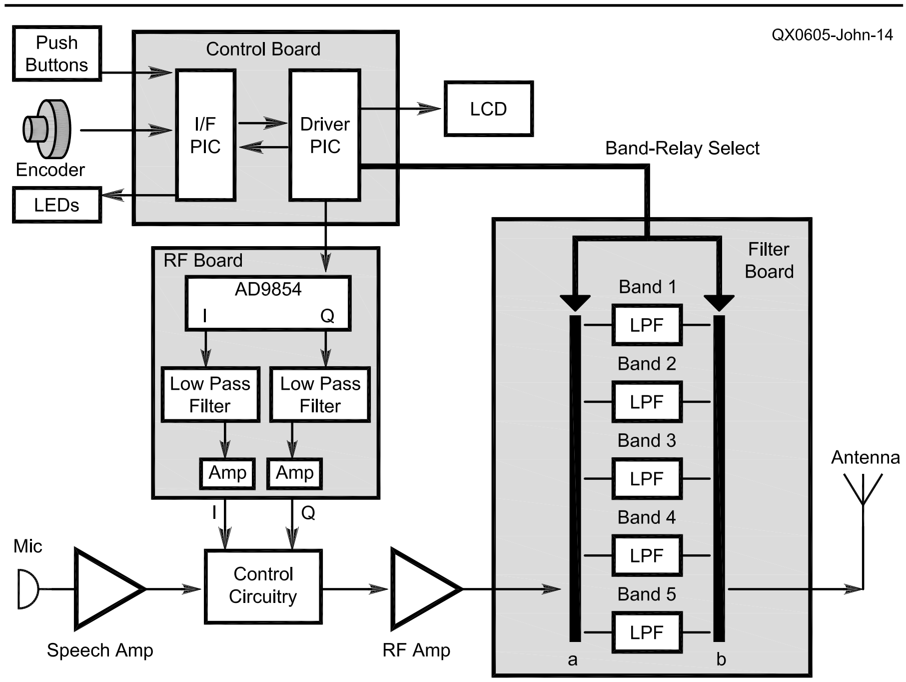


Figure 14 — Detailed block diagram of the IQPro in an SSB transmitter.

Example:

For a 125 MHz clock:

refosc3 is (2^{32} / 125 × 10^6) = 34.359738368 truncated to 34 (Hex 22)

refosc2 is the high byte of (.359738368 × 2^{24}) = 6035408.303

6035408.303 = Hex5C17D0, so the high byte is 5C.

refosc1 is the next byte of Hex 5C17D0, or 17

refosc0 is the last byte of Hex 5C17D0, or D0

The values for common oscillator frequencies are shown in Table 1.

Once again, we can think of this constant number as the step size that represents 1 Hz of movement. When we want to set the VFO's frequency to a specific number of hertz, we simply have to multiply this constant by the desired number of hertz to obtain a Frequency Tuning Word. We send this Frequency Tuning Word to the AD9854 and, *voila*, it generates the sine wave at the specified frequency. In fact, it generates two outputs exactly 90° apart.

Now it is obvious why the Calibrate function is necessary. When the PIC microcontroller is programmed, preset values are loaded into the PIC's EEPROM, based on the nominal frequency of the AD9854's external reference clock. However, it is never absolutely accurate as specified. The Calibrate function is used to adjust the 4-byte constant (step size) such that when the AD9854 uses it with this reference clock, the output frequency is accurate. The updated 4-byte constant is then stored back in EEPROM for future use in calculating the Frequency Tuning Word. It is not lost when the IQPro is powered down.

Inter-PIC Communications

A completely home-grown synchronous communications scheme is implemented in the IQPro to pass messages between the PICs. Yes, the PIC16F877 supports two industry-standard synchronous communications schemes (SPI and I^2C) as well as an asynchronous scheme (USART/SCI), but I chose not to use them. Why not? Because these communications schemes are really designed for efficiency in long data transfers but are rather inefficient for the single-byte messages of the IQPro.

The IQPro synchronous message-passing scheme uses the following general principles:

Use SYNC sequence: 111111111 (nine 1-bits). Note that this sequence of bits is impossible for a "normal" data byte plus parity.

Send a data byte consisting of 8 data bits followed by a parity bit.

Use EVEN parity

The first valid data code would be 00000000+p (which is 00000000 0)

The last valid data code allowed would be 11111111+p (which is 11111111 0)

Gives 256 usable codes. All 8 data bits are usable.

This communications scheme uses three data lines to send a message from one PIC to another. Two of these lines are controlled by the "originating" PIC (Data and Clock) and one line is controlled by the "receiving" PIC (Acknowledge). Two sets of lines are used to implement the two-way message passing.

Assume a message is to be passed from PIC-A to PIC-B. The timing diagram of Figure 15 shows the signals as one bit is transferred from PIC-A to PIC-B.

To transfer a single BIT from PIC-A to PIC-B the following actions are taken:

PIC-A detects B-ack is low. (If not low, wait until it is.)

PIC-A sets A-data to 0 or 1

PIC-A sets A-clk high

PIC-B detects A-clk high and extracts data from data line

PIC-B raises B-ack.

PIC-A detects B-ack high (within × cycles) and drops A-clk (and A-data)

PIC-B detects A-clk is low and drops B-ack.

Table 1
Hex Values for Common Oscillator Frequencies

Frequency	*refosc3*	*refosc2*	*refosc1*	*refosc0*
125.00 MHz	22	5C	17	D0
120.00 MHz	23	CA	98	CE
100.00 MHz	2A	F3	1D	C4
90.70 MHz	2F	5A	82	7A
66.66 MHz	40	6E	52	E7
66.00 MHz	41	13	44	5F
60.00 MHz	47	95	31	9C
50.00 MHz	55	E6	3B	88

Table 2
Messages Sent from the Interface PIC to the Driver PIC

Interface to Driver Messages (Hex)	*Description*
0	Keypad 1
1	Keypad 2
2	Keypad 3
3	MHz UP
4	Split Toggle
5	VFOA/VFOB Toggle
06-0F	(unused)
10	Keypad 4
11	Keypad 5
12	Keypad 6
13	MHz DN
14	Mode Cycle
15	VFOA=VFOB
16-1F	(unused)
20	Keypad 7
21	Keypad 8
22	Keypad 9
23	Band Up
24	Tone Set Toggle
25	Fast Tune Toggle
26-2F	(unused)
30	Keypad *
31	Keypad 0
32	Keypad #
33	Band Down
34	Backlight Toggle
35	Calibrate
36-3F	(unused)

Table 3
Messages Sent from the Driver PIC to the Interface PIC

Driver to Interface Messages (Hex)	*Description*
40	LED 1 ON (SSB) and LED 2 OFF
41	LED 2 ON (CW) and LED 1 OFF
42	LED 3 ON (VFOA) and LED 4 OFF
43	LED 4 ON (VFOB) and LED 3 OFF
44	LED 5 ON (SPLIT)
45	LED 6 ON (FTUNE)
46	LED 7 ON (CAL)
47-53	(unused)
54	LED 5 OFF (SPLIT)
55	LED 6 OFF (FTUNE)
56	LED 7 OFF (CAL)
57-5F	(unused)
80-8F	ENCODER messages

(Go back to step 1 for next bit)

Note that this scheme is completely synchronous so it is very fast. There are no "baud rates" to set up.

How well does it work? As implemented in the IQPro hardware and software, with the PICs running at 20 MHz, a message can be passed in 165 microseconds (measured).

Yes, there are some dangers in this scheme. The primary danger is if a collision occurs — with both sides originating messages simultaneously. Since the PIC code is not multi-threaded (it doesn't release control and switch between activities), if both PICs start executing the "SEND" code simultaneously, neither side will be able to receive a message so the message-passing will fail. This will result in a timeout. The Driver PIC is designed to time out first. The loop constants in the Driver receive_bit and send_bit code are smaller for the Driver PIC so the Interface PIC will detect the condition and will reset the Driver PIC. Obviously this is annoying so the occurrences must be extremely rare to make this acceptable.

How are collisions avoided? By being careful about when messages are originated in the Driver PIC.

Table 2 shows the messages that are sent from the Interface PIC to the Driver PIC. Table 3 shows the messages that are sent from the Driver PIC to the Interface PIC.

IQPro Operation

Optical Encoder

A good quality optical encoder is recommended for the IQPro, although a mechanical encoder will also work. Note that capacitors C1 and C2 are installed to eliminate the "noise" generated by the bounce of the switches in a mechanical encoder.

The recommended optical encoder has 128 positions per revolution. At design time, we decided we want the PIC to be able to capture all transition pulses when we turn the tuning knob (optical encoder) at a rate of 2 revolutions per second. When we turn the knob faster than 2 revolutions per second we allow some transitions to be missed. The IQPro was designed to send encoder messages from the Interface PIC to the Driver PIC every 25 ms. At this rate, the calculations for the maximum number of transitions per update is given this way:

512 transitions/rev × 2 rev/second

= 1024 transitions/sec

1024 transitions/sec = 0.98 ms / transition

$$\text{Then: } \frac{25 \text{ ms/update}}{0.98 \text{ ms/transition}}$$

= 25.5 transitions/update

We can send this number over to the Driver PIC with 5 "binary" bits. Since we have 6 bits (plus a direction bit) available we can exceed our design requirement and use the transitions when the tuning knob is turned even faster than 2 revolutions per second.

The IQPro has a communications scheme in which 8-bit messages are passed. (See the Inter-PIC Communications section for details regarding the inter-PIC communications

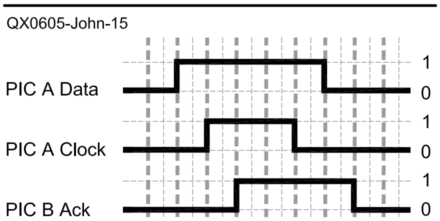


Figure 15 — Timing diagram showing the signals as one bit is transferred from PIC-A to PIC-B.

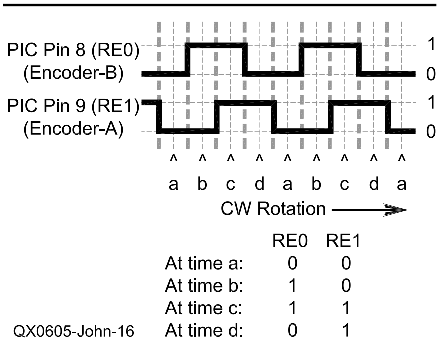


Figure 16 — Timing diagram that shows how the gray code works.

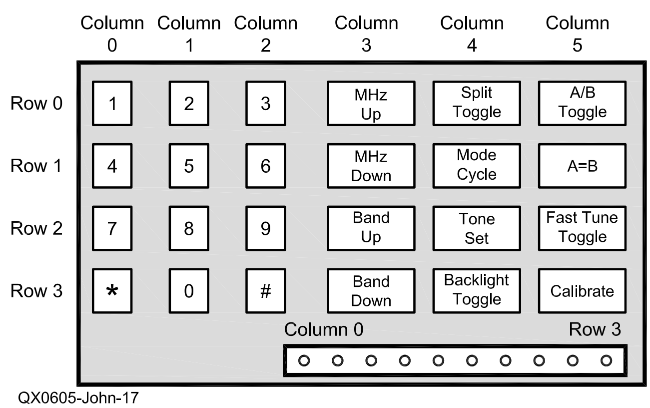


Figure 17 — The format of the 24 push buttons as they are arranged on the array board.

Table 4
How to Spread out the Tuning Rate

Tick Count Range	*Maximum Rev Per Sec*	*Normal Tune Mult Factor*	*Normal Tune Max Hz Per Update*	*Normal Tune Max Hz Per Sec*	*Fast Tune Mult Factor*	*Fast Tune Max Hz Per Update*	*Fast Tune Max Hz Per Sec*
0	0.1	1	1	200	1	1	200
1	0.2	2	2	400	8	8	1600
2-3	0.3	4	12	2400	16	48	9600
4-7	0.6	8	56	11,200	32	224	44,800
8-15	1.3	8	120	24,000	32	480	96,000
16-31	2.5	16	496	99,200	64	1984	396,800
32-63	5.0	16	1008	201,600	64	4032	806,400

scheme.) Since encoder messages are the most common message passed and efficiency is essential, a special scheme was implemented in which bit 7 of the message byte indicates that it is an encoder message. Then, bit 6 of any encoder message indicates the direction (1 = clockwise). This leaves 6 message bits (maximum value of 63) to pass the number of transitions that were detected in the past 25 ms period. A slight variation was implemented in which 0 indicates one transition and no message gets sent if no transitions are detected within the 25 ms period. Thus, 64 possible values can be sent in a message.

One additional factor that needs to be discussed is the PIC processing overhead that occurs on every update. The PIC processor must stop looking for encoder transitions when it is busy sending a message to the Driver PIC. Any transition that occurs while the PIC is busy sending a message will be lost. Fortunately message passing is very fast (about 165 µs) so the length of time that the PIC is taken away from looking for encoder transitions is very small (165 µs out of 25 ms).

The two encoder data pins are connected to the Interface PIC. Outputs A and B are the two encoder output signals and give gray code outputs (90° out of phase — quadrature). Figure 16 shows how gray code works while the algorithm that the PIC uses to determine the direction the encoder is rotating is as follows:

Going UP, the sequence is a,b,c,d,a,b,c,d,a, etc, so the sequence is: 00, 10, 11, 01, 00, 10, 11, 01, 00, etc.

Going DOWN, the sequence is a,d,c,b,a,d,c,b,a, etc, so the sequence is: 00, 01, 11, 10, 00, 01, 11, 10, 00, etc.

To determine if the sequence is UP or DOWN:

1) Take the "Right-Bit" of any pair.

2) XOR it with the "Left-Bit" of the next pair in the sequence.

3) If the result is 1 it is UP; if the result is 0 it is DOWN.

The encoder message is sent from the Interface PIC to the Driver PIC. If no acceleration algorithm were used and the transition count was used to increment or decrement the frequency, the maximum frequency change is about 64 Hz per 25-ms update cycle, or 2.56 kHz per second. That is because an encoder message can pass 6 bits of encoder change magnitude. This is too slow for comfortable operation so some acceleration is necessary. In fact, two tuning rates were implemented (normal-tuning and fast-tuning). At the same time it is sometimes important to be able to increment/decrement in 1-Hz steps. To spread out the tuning rate, the mechanism shown in Table 3 is used.

In Table 3, the acceleration constants of 1,2,3,3,4,4 are used for normal tune. For fast tune each of the acceleration constants is increased by 2. These acceleration constants are easily modified in the code and can be changed to suit the individual's preferences.

The Maximum Revolutions Per Second column shows how fast the encoder can be turned while keeping within that tick count range. Again, this assumes a 128 pulse/revolution (512 transition) optical encoder and an update interval of 25 ms.

The update frequency was chosen carefully such that updates are not done too often. In fact, the 25 ms update interval could have been reduced to 5 ms or less. However, shorter intervals between updates produce encoder magnitude values that are always small so acceleration is rarely invoked. Again, it means that the maximum tuning rate is less than what is usually preferred.

Push Button Array Board

The IQPro has 24 push buttons to control the VFO. Twelve of them form a "telephone-type" keypad that can be used for direct frequency entry while the remaining 12 push buttons control other VFO functions. Figure 17 shows the format of all 24 push buttons as they are arranged on the push button array board.

The push buttons are connected to the PIC on the main (control) board by way of a 10-pin header (HDR3) on the main (control) board. The 10 pins represent four rows of six columns of push buttons. The push button array board has connection positions for 24 SPST push buttons, arranged on a grid of four row busses and six column busses. One side of each push button connects to a row bus and the other side of each push button connects to a column bus. Figure 18 shows a schematic of the push button array board and its connections to the main (control) board via header HDR3.

The push button array board may be

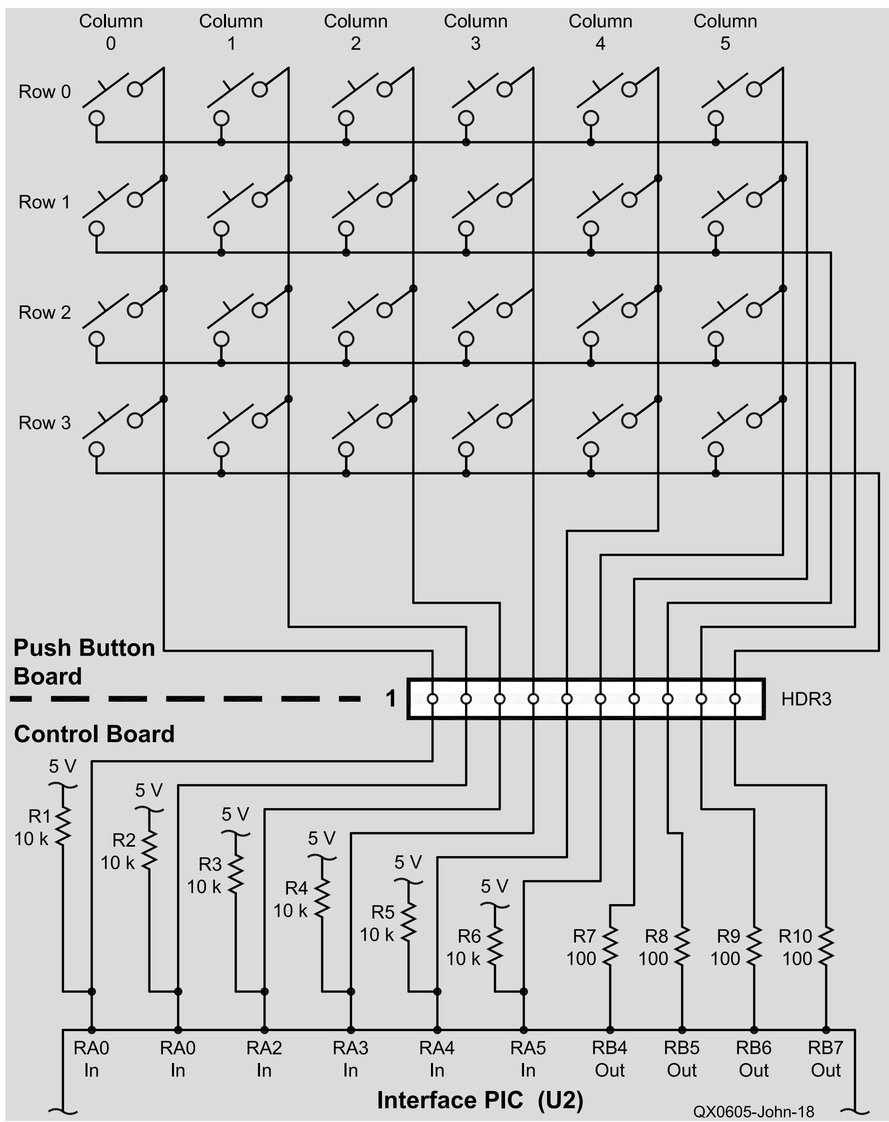


Figure 18 — Schematic diagram of the push button array board and its connections to the main (control) board via header HDR3.

directly plugged into the main (control) board's header HDR3 for testing purposes but it is intended to eventually be located on the front panel of the chassis and connected with a 10-conductor ribbon cable. The push button array board may be mounted directly behind the enclosure's front panel with the push buttons soldered into the push button array board and the push buttons protruding through holes in the front panel. However, this is not necessary and may not be optimum, given the close spacing of the push buttons on the push button array board. Individual push buttons may be relocated to the front panel and connected to the push button array board with a pair of wires.

The Interface PIC detects a push button press by executing the following algorithm:

Set all row pins high

Select one row by setting it low

Look at each column input to see if it's being pulled low by a push button.

If column input is low, that push button (row and column) is being pressed. Exit.

If all column inputs are high, select the next row and repeat

Push buttons Row-0 Col-0 through Row-0 Col-2, Row-1 Col-0 through Row-1 Col-2, Row-2 Col-0 through Row-2 Col-2, Row-3 Col-0 through Row-3 Col-2: Direct-Frequency-Input Keypad

This group of push buttons (four rows by three columns) forms a keypad that has the same format as a telephone keypad. It is used to allow the operator to jump to a desired frequency. The frequency is entered in kHz. To enter a frequency you start by pressing the "*" (star) button in the lower left corner of keypad. Then enter a number between 0 and 30,000. After entering 5 digits, the VFO will automatically jump to this frequency. To enter a frequency that has less than 5 digits, you can enter one or more leading zeros before the desired number, or you can enter the desired digits and then press the star key again to complete the direct frequency entry.

Push Button Row-0 Col-3: 1 MHz UP

Each time this button is pressed the frequency will increase by 1 MHz. Frequency maximum is 30.000 MHz.

Push Button Row-0 Col-4: Split Toggle

This push button allows the operator to toggle between NORMAL and SPLIT mode. SPLIT mode operates in conjunction with the keying interface header on the board.

When this keying circuit is activated (by grounding the key line or shorting the interface pins together) and it is in SPLIT mode, the PIC will switch from one VFO to the other. In NORMAL mode the frequency remains at the displayed frequency (VFO-A or VFO-B) as the keying circuit is activated.

The operator sets the desired transmit frequency in one VFO and the desired receive frequency in the other VFO. The current VFO selection is done by means of the Rx VFO Toggle push button. The LCD displays the frequency of the receive VFO when in receive mode and it displays the frequency of the transmit VFO when in transmit mode. The LCD displays the word XMIT when the keying circuit indicates transmit mode and RCV when in receive mode.

The LCD displays the word SPLIT when in SPLIT mode. This portion of the LCD is blank when in NORMAL mode.

Push Button Row-0 Col-5: VFOA / VFOB Toggle

Two VFO frequencies are used in the IQPro. This push button allows the operator to toggle between VFO-A and VFO-B as the currently active VFO. The LCD indicates whether VFO-A or VFO-B is active and the two VFO LEDs also indicate which VFO is active.

Push Button Row-1 Col-3: 1 MHz Down

Each time this button is pressed the frequency will decrease by 1 MHz. Frequency minimum is 0.000 MHz.

Push Button Row-1 Col-4: Mode Cycle

When this button is pressed the mode will change. The progression is from LSB to USB to CW– to CW+ and then back to LSB. The LED1 will be lit if it's LSB or USB while LED2 will be lit if it's CW– or CW+.

Push Button Row-1 Col-5: VFOA = VFOB

When this button is pressed the frequency of the VFO that is not currently being displayed will be changed to match the frequency of the VFO that is currently being displayed.

Push Button Row-2 Col-3: Band Up

Pressing this button will advance the frequency to the low end of the next higher US amateur band.

Currently the band table has entries for 9 amateur bands. (The 60 meter band was not included because of its unique spot frequency requirements.) The table can easily be updated to cover the bands for other countries.

Push Button Row-2 Col-4: Tone Set

This push button is used to change the CW offset tone. Pressing the push button will bring up a display on the LCD that shows the current tone, as retrieved from the PIC16F877's EEPROM. If the mode is LSB or USB, the tone value cannot be changed. However, if the mode is CW– or CW+, the tone value can be changed (within the range of 0-32,767 Hz) by turning the encoder. When the tone button is pressed again, the tone value is saved in the PIC's EEPROM.

When the VFO is in RCV mode, the tone is *added* to the base (transmit) frequency if the mode is CW+ while it is subtracted from the base frequency if the mode is CW–. The tone is not added or subtracted if the VFO is in XMT mode.

Push Button Row-2 Col-5: Normal/Fast Tuning Toggle

Pressing this push button toggles the optical encoder tuning rate between normal and fast tuning.

The default tuning rate allows fine tuning in 1-Hz steps when the encoder is turned very slowly. When turned slowly, one complete revolution of the encoder shaft will change the frequency by 512 Hz. However, the acceleration algorithm provides greater rate of frequency change when the encoder is turned faster. In normal tuning mode the maximum amount of frequency change in one revolution of the encoder shaft has a practical limit of about 25 kHz.

After toggling the push button to fast tuning mode, the operator can still tune in 1 Hz steps by turning the shaft very slowly. However, the acceleration algorithm ramps up much faster, so the maximum tuning rate is about 150 kHz per revolution of the encoder shaft.

When in fast-tuning mode, the right-most character on the first line of the LCD displays a "+" (plus sign) if fast-tuning acceleration is in use.

Push Button Row-3 Col-3: Band Down

Pressing this button will advance the frequency to the low end of the next lower amateur band. The bands are described in the section describing the BAND UP push button operation.

Push Button Row-3 Col-4: Backlight Toggle

This push button is used to toggle the LCD backlight on and off.

Push Button Row-3 Col-5: Calibrate

The CALIBRATE button is used to put the VFO into a special mode to ensure the DDS is generating a sine wave for the exact frequency that is displayed on the LCD.

When the CALIBRATE button is first pressed, the VFO switches to 10 MHz and the optical encoder tuning rate is set to a very small number. Now tune a nearby receiver to WWV at 10 MHz and listen, in CW mode, to WWV's tone at some offset (eg 700 Hz above). You can hear the IQPro putting out a faint signal at what it "thinks" is 10 MHz with the same tone offset. While listening to the WWV tone and the IQPro tone, turn the encoder to adjust the IQPro's tone. As the two frequencies get closer together, you can hear the beats get slower and slower. Listen for zero beat, which indicates the tones are identical. When the two tones are identical it is calibrated. Now press the CALIBRATE button again to exit this mode. The updated constant is now saved in EEPROM.

Other Controls

A keying circuit interface header is provided on the board for a connection to external keying circuitry. It is used for switching between the two VFO frequencies when in SPLIT mode. The LCD changes from RCV to XMIT when the keying interface pin is grounded (or the two header pins are logically connected together). In addition, if SPLIT mode is selected, the PIC changes to use the opposite VFO frequency when the keying interface pin is activated (grounded).

Test this feature by entering different frequencies in VFO-A and VFO-B. Select SPLIT operation. Now short the two header pins together and watch the display change from one VFO to the other and the display change from RCV to XMT. By design, the currently selected VFO is always the "receive" VFO, and the receive frequency is displayed. The other VFO contains the "transmit" frequency and it is displayed when transmitting.

Liquid Crystal Display

The LCD shows the current frequency that is being generated by the DDS. It also shows the state of the SPLIT toggle, the current transmit/receive mode, and the currently active VFO (A or B). When direct frequency entry is being entered via the push buttons, the digits are displayed on line 2.

A pair of resistors (R15 and R16), forming a voltage divider, are used to adjust the LCD brightness/contrast. These resistors are not necessary for all LCDs but are often required for proper contrast. For some types of LCDs, pin 3 may be tied directly to ground, while in other LCDs a small positive voltage is required on pin 3 for maximum brightness. The resistor divider has R15 (820 Ω) and R16 (5.6 kΩ) connected in series, with R16 connected to VDD (+5 V) and R16 connected to ground. pin 3 is connected to the junction of R15 and R16. Thus, the voltage at LCD pin 3 is calculated in from:

$$V = 5 \times \frac{820}{(5600+820)} = 0.64 \text{ V}$$

If you use an LCD other than the one supplied in the IQPro kit you may have to determine the correct bias voltage empirically.

If 0.64 V is not correct for the contrast control for your LCD, simply adjust the size of R15. R15 can be shorted out (replaced by a piece of wire) if 0 V is the appropriate voltage for pin 3. In this case you may leave out R16. Obviously you don't want to short out R16.

Virtually any 16 × 2 LCD that uses a HD44780 controller can be used. However, the backlight current requirements need to be considered. The current-limiting resistor (R68) in the backlight's power line is 6.2 Ω for the LCD supplied in the kit. If another LCD is used, this value may have to be adjusted.

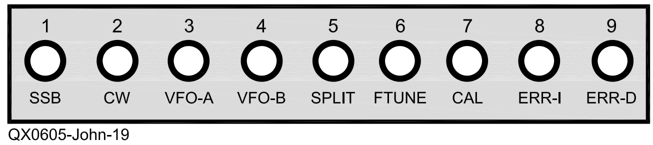


Figure 19 — The LED layout.

The LCD backlight is toggled on and off by a PIC pin driving a MOSFET. A push button (Row-3, Column-4) toggles a PIC pin that drives the gate of MOSFET Q9 through R69. When the PIC pin is set high, the Q9 is "turned on" so current flows through the MOSFET from drain to source. When the PIC pin is set low, Q9 is "turned off" so current flow is shut off. An IRLML2502 MOSFET was selected for this application because of its very low resistance (0.045 Ω) when the device is turned on, allowing the high current flow with the 5 V supply.

LEDs

There are no light emitting diodes (LEDs) on the Main Board. Instead, LED connections are made via Header 2 on the Main Board. This header has nine active pins (one per LED) plus a ground. Each header pin goes to +5v when the LED is turned on. Figure 19 illustrates the LED layout.

LED1 indicates that Single Sideband mode is active. LED2 indicates that CW mode is active. LED3 is turned on when VFO-A is active in receive mode. LED4 is turned on when VFO-B is active in receive mode.

LED5 is turned on when SPLIT operation is active. LED6 is turned on when Fast Tuning is active.

LED7 is turned on when Calibrate is active. LED8 is turned on when an error is detected in the Interface PIC.

LED9 is turned on when an error is detected in the Driver PIC.

IQPro Spectral Measurements

The question of spectral purity is always the subject of a lot of discussion in any forum in which DDS VFOs are discussed. This is a very legitimate question.

The IQPro has 30 MHz low-pass filters for the AD9854's I and Q signals. These filters are located between the toroids and the transistor amplifiers. The toroids or transistor amplifiers may be bypassed, depending on the amplifier option selected. See Section 3.6. The main purpose of these low-pass filters is to remove the DDS fold-back signals — signals that occur at the DDS reference clock frequency minus the fundamental frequency — that are produced by any DDS. The amplitude of these fold-back signals follow the Sin(x)/x curve. See the AD9854 specification sheet for more details.

Measurements of the IQPro's DDS outputs were made with a TEK 2712 spectrum analyzer. For these measurements, the toroids and transistor amplifiers were in use and the signals levels were set at the maximum.

Harmonics were seen in the output spectrum. When the DDS frequency is 10 MHz, the second harmonic of about –26 dBc was seen as well as a third harmonic of about –31 dBc. When the DDS frequency is set to 25 MHz, the second harmonic is about –27 dBc but the third harmonic is about –24 dBc.

Harmonics can be produced by nonlinearities anywhere in the RF chain so band-specific filtering on the DDS board would not be sufficient. For a receiver, harmonics at these levels should not be detrimental. For a transmitter, the output signal must be filtered by a band-specific low-pass filter at the transmitter's output so, once again, these harmonics should not be detrimental. Obviously, one low-pass filter cannot cover all of the HF ham bands. A low-pass filter that is used to remove the second harmonic from a 40 m (7 MHz) fundamental could not be engaged when running at 20 meters (14 MHz fundamental), and so forth.

I found that the spectrum was very clean with no noticeable spurs above the noise level until the frequency reached 125 MHz. Then the reference clock was seen (about –38 dBc).

The IQPro supplies signals at a header (HDR5) that facilitate the engagement of a band-specific low-pass filters at the transmitter's output, just before the signal is sent to the antenna. Low-pass filtering at this level will bring the transmitted spectrum to a level that is FCC compliant.

Questions/Support

For up-to-date details and documentation regarding this project, please see my Web page, **www.cbjohns.com/aa0zz**. The latest version of the code for the two PIC microcontrollers can also be found on the site. For additional support questions, see the YAHOO group DDS-VFO (**www.groups.yahoo.com/group/dds-vfo**). Latest pictures, assembly manual, PIC code (source and machine code) files are available on the YAHOO group and on my Web page. User-contributed files are also available in the FILES section of the YAHOO group.

Conclusion

Is a DDS-based VFO right for your radio? It really depends on your rig. There are trade-offs in any design and clearly, a DDS-based VFO is not right for every application. However, the ability of the AD9854 DDS to produce I and Q signals throughout the 0-30 MHz tuning range is intriguing.

It is my opinion that most homebrew projects are never completely done. By the time you finish one version you are tempted to do it again — to make it even better. The IQPro is one building block of a radio that can be customized forever. That's why the source code is distributed. As we all experiment with various configurations and share them with others, we learn together. That's the real value of building the IQPro DDS VFO.

Craig Johnson, AA∅ZZ, is an electrical engineer, having graduated from the University of Minnesota in 1971. He has worked at Unisys Corporation in Roseville, Minnesota, for his entire career (34 years and counting), working on the design/development of large-scale, mainframe computers. He got his first ham license in 1964 at the age of 14 and proceeded to General class license 6 months later. After college he let his license lapse (to concentrate on computers). In 1995, two of his three children expressed interest in ham radio so he told them he would teach them the theory and they could all get their licenses together. His daughter (then age 14) and son (then age 10) went with him for the license exams and they all got their Technician Plus licenses. After that he earned his Amateur Extra class license. Today, his entire family of five has ham licenses. □□

Dual Directional Wattmeters

An in-depth look at power on transmission lines, how it is measured using the dual directional wattmeter and the basic design principles of RF wattmeters.

Eric von Valtier, K8LV

Next to the transmitter itself, the RF wattmeter is probably the most ubiquitous accessory in Amateur Radio stations. We rely on it as an overall indicator of the health of our station. Wattmeters come in an assortment of designs and styles, all of which share a few basic characteristics. Despite their simplicity of use, they may add some confusion to the overall goal: accurate measurement of the power into the transmission line and then into the antenna itself.

In this article, I will explain the inner workings of the dual directional RF wattmeter (DDW). This will lead naturally to simple and direct answers to some of the common, perplexing questions that have arisen along with the evolution of the device itself. That evolution actually parallels the development of coaxial cable, which began in the 1940s. By the early 1950s coax began to appear in the off-the-shelf electronic markets, and it rapidly became the transmission line of choice for RF power.

Most of the mystery surrounding RF wattmeters is a result of the intricacies of wave propagation and power flow along transmission lines. Within this article we will focus specifically on the coaxial line, and I would like to begin with a very short review of some basic principles. This review will present a formulation of the main ideas of bidirectional power flow and its measurement.[1]

Impedance and Bidirectional Power Flow in Coaxial Lines

Two commonly asked questions are: "Why does coaxial cable have a specific impedance?" and "What is the significance of impedance matching?" The answer to the first question is direct and uncontroversial. The second is a little more difficult. When the basic electrical theory of a transmission line is analyzed, as in any of the typical references, it begins by assuming the line to be a linearly distributed arrangement of inductance, capacitance, and resistance — the three basic building blocks of electronics. It is further shown that the line can be described as having a unique value of L and C for every infinitesmal section, with values for L and C depending totally on the physical materials and dimensions of the line. Given this value of L and C for every section, a straightforward calculation of the voltage across, and the current through, that section shows that V and I are not independent of each other. For a given value of current, I, flowing in any section, the voltage across the section must be V, where V and I are related by the basic formula for a uniform lossless transmission line:

$$V = I \bullet \sqrt{\frac{L}{C}} \qquad \text{(Eq 1)}$$

Since this looks exactly like Ohm's Law for ac circuits, it is natural to identify the quantity $\sqrt{L/C}$ as an impedance. It is given a special name — *characteristic impedance* — and it directly controls the process of power flow through the line.

$$Z_0 = \sqrt{\frac{L}{C}} \qquad \text{(Eq 2)}$$

where Z_0 is the characteristic impedance.

In terms of circuit quantities, its interpretation is specific, as it states that if we wish to send a sine wave of power through the line to a load, there is a restrictive condition on that power flow, as expressed by Equation1. It requires that for every ampere of current through the line, we must apply 50 V of signal amplitude, assuming a 50-Ω line. This relationship between V and I is very tightly locked by the line impedance, which is a result of the microscopic wave physics of wave motion and power flow, and it potentially imposes a serious restriction. It literally forces all power flow to occur in 50-Ω waves on the line, implying the impossibility of driving a load of some other value; such as 95 Ω.

In the typical treatment of transmission-line theory, the discussion at this point would be punctuated by a statement, in a totally *de facto* manner, that the presence of a terminating load will generally result in reflection of some of the applied power. The succeeding analysis then proceeds to establish the effects of this line discontinuity, or mismatch, as we call it, on all aspects of the power flow in the line. It has always been unsettling to me to see this statement appearing as a given fact and used as the basis of the ensuing calculations and formulas. It would be far more satisfying, both practically and theoretically, to establish from the start why this reflection must occur and use that underlying physical reasoning to guide our intuition. I have formulated the following review of basic transmission-line theory that will fully provide that satisfaction. I hope that many readers will find their grasp of the dynamics of transmission lines enhanced and clarified. That, in turn, will facilitate the trip through the mysterious land of DDW technology.

The Need For Modal Wave Flow

If the line can *only* transmit power in waves that have an impedance of 50 Ω, what happens if we desire to use this line to transmit power to a 75 Ω load? Or, to a mismatched load like 45 + j20 Ω? On the basis of the previous reasoning, that would be impossible, as only 50-Ω waves seem to be allowed. What mysterious thing is going to take place when we connect a mismatched load

[1]Notes appear on page 31.

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to the end of the 50-Ω transmission line?

The answer to that lies in another important fact regarding wave flow, which is that every line can actually support two such waves of a given frequency traveling in opposite directions. One seemingly carries power in the forward direction, from source to load, and the other carries power in the reverse direction. The basis of this lies in the physics of wave propagation and the equations that express them. What the relevant equations actually show is that a wave can travel in either direction while at the same time satisfying the basic $V \bullet I$ relationship in Equation 1. More important is the fact that *both waves can exist simultaneously, independently of each other, and with independent phases*. The true field within an actual line is a mixture of both of these wave modes, which greatly affects the process of power flow on the line. This is the crucial fact that suppresses the potential limitations of Equation 1.

These waves are typically referred to as the forward and reverse waves. Their existence is the result of the completely bidirectional symmetry of the physical line, which is intuitively obvious. I will discuss the physical nature of the forward and reverse waves a little later, but for now their importance is as follows:

The impedance relationship of Equation 1 applies to each of these waves individually, but not to their sum. The way in which the two waves combine to form the total physical wave involves some complex arithmetic, which results from the fact that as these two waves move along the line in opposite directions; the relative phase between them is constantly changing. Hence, their sum and total impedance, at any point on the line, is also changing and takes on a wide range of values, varying above and below the characteristic impedance of the line. In other words, the impedance of the composite wave carrying the energy to the load gets *transformed* throughout a range of values, which differ in their position along the line. It is this fact alone that allows any particular coaxial line to be used as a transformer that can transmit power from a source to a load of any desired impedance. In other words, it allows efficient power transfer even when mismatched. The 50-Ω line can really be used to transmit power to loads of virtually any impedance, which happens often in the practical world. I refer to this process as impedance micro-matching, because it microscopically blends (matches) the two modal waves so that they match the correct composite $V \bullet I$ relationship forced at the termination by the load impedance. It occurs naturally when the load is *not* precisely matched to the line impedance by design, and forces a match of all power flow, at the microscopic level anyway.

This is not the conventional view of transmission line impedance matching, which concentrates typically on the side effects of mismatching. I am trying to illustrate a fact, not widely known and understood, that the bidirectional wave propagation actually is our friend, in that it allows us to use the coaxial line efficiently for feeding loads that are different from the characteristic impedance: mismatched lines. The only way that power can flow to a non-matched load is by the simultaneous presence of these two opposing modes. In theory, *any* load impedance can be fed through a given cable impedance. The two wave modes will always adjust themselves so that each is a 50-Ω wave but their sum will produce the applied load impedance at the load end. The wave at the driving end will have yet another value and is the value presented to the generator, commonly referred to as the driving-point impedance or the transformed impedance.

V = Line voltage
I = Line current

$$Z = \sqrt{\frac{L}{C}} \qquad \text{(Eq T1)}$$

$$\beta = \omega\sqrt{\frac{L}{C}} \qquad \text{(Eq T2)}$$

$$V_F = I_F \bullet Z \qquad \text{(Eq T3)}$$

$$I_F = \frac{V_F}{Z} \qquad \text{(Eq T4)}$$

$$V = V_F \; e^{j(\omega t - \beta x)} + V_R \; e^{j(\omega t + \beta x)} \qquad \text{(Eq T5)}$$

$$\Gamma = \frac{Z_{LOAD} - Z}{Z_{LOAD} + Z} \qquad \text{(Eq T6)}$$

$$V_F = \frac{V}{1+\Gamma} \qquad \text{(Eq T7)}$$

$$V_R = \frac{V \bullet \Gamma}{1+\Gamma} \qquad \text{(Eq T8)}$$

$$SWR = \frac{1+|\Gamma|}{1-|\Gamma|} \qquad \text{(Eq T9)}$$

$$SWR = \frac{2Z_{LOAD}}{Z_{LOAD} + Z} \qquad \text{(Eq T10)}$$

$$P_F = \frac{V_F^2}{Z} \qquad \text{(Eq T11)}$$

$$P_R = \frac{V_R^2}{Z} \qquad \text{(Eq T12)}$$

$$P_{TRANS} = V_F^2\left(1-\Gamma^2\right) \qquad \text{(Eq T13)}$$

$$P_{TRANS} = \frac{V^2\left(1-\Gamma\right)}{\left(1+\Gamma\right)} \qquad \text{(Eq T14)}$$

Figure 1 — General transmission line formulas.

The forward and reverse waves on the line combine to produce a pattern in space along the line that shows peaks and valleys. These occur at the points where the phases of the two waves, which change oppositely with direction of travel along the line, constructively and destructively reinforce each other. For every peak, there is a valley a quarter wavelength away, and the peaks repeat every half wavelength. This pattern is typically called a standing wave, because it *stands there* in space instead of propagating along the line. It does not represent true power flow, but is still subject to normal Ohmic loss on the line. That is the main reason that standing waves are undesirable: They produce loss that is parasitic to the desired power transmission. The ratio of the maximum to the minimum amplitudes of this pattern is directly related to the reflection coefficient, as given by Equation T6 (see Figure 1), and is called the standing wave ratio (SWR). It is a commonly used measurement of line impedance matching, but its significance is not as well understood.

The basic formulas of transmission line theory, as applied to the generic case of a generator driving a single load over a lossless cable, are summarized in Figure 1. The most important quantity, next to the load and line impedances, is the voltage reflection coefficient — typically symbolized as Γ. This quantity is uniquely determined once a given line is terminated.[2] See Equation T5. Various functions of Γ then describe every important wave relationship on that terminated line. Γ drops to 0 only for a perfectly matched line, –1.00 for a shorted line, and +1.00 for an open line.

Contrary to popular views, the forward and reverse waves on a transmission line are not separate fields. There is only one measurable electromagnetic field present in the line, assuming a single CW power source at the sending end. That is, if you were able to carefully probe the interior of the line under power, at every point there would be one specific value for E and one for H. Similarly, any electrical measurements on the line with ordinary voltmeters and ammeters would reveal single, well-defined values of voltage and current (these would typically be called the line voltage and line current). The process of analyzing the circuit using bidirectional waves is a theoretical concept, which turns out to be a very useful one. It is a simple case of what field theory refers to as a "modal expansion," in which an EM field is analyzed as a sum of modal waves. Each modal wave is chosen to satisfy the basic equations for the case at hand, which in this case is

Equation 1 and the formulas from Figure 1.

I wish to emphasize the fact that the forward and reverse waves really do not exist separately, and the idea of sending power into a line, only to have it reflected back, is not strictly true. It is mainly an analytical device that allows one to disobey the overly simplified view of line impedance by permitting power flow where 50 V does *not* produce 1 A of line current. The separation of the true field within the line into forward and reverse components can happen physically only by the intervention of some device which is able to resolve these two components and measure or demodulate them separately and independently. The directional wattmeter is one type of device that performs this function by electronic means.

From a practical standpoint, the user of an RF wattmeter is not usually interested in all of these details. He simply wants to know what the true power flow is from the transmitter into the line. The problem is that as usual, "It's not that simple." Specifically, the net power flow to the load is the algebraic sum of the forward and reverse wave powers, which is synonymous with the familiar rule, "Transmitted power equals forward power minus reflected power". It really is necessary to analyze the actual wave on the line into its forward and reverse components by some means, and measure each separately. Any scalar measurement is meaningless by itself. There are two generic methods of making such measurements, as I will explain.

RF Power Measurements: General Methods

The simplest method of transmission line power measurement is shown in Figure 2A. It is based upon the ability to make an accurate measurement of the vector impedance at the line input.

If the exact vector impedance is known (Z and Θ), the power flow is determined from any of the following formulas:

$$P = \frac{V^2 \bullet \cos(\Theta)}{Z} \quad \text{(Eq 3A)}$$

$$P = I^2 \bullet \cos(\Theta) \bullet Z \quad \text{(Eq 3B)}$$

$$P = \frac{R_{LOAD} \bullet |V|^2}{|Z|^2} \quad \text{(Eq 3C)}$$

$$P = R_{LOAD} \bullet |I|^2 \quad \text{(Eq 3D)}$$

where:

Z = impedance magnitude

Θ = impedance phase angle

The voltage or current is measured with an ordinary RF voltmeter or ammeter. During the earlier eras of ham radio, it was common practice to install an RF ammeter right at the transmitter output, and this reading was used in Equation 3D to calculate the feed line power using the known feed line impedance, according to that equation. For example, a 50-Ω matched feed line at 1 kW power output would read 4.49 A on the RF ammeter.

The disadvantage of this basic method is that it requires accurate measurement of the vector impedance, which is not so simple. Moreover, RF impedances are typically not constant except over very narrow bandwidths and it would be necessary to keep on hand a chart of impedance (Z, Θ) or ($R + jX$) at all frequencies of interest. Hence it is rarely used, in favor of a wattmeter of one of the types discussed here.

The second method, as shown in Figure 2B, is to use some means of separating the transmission line wave into its two modal components, and measuring the amplitude of each separately. This is the purpose of the DDW, and has become the common standard in ham stations since its origin over 40 years ago. The inner workings of the DDW will now be revealed in some detail. For this purpose it is natural to divide the discussion into two major areas: RF sensing and detection, and measurement. The majority of the complexity will be found in the first area and it will receive maximum attention. The second area is considerably simpler and requires much less explanation.

RF Sensing Methods

RF sensing requires that the field(s) within the line be sampled, which results in the extraction of sample signals that are related to the actual ones by known scale factors. For example, in one design to be described later, every watt of power in either forward or reverse wave will produce 1 milliwatt of sample signal across the internal dummy loads. The sampled field is assumed to be a perfect replica of the measured field with a known attenuation factor (30 dB in this case). The sensing circuitry produces both the forward and reverse samples, whose amplitudes can then be measured separately by simple (scalar, not vector) voltage measurements. Once the sensing has been done and the RF samples are available, the design domain changes from RF waves to ordinary low-frequency ac/dc circuitry.

From a rigorous theoretical standpoint, the devices used by hams for power measurements are not really wattmeters, and never have been. A true wattmeter must measure the time-averaged value of instantaneous voltage times instantaneous current, where the average is typically taken over many RF cycles. In practice this requires a circuit that is capable of performing multiplication (of instantaneous voltage times current) and subsequent averaging. Although possible to build, such circuits are quite exotic and difficult to design for wideband use. Hence, they are rarely used in practical RF power measuring equipment. A simpler alternative is to measure the rectified RMS values of V or I at the sample load, using

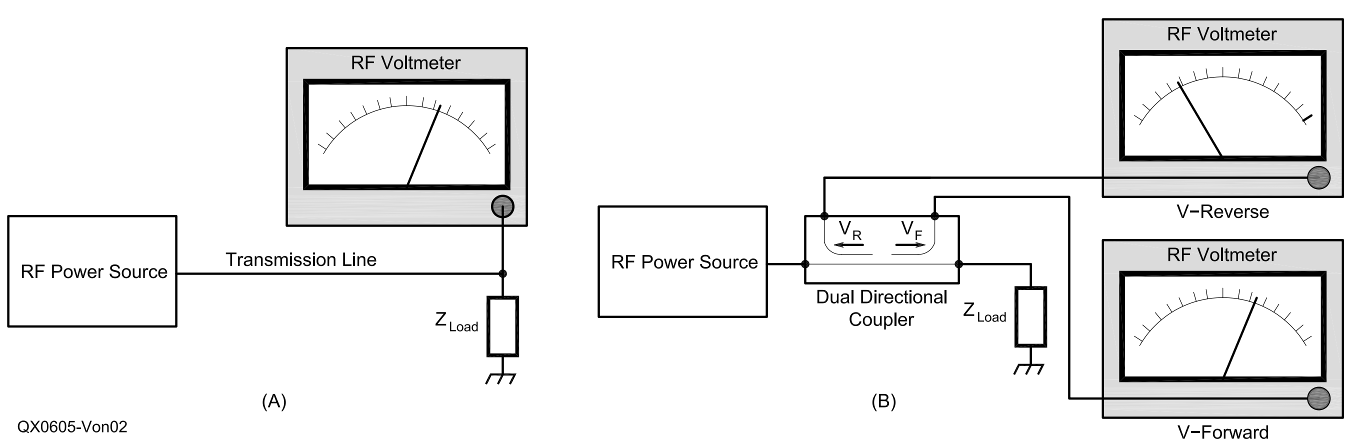

Figure 2 — Part A shows scalar power measurement. Part B shows dual directional power measurement.

a precision resistor dummy load that validates the following equations for the sampled power:

$P = I^2 R$ (Eq 4)

or

$P = V^2 / R$ (Eq 5)

These equations are only valid for steady-state CW signals. For example, a few percent of second harmonic RF component (not second-order IMD!) in that signal can result in a few percent error in the calculated power. Practically speaking, though, the assumptions required for the validity of Equations 4 and 5 are well satisfied in Amateur Radio RF applications, and these equations are used ubiquitously in virtually every DDW design that is marketed to hams. Without this simplification, the world of wattmeter design would involve one more level of complexity. Since Equations 4 and 5 are only valid for a CW signal in steady state, SSB and rapidly varying signals require additional factors to be considered in determining what is actually being measured. For the present we will assume the validity of these equations, and further discuss the intricacies of the actual measurement.

There are three common design themes for DDWs. The circuit analysis of each is summarized in Appendix A, with an emphasis on illustrating what quantities are actually being measured. For all of these designs the sample signals bear exact relationships to the desired quantities of measurement and that relationship is stated by means of the given formulas.

Typical DDW Sensing Circuitry

The analysis of each of the circuits was based upon the formulas accompanying each schematic, which may in some cases involve simplifications to provide clarity. Such approximations allow calculations that will be accurate to within 1 to 2%, which is adequate for common power measurements. More accurate calculations are readily available, which account for second-order and parasitic effects. Some practical notes on these designs follow.

The "Bruene circuit," named after the innovative Collins engineer and radio ham, is used in just about every SSB transceiver ever produced for hams. See *QST*, April 1959, "An Inside Picture of Directional Wattmeters," W Bruene, WØTTK. It is also used in most of the various wattmeters marketed by the popular manufacturers of ham gear. It is simple in design and construction, and quite easy to calibrate, which accounts for its commercial success. It is sometimes packaged alternatively as an SWR meter with a single meter, which requires the user to manually adjust the meter for full scale using a control on the front panel. The scale can then be calibrated directly in SWR units.

The hybrid-coupler design is potentially more accurate over a large bandwidth than others. Its scale factor is set by the transformer ratio, N, which is a precise quantity. It is somewhat sensitive to capacitive coupling in the current-sense transformer, which can be minimized by good physical design. I favor this design and find it to easy to build for use up to 100 MHz.

The unidirectional VI design has been used by several famous manufacturers of high-grade wattmeters. Its disadvantages are narrow bandwidth, narrow dynamic range, and extreme sensitivity to harmonic distortion. It is also an inherently unidirectional design and reverse wave measurements are made by turning the entire sensing circuit around 180° by mechanical means. This ingenious feature allows use of a single meter for a DDW, but will not allow remote applications. That would require two separate sensing heads, one for each direction.

A complete and rigorous analysis of these circuits is too complex to present here in its entirety. A number of useful approximations have been made for clarity. These approximations produce very little error in the final formulas while greatly simplifying the detailed circuit analysis used.

1) The DDW is assumed to be small enough that it does not physically modify the fields on and within the transmission line being measured.

2) All transmission line effects within the DDW itself are neglected and it is treated as a pure lumped-element circuit with one input node and three output nodes.

3) The insertion loss of the DDW itself is negligible.

4) The load is assumed to be resistive.

With regard to item 3, note that all practical designs produce some small loss, primarily due to voltage drop across the current sensing element. This loss is lowest for the single directional design, which may be regarded as the most precise from that standpoint. The current transformer designs present an equivalent series resistance from input to output of approximately R_s / N^2 (see Equation A2). For typical designs this amounts to a small fraction of an ohm and produces less than 1% error. Item 4 produces a great simplification in the circuit analysis while still showing the basic calculations. The final results regarding power flow are all still valid for complex loads, which interested readers may verify for themselves by repeating the calculations using complex phasors for all circuit quantities.

RF Power Measurements: The Dual-Directional Wave Approach

A complete DDW performs the following distinct functions:

1) Extract small samples of the EM wave flowing through the line.

2) Separate this field into its forward and reverse waves, each applied to a dummy load.

3) Measure the amplitude (i.e. voltage) of each wave modal field.

4) Convert these voltage measurements into power levels and display them.

As shown in the Appendix, the first two items are performed by a piece of circuitry that is generally referred to as a "directional coupler." In typical designs, it produces two demodulated output voltages, equal to the amplitude of the forward and reverse waves, times a scale factor that is common for both. It will now be shown how measurements of the forward and reverse waves, using steps 1, 2 and 3 in the list, are used to produce the final readings.

The design formulas show that the net

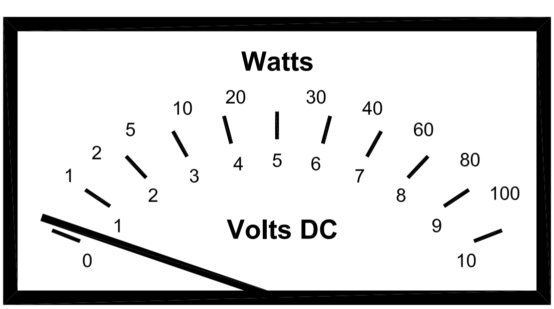

Figure 3 — Voltmeter calibrated in watts.

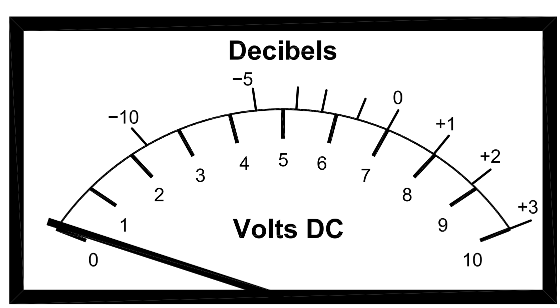

Figure 4 — Voltmeter calibrated in decibels.

power flowing to the load gives rise to the sample signals v_1 and v_2, proportional to the forward and reverse waves. To relate this to power, it is necessary to further manipulate these values by squaring the directional formulas (A5) and (A6) or (B1) and (B2), since the power is proportional to v^2. This results in the important formulas (D5) and (D7). They establish a rigorous proof of the well-known rule applied to all DDW meters: transmitted power is equal to P_{FWD} minus P_{REV}. The formulas show clearly why that is true and how it forms the basis for the DDW.

It is convenient to regard the sensing circuitry as producing signals with internal scale factors of watts-per-square volt. These are then applied to ordinary dc voltmeters, which are calibrated in units of volts squared, instead of linear volts, times some scale factor. For example, if the internal scale factor is 1.0 V^2/W, 25 W would produce a forward power signal of v_1 = 5 V. The meter scale would be calibrated so that the 5 V mark would be labeled as "25 W", 10 V as "100 W," and so on. This is shown in Figure 3.

Another popular style of meter calibration and labeling is shown in Figure 4. The reading is converted directly into dB referred to some standard load. One popular standard is the "decibel-milliwatt," which establishes 0 dBm as one milliwatt into 50 Ω. This corresponds to a voltage of 0.223 V_{RMS} across a 50 Ω load. Many lab instruments use meters that are scaled so that the 3 dBm point (0.316 V) is full scale. This puts the 0 dBm point at 0.707 of full scale. The resulting meter then has a usable range of about –15 to +3 dBm.

Demodulation and Detection

It is typically not effective to attempt to directly measure the RMS value of RF voltages. The earliest approach to making such measurements was to measure the heating effect of the signal on a special sample. The primary transducer was called a bolometer and consisted of an element made of some high resistance material, thermally connected to a temperature sensor. It is especially useful for making accurate RMS measurements.

A desirable alternative is to demodulate the RF signal, which for a CW signal source produces a dc output voltage proportional to the signal amplitude. The most common way of doing this is by rectification, using a small diode capable of properly switching on and off at high frequencies. In typical DDW designs the demodulation process takes one of two forms: ideal rectification, and square-law detection.

The ideal rectifier simply chops off the negative or positive half of the signal. The resulting rectified wave is fed into an RC filter with a relatively long-discharge time constant equaling many RF cycles. This in effect forms a peak-detector, which produces an output dc signal equal to the peak value of the RF signal. The RMS value is then inferred by including a factor of 0.88, the ratio of average-to-RMS values for a sine wave, into the meter calibration.

As described previously, the meter is calibrated to actually read the square of the demodulated voltages, as required by Equations 4 and 5, to produce power readings. This always results in a non-linear meter scale with 25% power (50% voltage) at center scale. The main source of error in this demodulator is the forward voltage drop of the rectifier, which causes increasing error as the signal voltage decreases. The best diodes from this standpoint are the older germanium types, which have quite low forward drops compared to modern silicon diodes. A certain amount of the diode-drop error can be decreased through an intentional distortion of the meter scale at its low end. As the RF peak voltage decreases to the range of the diode voltage (a few hundred millivolts) it simply becomes useless. For this reason most wattmeters have a low-power threshold, below which they are not accurate.

The one main advantage of the simple rectifier is its wide range of voltage inputs. This is beneficial in designs that are intended to cover as wide a range of power levels as possible. Modern Schottky diodes can tolerate up to 30 or 40 V of RF input while producing forward drops of only a few hundred millivolts. Hence, they have become quite popular in modern DDW designs covering power ranges of a few watts up to several kilowatts.

The square-law detector, on the other hand, uses the fact that for low values of forward voltage, most diodes exhibit a quadratic relationship between I and V. This fact can be used to produce a simple squaring circuit, which converts the input voltage directly to a voltage level that represents the voltage-squared times a scale factor, unique to each individual diode type. This factor is absorbed into the overall meter calibration factor, resulting in a meter reading of power that is close to linear. High-quality designs then add a small amount of additional scale "warping" to the meter face to make it appear linear over as wide a range as possible.

The disadvantage of this method is that it has very little dynamic range, typically around one full decade of power. As it goes beyond the upper limit of the power, the forward voltage applied to the diode starts to approach saturation and the quadratic I-versus-V curve of the diode slope saturates. This results in the need to limit the power range of the meter to much less than two decades, resulting in a design where the sensing elements, including all of the RF sensing components, are plug-in modules targeted for specific power and frequency ranges. For example, to change from 100 W full scale to 500 W full scale may require the user to change the sensing "slug," as they are called.

Filtering: CW and PEP Readings

Historically, DDWs all used mechanical meter movements for power display, since LEDs and LCDs had not yet been invented. With the advancement of electronic technology came the ability to provide much more sophisticated displays, and DDW products continue to evolve. In particular, the mechanical meter places some severe limitations on its ability to make measurements of rapidly varying sources, such as voice signals. Without electronic enhancements, such meters are nearly useless for accurate power measurement. The enhancement is realized by additional filtering of the demodulated sample signals. This is commonly referred to as post-detection filtering and we will look at several generic types.

The ideal wattmeter design would use a perfect RMS demodulator of the signal. This means that it would measure the true RMS value of the waveform at all signal levels, but such a demodulator is far more complex than desirable for this application. What is typically used instead is a basic rectifier and filter, which attempts to measure the RMS value by indirect means. The next best choice would be a circuit that measures the average value of signal. The RMS value would then be derived from this by the well known factor relating the RMS-to-average ratio of a sine wave. This known ratio is dependent upon the purity of the sine wave signal, which is usually adequate. It does require the use of a

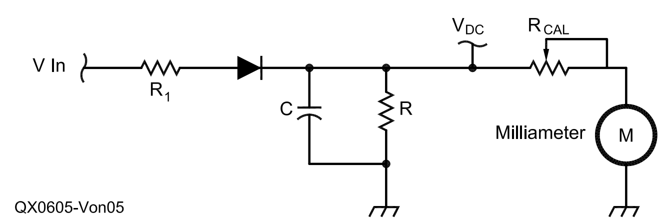

Figure 5 — Typical rectifier and meter driver circuit.

rectifier, however, as only one half of the sine wave can be averaged, since the full-cycle average is always zero! This produces the need for a rectifier, with its inherent error at low values due to forward drop. There are additional problems with half-cycle averaging due to other non-ideal diode characteristics, but they are too complex to deal with here. So, we will concentrate on the next best detector for this purpose: the peak detector.

It turns out that the simple peak detector is the best compromise between accuracy and complexity, and it depends on the known ratio between peak and RMS values of a sine wave. Assuming low distortion in the signal, the output of a peak detector will always be 1.414 times the RMS value (for sine waves), and the peak detector output can then be used to directly drive a display. The peak-to-RMS factor is simply absorbed into the overall calibration factor. This method is used in almost every DDW I have ever seen, and we will take a more detailed look at it.

The generic diode rectifier/demodulator is shown in Figure 5. The characteristics of this circuit are mainly determined by the relative values of R_1, R_2, and C.

For fastest peak response, the value of R_1 must be minimal and it is usually omitted. In conjunction with C it tends to low-pass filter the input signal, which is not desirable for fast RF peak response. Unfortunately, the diode parasitic RLC elements, which are all significant in real diodes, prevent it from actually performing well as an averaging detector (i.e., integrator) over a wide bandwidth. It just cannot be readily optimized to form into a useful averaging detector.

Therefore, R_1 is eliminated and a basic peak detector results with C charging rapidly to close to the peak value of the positive half-cycle, minus the diode drop. In the typical circuit, the available charging current is quite high so highly efficient peak detection readily occurs. The discharge time is set by R_2, with typical values producing "hold" times of hundreds to thousands of RF cycles.

The minimal filter required for both meter channels is a low-pass RC filter, which removes all residual RF variation and leaves a dc voltage equal to the peak amplitude of the RF signal. If the RC time constant is a moderate multiple of the RF period, it will closely resemble the peak value of the RF voltage over a short time. The voltages from each channel of the DDW front end can next be applied directly to meters or electronic voltmeters for final display. The latter consists of a fixed resistor in series with a low current milliameter, with part of the resistor being made variable. At this point in the circuit, all of the various scale and proportionality factors that have entered into the overall meter reading, for a given reference power level, can be reduced to this one adjustment, which forms the final calibration variable. In practice a known power level (e.g. 200 W) is fed thru the DDW to a matched load and then the calibration pot is adjusted for a reading of 200 W forward. A known mismatch is then applied and the reverse power calibration pot is set to produce a reading that satisfies Equation 6:

$$P_{REF} = P_{FWD}\, \Gamma^2 \qquad \text{(Eq 6)}$$

Power Display for Rapidly Varying Signals

There is an ongoing debate about the display of transmitter output during SSB and some types of data modulation. The basic problem here is that there is no meaningful measure of power that meets everyone's needs. The actual averaging interval is very influential on the result but there is no basic time interval that stands out as ideal. It is natural to seek maximum power and energy flow on very general grounds, for which numerous methods of speech processing have evolved. I do not have any significant information to add to the debate except for the following.

One hard limit that must be faced by any transmitter operator is the clipping point. Above this point the amplifier begins to generate copious amounts of distortion, which is illegal and highly antisocial. Hence, it would be best if the amplifier power were held just below the clipping point. This is precisely what is done in well designed transmitters using ALC. A well designed ALC circuit will not allow significant excursions into the distortion region. Under ALC control, every SSB operator knows that the power meters read only a fraction of the available PEP of the transmitter, leading the operator to believe that he is being cheated out of his full power. The natural reaction is to try to drive the audio harder and push the meter reading up, which a properly designed transmitter will not allow. The best solution to this problem, which is basically psychological, is to allow the wattmeters to respond to the peaks of the RF envelope, which undoubtedly reach PEP frequently.

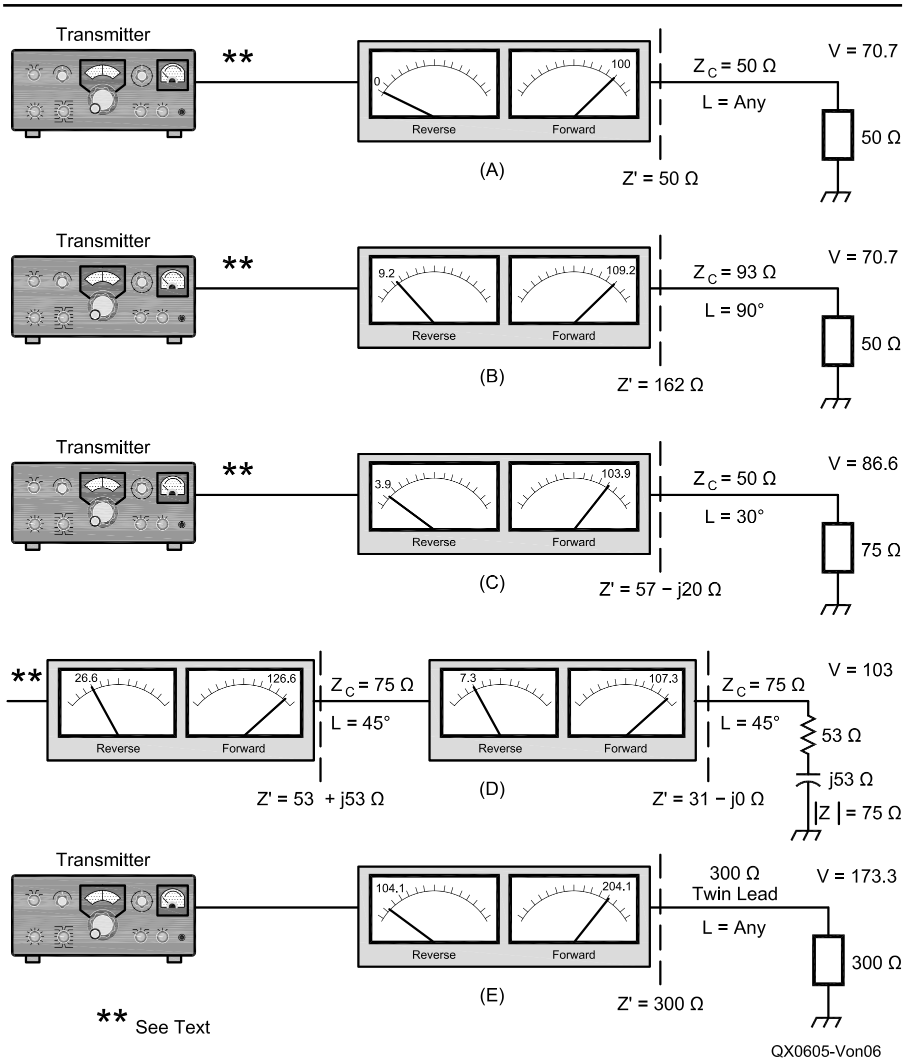

Figure 6 — Typical readings by a 50 Ω DDW for various conditions of line and load. All cases shown produce 100 W of real power into the load.

In terms of DDW technology this is a simple matter to resolve. As shown, these instruments are inherently peak reading in nature and their rectifier/demodulator circuits readily produce voltages that accurately indicate the RF power averaged over a moderate number of cycles. All that is needed is to increase the discharge time constant of the RC filter as shown in Figure 5. This may not be feasible in some cases because the resulting value of C becomes too large to properly peak-detect the RF signal.[3] This can be circumvented by employing a second peak detector circuit, which is powered by a power supply instead of the RF sample signal. Either way, allowing the operator to monitor the PEP more directly is preferable to producing the incorrect illusion that full power output is not being achieved. As the debates on proper or desirable averaging times and algorithms continue to evolve, my preference is to rely on basic peak detection until a more meaningful power measure has proven itself.

DDW Impedance and What It Really Means

As shown at the beginning of this article, the concept of directional, modal waves was a useful, but not necessary, device to explain the process of power flow on mismatched lines. If all lines were perfectly matched, power measurements would be reduced to simply measuring the line voltage and using Equation 5. A wattmeter would just be an RF voltmeter calibrated in units of V^2 / Z_0. But the presence of mismatch naturally leads to standing waves, which cause the line impedance to vary along its length in a complex way. The resulting total power flow, which must still remain constant all along the line, because of conservation of energy, can be effectively analyzed by the dual wave theory. The theoretical analysis shown here proves that these modal waves can be easily measured and the results used to accurately determine the power flow. I would now like to answer two very central questions in DDW theory and practice that have troubled many users, including me: What does it really mean for a DDW to have an impedance value? and What is the relationship of this impedance to the measurements it makes?

I will answer those questions in two different, but related, ways. First, the DDW impedance, Z_0, is actually an arbitrary value that was incorporated into the design, as shown by the steps related to Equations A3 or B1.[4] The only tangible effect is to cause a null in the reflected power reading when measuring power on a matched line whose impedance is equal to Z_0. The device will still make accurate power measurements on other transmission lines, matched or unmatched. To see that this is true, just consider that any line with any load will present a specific Z_{LOAD} to the DDW output port. Power will flow into this load completely unaware of the presence or effect of the DDW. That is, the DDW *cannot* force its own impedance, Z_0, onto that load in any way. Equation D4 proves that the power flow to any load will be accurately measured as the difference between the two meter readings, and Z_0 was simply a number that went into the calibration of the meters.

When the DDW is used on a line equal to its calibration impedance, if the load is matched, the reflected power nulls out and the forward meter reads the exact transmitted power, also called the *true power*. The best that can be said is that on a matched line of Z_0 impedance, the forward power meter becomes a true power meter. See Figure 6A. The fact that the DDW usually has SO-239 connectors for input and output does *not* mean that it cannot be used for non 50-Ω lines.

Second, consider what is happening on a mismatched line. The line impedance (V / I) starts out at the load end as equal to Z_{LOAD}, but as the line is traversed back towards the generator, Z_{LINE} varies greatly, repeating itself every half wavelength. This impedance variation is described by the common equations from transmission line theory, such as T5, and can be easily visualized on a Smith chart. A DDW inserted at any point along that line will read the correct power flow, regardless of the actual impedance of the cable as the examples in Figure 6 show.

The action of the DDW may be visualized as resulting from the insertion of a very short piece of cable with impedance Z_0 into the actual transmission line. This will cause an immediate mismatch on the line based upon the resulting value of Γ; but a little circuit analysis, omitted here, shows that an inverse transformation happens at the input side, restoring the line to its original impedance. In the limit of an infinitesimally short DDW, the overall line and load remains undisturbed but the forward and reverse measurements are made at an impedance of Z_0, the calibration impedance. That is why it will produce the correct values regardless of the line or cable impedance, by resolving the actual wave on the line into two 50-Ω modal components. Its internal calibration is based upon 50-Ω waves.

Measurements on Mismatched Lines

Practical lines are always mismatched to some degree. The question arises as to what effect this has on DDW readings. One purpose of this article is to answer that specific question. So far, it has been shown how the DDW is able to produce two voltages (V_F and V_R), which are each equal to a scale factor times the amplitudes of the directional waves. It has also been shown how these voltages are used to establish the transmitted power, which is the primary goal. A finished DDW consists of this voltage-sensing section, whose outputs are rectified, filtered and applied to meters calibrated according to Equations D5 or D7.

Another useful purpose it can serve is to analyze the closeness of the match between the line and the load, which is usually an antenna. The basic quantities which describe matching error are the reflection coefficient Γ, which is defined in Equation T5, and the SWR defined in Equation T8. These quantities are derived from the DDW readings by more application of the external analog computer — you! In particular, the value of Γ is determined by the following ratio of two readings:

$$|\Gamma| = \sqrt{\frac{P_{REF}}{P_{FWRD}}} \qquad \text{(Eq 7)}$$

From this, the value of SWR is readily determined by applying Equation T8. It is interesting to note that this formula shows that the DDW is indirectly making a measurement of Γ and for this reason it is sometimes called a "reflectometer."

It has been erroneously claimed that a DDW of the types described here is capable of making vector impedance measurements. (See the references in Note 1.) This is a result of overly ambitious application of the concepts of directionality and vectors to the situation. In fact, the DDW performs the equivalent of a scalar impedance measurement of any loads. A basic description of the process, from this viewpoint, is that the load is measured in terms of its SWR relative to Z_0. Z_0 is the known calibration impedance of the DDW and SWR is obtained from application of Equation T6. This localizes the load impedance to all points on the SWR circle of radius S on a Smith chart centered on Z_0. For small SWR values, this is a narrow range of values, but at larger SWR values it can cover quite a large range of impedances.

Vector impedance measurements, especially at VHF and microwave frequencies, are frequently made using a directional coupler *and* a vector voltmeter. The latter is a sophisticated instrument that is considerably more complex than a DDW.

Some modern DDWs contain microprocessors capable of doing the kind of arithmetic on v_1 and v_2 implicit in the preceding discussion, so that they are able to directly display digital values of SWR and Γ, based upon their measurements of those two demodulated signals. Future DDW designs and products will certainly evolve toward more advanced displays, with this additional information directly displayed. For the users of many of thousands of DDW boxes already in existence, with just the two generic power meters available, these rules will continue to be of value for reference.

Some Interesting DDW Measurements

The following examples show a DDW connected at various places in typical transmission lines. They all were chosen to examine the power flow on a transmission line delivering 100 W into the associated load . These examples intend to portray the actual measurements that will result when the DDW is inserted anywhere along the line, since the portion of line between the DDW and the transmitter has no effect on the readings. Of course, the input cable can and will affect the actual impedance presented to the transmitter, possibly quite significantly. It is assumed that this additional mismatch is compensated for by readjusting the transmitter until the load power equals the reference value of 100 W. This may seem to be a trick at first glance, but it is perfectly valid. We are attempting to show that the DDW always reads the correct transmitted power as the difference between P_{FWD} and P_{REV}, its two meter readings.

The examples were selected to demonstrate different types and degrees of mismatch, and would probably have significant effects on transmitter power output. That is a shortcoming of the transmitter in its ability to tolerate a wide range of output loads. These results show that although this shortcoming may actually be in control of the power ultimately reaching the antenna, the DDW will still faithfully monitor this power flow and tell the user the true power flow.

The fourth example was selected to give a glimpse of what you might see as a DDW is moved along the given line. In each example, the value Z' is the line impedance at the DDW insertion point. It can be calculated in several ways, the easiest being to use a Smith chart for impedance Z_C to transform the load impedance from its terminal position back to the DDW by the number of degrees shown. Alternatively, the transformations can be calculated easily due to the choice of angles used, using the following equation, where all impedances are normalized to Z_C:

$$Z' = \frac{Z'_{LOAD} + j\tan\Theta}{1 + j\ Z'_{LOAD}\tan\Theta} \qquad \text{(Eq 8)}$$

where

$$Z'_{LOAD} + = \frac{Z_{LOAD}}{Z_0} \qquad \text{(Eq 9)}$$

The last of the examples might not be practical as shown, as the balanced line could interact with the DDW because it has considerable external fields, especially at VHF. This could be readily alleviated by inserting isolation chokes or a current balun at the interface. The example is intended mainly to reinforce the notion that the DDW is capable of measuring the true power flow into any load impedance whether it be a lumped load or a transmission line.

Final Comments

This article was the result of my study of the subject over many years of Amateur Radio experience. I can clearly recall the first *QST* article that was published on the subject, complete with a front-cover photograph, using the early design that had a 6-inch piece of transmission line for the current-sense transformer. I can also recall being quite mystified by how it worked — a situation that I endured for many years. I have seen very little published material that offered clear and complete explanations as I have tried to present here.

I recently ventured into the area of autotune amplifiers and antenna tuners, which is fertile ground for hams, such as myself, who are seriously interested in the hardware side of electronics. One issue that emerges immediately in the auto-tune world is the measurement of power, phase angle, and related quantities to RF signals. While pursuing this area, I happened to acquire a Collins 180L3 automatic antenna tuner. This proved to be quite an interesting device, sought after by some collectors primarily as a source for a very high-quality vacuum variable capacitor in its RF section.

As I dissected the circuitry of this product, it amazed me with examples of primitive techniques of both analog computation and digital circuit control (via relays!). In particular, there appeared an early version of the famed Bruene directional coupler, used in the design to decide whether the load reactance is positive, negative, or zero. Upon analyzing this circuit, my years of mystery about the DDW quickly disappeared and was replaced by a great interest in the technology of impedance and power measurement. That interest led to a thorough study of the subject and this article is one of the results of that study. The main result, however, was the ensuing assortment of hardware designs for DDWs, impedance meters, and auto-tuners, and the construction projects that resulted. Some of these may be subjects of future articles in this magazine. I would like to conclude with some observations that were made possible only by a thorough study of the DDW as has been presented here.

One source of confusion burdened me for many years, as I am sure it burdens most users. It results from the fact that every discussion of the DDW that I ever encountered used a bidirectional wave analysis, which is a natural and intuitive way to proceed for power flow on transmission lines. But we also know that we can easily connect any kind of a load on the DDW output port and get a correct reading of the power, as proven here. For this case (such as a simple resistor of any value) there is really no apparent wave motion. The methods of ac circuit analysis do not generally use wave fields, but rather pointwise voltages and currents. So with no waves in sight, what is the meaning of the traditional DDW analysis? As I have shown, the mere presence of a 50 Ω coaxial connector on the output port really has no significance in the measurement theory.

There is an interesting answer to this apparent contradiction. It relies upon a somewhat conjured-up definition of forward and reverse waves within lumped ac circuits and my first encounter with it did not result in enhancement of my overall grasp. Reviewing the historical development of the DDW, and the corresponding development of the theory in various articles (such as this one!), my opinion is that the DDW is a simple and useful design for measuring RF power, but the directional wave approach is really somewhat of a compromise. Like many similar cases, it was brought about by the needs and limitations of the technological era in which it was born. I question its usefulness for the future, however.

The dirty little secret of the DDW is that formally speaking, it does not do the job. It measures two quantities, v_1 and v_2, and then forces the user to input them to the external analog computer — himself! He must perform the calculation of $kv_1^2 - kv_2^2$ to obtain the final value of transmitted power. An apt analogy would be an ohmmeter, which really measured only the current through the unknown resistor after applying a fixed voltage to it. The user would then be given this current value and told to obtain the resistance by using his "analog computer" for solving Ohm's Law. In the days before the ohmmeter was born, precisely that was done.

When the DDW reports the two numbers P_{FWD} and P_{REV} to its user, these numbers are somewhat arbitrary. They relate the actual, measured power to two waves on a fictitious transmission line, one with impedance Z_0. These numbers do not have an absolute significance, and the concepts of forward and reverse power have been conjured up to simplify things for the user. In fact, the only absolutely correct quantity is the difference (See Equation D6) and there is not really a need to even label the meters *forward* and *reverse.* These labels are true only if the characteristic impedance of the line being measured is exactly Z_0.

The essence of power measurement deeply involves both phase and amplitude measurements, and the DDW only partially achieves this. As was shown earlier, it locates the impedance to some point on the circle whose radius equals S on the Smith chart. This is adequate for accurately measuring the true power, but not for other types of measurements, especially load angle or power factor. Technology has advanced enormously since the days when Mr. Bruene and his team had to devise ways of measuring load power and phase using a few LCR components and a mechanical meter movement. This newer technology will soon begin to appear in our wattmeters and these discussions of forward and reflected waves will disappear. The wattmeter will become what we really wanted it to be all along: an instrument that directly reads power flow to any load, and for a nominal additional fee, the phase angle of that load.

Notes

[1]Some excellent references on transmission lines for beginning, intermediate, and advanced levels are:

J. D. Kraus and K. R. Carver, Electromagnetics, McGraw-Hill Book Co, New York, 1973.

S. Ramo, J. R. Whinnery, and T Van Duzer, Fields and Waves in Communications Electronics, John Wiley and Sons, New York, 1967.

J. R. Colin, Foundations for Microwave Engineering, McGraw-Hill Book Co, New York, 1974.

[2]Throughout this article there will be minimal use of complex values for quantities such as Z and Γ, to keep the explanations as simple as possible. All of the power calculations keep their same basic form when done with full phasor arithmetic, which an inquiring reader can readily verify, using the non-complex form r of the reflection coefficient in place of Γ. The main intent here is to outline the methods of analysis that lead to the desired results for power flow.

[3]The accuracy of the peak detector is influenced by the value of R1, which includes the internal impedance of the sensing circuitry. So even with no R_1 there is some current limiting by the sensing transformer and capacitors C_1 and C_2. In addition, larger values of C result in larger charging current which increase the diode drop and its resulting error.

[4]The quantity Z_0 becomes absorbed into the overall calibration factor along with all of the circuit components. It is no more or less relevant to DDW operation than any of these and its uniqueness as a DDW parameter is limited to specifying that one unique value of Z_{LOAD} for which C × v^2 will null out.

Appendix

The following paragraphs present a straightforward ac analysis of three popular DDW designs. I have assumed for this analysis that the physical circuitry meets the assumptions stated in the text. A careful reading of these analyses will reveal important similarities in their basic operation, in that they all involve methods of sensing and measuring both voltages and currents and then combining these readings to display values of power flow.

The Bruene Wattmeter

The current sensing transformer is typically wound on a toroidal core for HF, but frequently uses directly coupled short lengths of transmission line at VHF. The transformer develops a sample voltage, V_S, as shown in Figure A1, as a result of the net load current ($I_f - I_r$). The capacitive voltage divider develops a sample of the line voltage, which is applied in series with the current transformer secondary. The resulting signals, v_1 and v_2, form the two RF output points of the circuit, with measurable voltages as follows:

$$v_1 = V_F\left(\mathrm{k}_1 + \frac{\mathrm{k}_2}{2\,Z_0}\right) + V_R\left(\mathrm{k}_1 - \frac{\mathrm{k}_2}{2\,Z_0}\right) \qquad \text{(Eq A1)}$$

$$v_2 = V_F\left(\mathrm{k}_1 - \frac{\mathrm{k}_2}{2\,Z_0}\right) - V_R\left(\mathrm{k}_1 + \frac{\mathrm{k}_2}{2\,Z_0}\right) \qquad \text{(Eq A2)}$$

The impedance calibration of the circuit is accomplished by requiring that v_2 nulls out for a perfect match at the calibration impedance Z_0. Equation A2 shows this to be possible only if the following calibration condition is imposed, by properly "adjusting" either k_1 or k_2:

$$\mathrm{k}_1 = \frac{\mathrm{k}_2}{2 Z_0} \qquad \text{(Eq A3)}$$

which, in terms of circuit parameters, is as follows:

$$\frac{C_2}{C_1 + C_2} = \frac{R_s}{2 N Z_0} \qquad \text{(Eq A4)}$$

From a practical viewpoint, this is how the DDW acquires its calibration impedance. It allows some degree of freedom in choosing the values N, R_S, and the capacitor values, but this detailed tradeoff will not be discussed here. By imposing condition expressed in Equation A3 on the basic network functions given by Equations A1 and A2, the formulas for the output voltages reduce to the final formulas:

$$v_1 = 2\,\mathrm{k}_1\,V_F \qquad \text{(Eq A5)}$$

$$v_2 = 2\,\mathrm{k}_1\,V_R \qquad \text{(Eq A6)}$$

These are the complete formulas for the output of the RF sensing section of the circuit. What they show is that the sampled signals exactly represent the forward and reverse wave amplitudes on the transmission line. The wave has been decomposed (resolved, in scientific terminology) into its two modal components. The text explains how these low-level ac signals are processed to produce the final readout in watts, S, and G. The hybrid-coupler will next be reviewed and seen to function very similarly.

The Hybrid-Coupler Wattmeter

The circuit shown in Figure A2 is an elegant method of separating the sampled line field into forward and reverse components, using a method originating from early telephone technology. It uses two identical transformers connected as shown, which can be analyzed similarly to the Bruene circuit. The two basic circuit equations for the two unknown nodes are:

$$v_1 - v_2 = \frac{V_F + V_R}{N}$$

$$v_1 + v_2 = \frac{R\,\left(I_F - I_R\right)}{N}$$

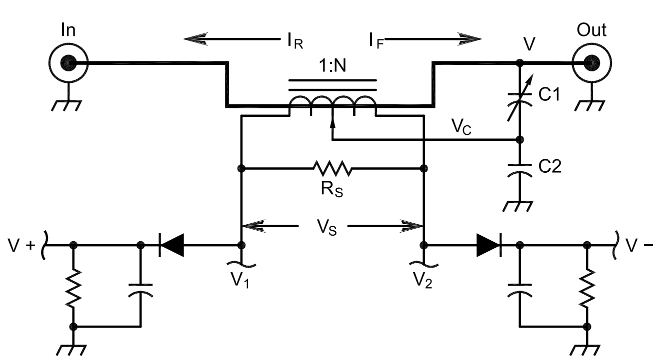

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Figure A1 — The Bruene wattmeter.

$$k_1 = \frac{C_2}{C_1 + C_2}$$

$$k_2 = \frac{R_s}{N}$$

$$I_F = \frac{V_F}{Z_0}$$

$$I_R = \frac{V_R}{Z_0}$$

$$V = V_F + V_R$$

$$V_2 = k_2\left(I_F - I_R\right)$$

If R is selected to match the calibration impedance Z_0, then these equations reduce to the desired format:

$$v_1 = \left(\frac{1}{N}\right) V_F \quad \text{(Eq B1)}$$

$$v_2 = \left(\frac{1}{N}\right) V_R \quad \text{(Eq B2)}$$

These are identical in form to Equations A5 and A6, except for a different scale factor. In fact, they have another very useful property in that the scale factor is a precise design parameter. Since N is the number of turns on the current sense transformer, it forms, in conjunction with the precision resistor of value Z_0, an exact scale factor for this sensing circuit. Equations B1 and B2 show that the sampled voltages are attenuated versions of the two modal waves, with an attenuation factor of 1/N. Hence, it is common to refer to this as a "coupling factor" whose value A is given by the standard formula:

$$A(dB) = 20 \ \log\left(\frac{1}{N}\right) \quad \text{(Eq B3)}$$

For example, one very popular design uses a transformer with 33 turn toroidal coils, and this formula shows that the resulting coupling factor is –30 dB. (i.e. 1000:1) This has the intuitive interpretation that every watt of forward (reverse) power in the measured line will produce one milliwatt of power in the forward(reverse) sample loads, where each is terminated with Z_0 ohms.

This generic design is also commercially marketed, without the detector circuits, under the name *Directional Coupler*. The meaning of this should now be clear from the above, which shows how it samples a signal within a cable into two modal components, which can then be analyzed with additional test equipment of any desired accuracy and complexity. One such application of this is the measurement of vector impedance using a directional coupler and a network analyzer or vector voltmeter. The latter instrument is required to determine the phase angle in addition to the magnitude of *Z*.

Single Wave Reversible Circuits

This is the type of circuit used in the popular Bird reversible wattmeter, which was one of the first commercial high-accuracy DDW products. Many hams still "religiously" insert a Bird wattmeter in the coax feed line out of their power amplifiers for monitoring of both amplifier and antenna.

This circuit functions similarly to the Bruene type, except that the current-sense element is only single ended. It is a small loop placed close to the center conductor of a coaxial line element, and it is characterized by both mutual and self inductances. It derives a sample of the current by means of its mutual inductance, M, with the center conductor, and a sample of the voltage by the direct capacitive coupling of the same sense loop to the line, with a capacitive coupling, C. Since the output voltage v_1 is simply the sum of v_C and v_S, the above formulas can be combined to give the following formula for v_1:

$v_1 = V_F\,(k_1 + k_2) + V_F\,(k_1 - k_2)$

$v_1 = V_F\,(k_1 + k_2) + V_R\,(k_1 - k_2)$

Impedance calibration is established by forcing $k_1 = k_2$, resulting in the final formula:

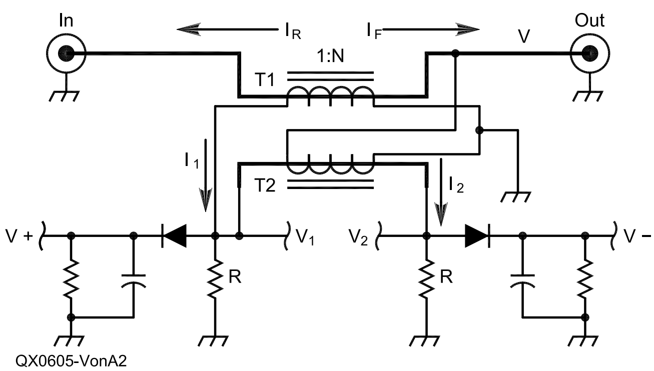
$$v_1 = 2\,k_1\,V_F \quad \text{(Eq C1)}$$

In terms of circuit quantities, this establishes the design condition that:

$$\frac{C_2}{C_1 + C_2} = \frac{R \bullet M}{Z_0 \bullet L} \quad \text{(Eq C2)}$$

In practice, most of C_1 is the parasitic coupling capacitance to the line and M / *L* is fixed by the geometry of the loop design. Since Z_0 is fixed, usually 50 Ω, the final "tweaking" is achieved by selection of *R* and C_2, the latter using some variable *C*.

This design produces a single output, which is the forward wave sample in the example shown. Nevertheless, it is truly a bidirectional device at its basic level, for otherwise it could not properly sense only one modal component. It may be seen as one half of the previous designs, and the question naturally arises: How does it sense the other component? The answer lies in the complete symmetry of the circuit with regard to wave direction, so that if input and output were interchanged, the sample would change modes from forward to reverse, and so on. This process is facilitated by placing the entire sensing circuitry in a small module with the sensing loop located near its end. The module is mounted in such a manner that it is in close proximity to the coaxial center conductor but can be physically rotated 180°, placing the loop parallel or anti-parallel to the coaxial line. Forward and reverse readings are made separately



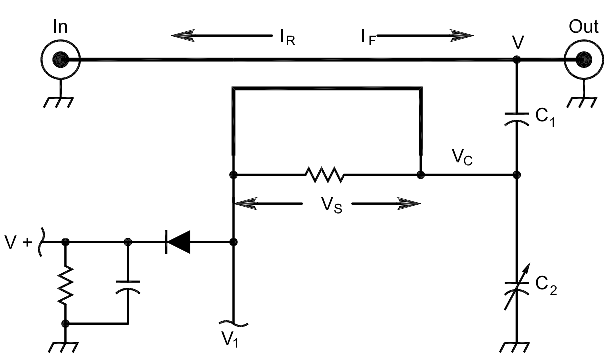
$$I_1 = \frac{I_F - I_R}{N}$$

$$V = V_F + V_R$$

$$v_1 - v_2 = \frac{V_F + V_R}{N}$$

Figure A2 — The hybrid coupler wattmeter.

M = Mutual Inductance
L = Self Inductance



$$k_1 = \frac{C_2}{C_1 + C_2}$$

$$k_2 = \left(\frac{M}{L}\right)\left(\frac{R}{Z_0}\right)$$

$$I_F = \frac{V_F}{Z_0}$$

$$I_R = \frac{V_R}{Z_0}$$

$V = V_F + V_R$

Figure A3 — The unidirectional wattmeter

by locking the "slug" (which is the common name for this module) into either of its two positions, which maximize M.

I have seen at least one commercial VHF design in which the sensing loop uses a small resistor instead of a wire, parallel to the line, combining the functions of R and M into a single component to reduce parasitic reactances. This results in a very compact sensing assembly, and the complete sensor assembly uses two identical modules placed on opposite sides of the sensed line, pointing in opposite directions.

"Waveless" Wattmeters

It is possible to explain the operation of the DDW without having to involve the action of bidirectional wave fields on transmission lines. The analysis begins exactly as for the Bruene DDW, except that we now assume ordinary circuit definitions of voltage and current at the load, where the power is to be measured. The resulting formulas for v_1 and v_2 are carried out as above, except now there is only one relevant value of V: the total line voltage. This is the voltage that would be measured by an RF voltmeter. Assuming that the load impedance is connected right at the output of the DDW, with no transmission lines present, a slight rearrangement of Equations A1 and A2 in terms of V and I at the load results in the equations:

$$v_1 = k_1 \bullet V + \frac{k_2}{2} \bullet I \qquad \text{(Eq D1)}$$

$$v_2 = k_1 \bullet V - \frac{k_2}{2} \bullet I \qquad \text{(Eq D2)}$$

It is significant that V and I now refer to the terminal voltage across and current through an arbitrary load. They are in no respect constrained by transmission line relations, but by the basic circuit definition of impedance:

$V = I \bullet Z_{LOAD}$

At this point the typical condition expressed in Equation A3 is now imposed, with a clear understanding that Z_0 is parameter that was just selected arbitrarily — it is not related to any aspects of the circuit under discussion.[1] The latter consists of just the assumed load and the (negligible) DDW circuitry with *no* transmission lines.

Substituting this into the "directional" formulas (Equations D1 and D2) produces a new format for the sample signals:

$$v_1 = k_1 \bullet V \bullet \left(1 + \frac{Z_0}{Z_{LOAD}}\right) \qquad \text{(Eq D3)}$$

$$v_2 = k_1 \bullet V \bullet \left(1 - \frac{Z_0}{Z_{LOAD}}\right) \qquad \text{(Eq D4)}$$

These results show the meaning of v_1 and v_2 with respect to V and I for an arbitrary load, but their significance as identifiable quantities is not at all obvious, as it seemed to be using modal wave analysis. To complete the analysis it is necessary to look at the quantity $(v_1^2 - v_2^2)$, which was shown in the article text to represent $P_{FWD} - P_{REV}$ in the transmission line. For the waveless case, we find this quantity by squaring both Equations D3 and D4, and subtracting them, with the result:

$$v_1^2 - v_2^2 = 4 k_1^2 \bullet V^2 \bullet \left(\frac{Z_0}{Z_{LOAD}}\right) \qquad \text{(Eq D5)}$$

$$v_1^2 - v_2^2 = \left(\frac{1}{C}\right) \bullet P_{LOAD} \qquad \text{(Eq D6)}$$

where

$$C = \frac{1}{4 k_1^2 \bullet Z_0}$$

All of the circuit quantities have been combined into a single constant, C, whose dimensions are square volts per watt.

If the voltages v_1 and v_2 are now applied to voltmeters with scale factors of C watts-per-square volts, a complete wattmeter results, according to the following inverted form of Equation D6:

$$P_{LOAD} = C\, v_1^2 - C\, v_2^2 \qquad \text{(Eq D7)}$$

where C v_1^2 and C v_2^2 are the observed meter readings in units of watts. This formula is highly significant in that it shows that a DDW based upon these designs will *always measure the true power to any load*. We have made no assumptions about transmission line waves or lumped impedances to arrive at this result. To make this perfectly clear, note that for the directional wave analysis, the final results given by Equations A5 and A6 can be similarly combined into the form:

$$v_1^2 - v_2^2 = 4\, k_1^2\, Z_0\, (P_F - P_R) \qquad \text{(Eq D8)}$$

Since $P_F - P_R$ is known to be the power transmitted to the load using the modal method, this shows that the two methods indeed produce the same result. The equality of Equations D6 and D8 proves it explicitly. This result is expressed best by Equation D7 in terms of a factor C, relating transmitted power to the difference of squares of the two voltages. This factor is determined by the hardware circuit parameters as shown in the above circuit analyses, with the result that any given design may be regarded as consisting of an RF sensing front-end, with a sensitivity factor of 1/C in W/V², driving voltmeters that have been recalibrated in units of watts.

The previous discussion should now clearly establish that the familiar dual directional wattmeters that we all have become so familiar with are not inherently dependent upon 50 Ω transmission lines. In fact, a majority of the measurements made by them occur on lines which do indeed possess a 50 Ω characteristic impedance, but the actual field on the line is not a pure 50 Ω wave. When it is not, the goal is to measure the actual power. If it could not do that it would be a useless instrument. The only significance of the 50 Ω coaxial connectors used for input and output feeds is that they provide a minimum of impedance discontinuity on 50 Ω coaxial lines, the typical application.

The practical ramifications of these interesting results are further discussed in the article, along with some interesting examples of measurements that might be hard to interpret without this demonstration of the equivalence of the wave theories and circuit theories.

Note

[1] An interesting exercise would be to construct a DDW with an adjustable Z_0 and observe its readings under the same cases shown here. As the circuit analysis here proves, adjusting this value of Z_0 while a given load is being driven would result in changing values on the two meters but their difference would remain constant, equal to the true load power. □□

Low-Profile Helix Feed for Phase 3E Satellites: System Simulation and Measurements

The authors modeled, built and measured the performance of a helix feed system for a parabolic dish suitable for 2.4 GHz satellite operation.

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Introduction

An important push toward the intensive use of helix antennas or helix-fed disks was born with a revolutionary new generation of satellites: the Phase 3 class. At the beginning of 2004, unfortunately, the Phase 3D primary battery failed, and AO40 is nonoperational. The new Phase3E (see Figure 1) is being built in Germany by AMSAT DL, and is expected to launch within the next couple of years.[1, 2, 3] The declared goal of the P3E is to offer newcomers access to high-orbit satellite communications as well as a working platform for those already operating through AO-40 in S and K modes.

On top of that, future communication links to a P5A Mars station at 10.5 GHz will be simulated and tested, which means more antennas and modules for microwave bands.[4] Starting from these considerations and specifically for people interested in communicating with the new satellites, for the first time a complete antenna system (disk plus helix feeder) is mechanically designed by *AutoCAD* (Figure 2), simulated by NEC2 and NEC Synth and finally measured in the field.

For circular polarization applications, the axial-mode helix antenna is a natural candidate because its good polarization performance is an inherent attribute of the antenna shape, without the need for a special feeding arrangement. Polarization properties of the helix have been the subject of several publications since the early work of Kraus.[5, 6] Traditionally, the pitch angle — an important parameter of the helix — may range from about 12° to 16°. The pitch angle, α, is the angle that a line tangent to the helix wire makes with the plane perpendicular to the axis of the helix.

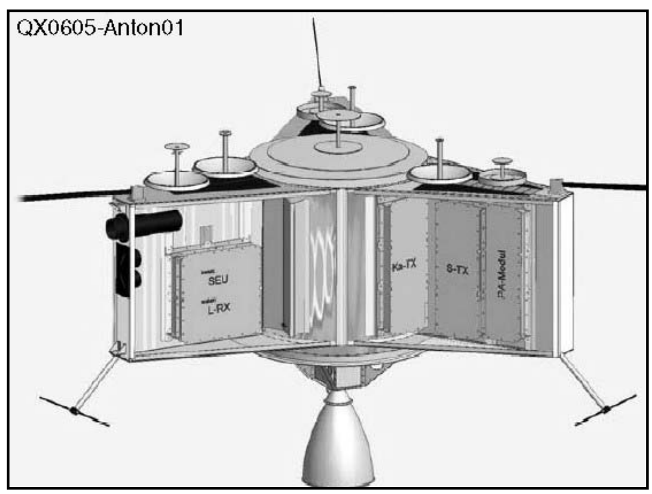

Figure 1 — A new satellite scheduled from Amsat : Phase 3E.

In the past, low-pitch helices have been recognized as ineffective radiating elements for a circularly polarized wave. Numerical results using *NEC-Win Pro* and intensive field measurements, however, lead to some low-pitch helices with gains comparable to that of a conventionally long helix.[7] The proposed complete system for the 2.4 GHz band includes a very short feed helix and a 60-cm diameter disk. Moreover, the originality of this project includes the use of the uniform

[1]Notes appear on page 39.

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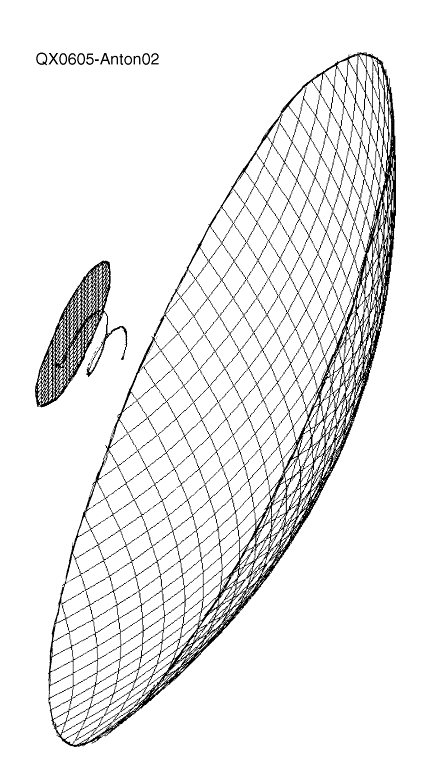

Figure 2 — Mechanical design (DXF) of the complete disk system.

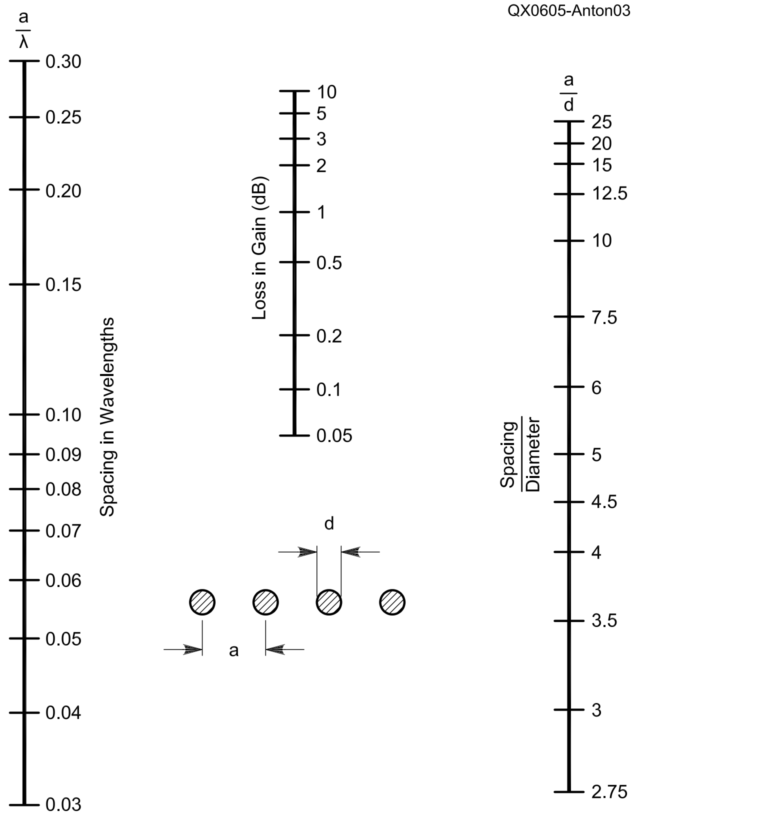

Figure 3 — Loss in gain for a mesh reflector as a function of mesh size.

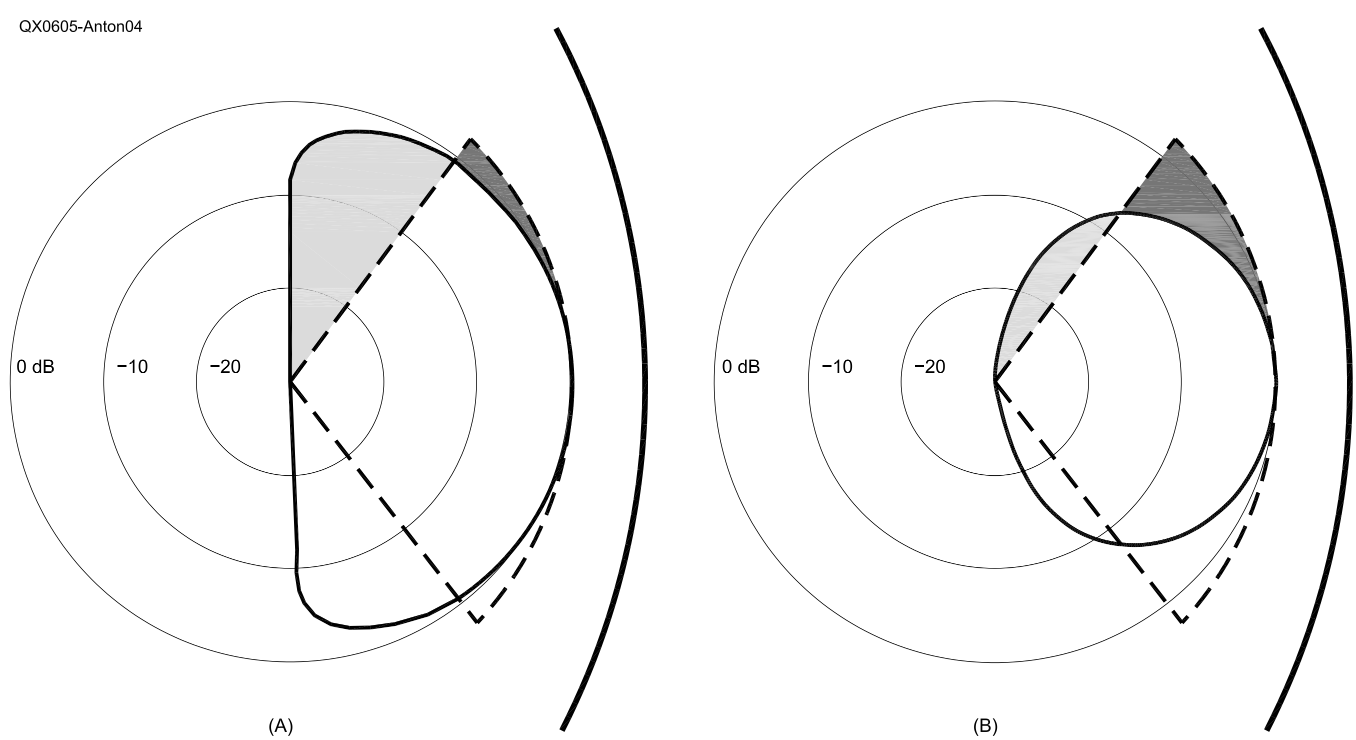

Figure 4 — Disk illumination and spillover losses (taper = 3 dB) are shown at A. Part B shows the disk illumination and spillover losses (taper = 10 dB).

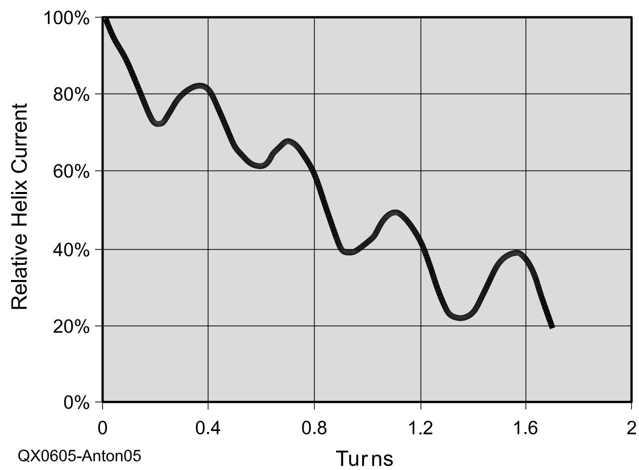

Figure 5 — Current simulation (low profile helix) using NEC-Win Pro.

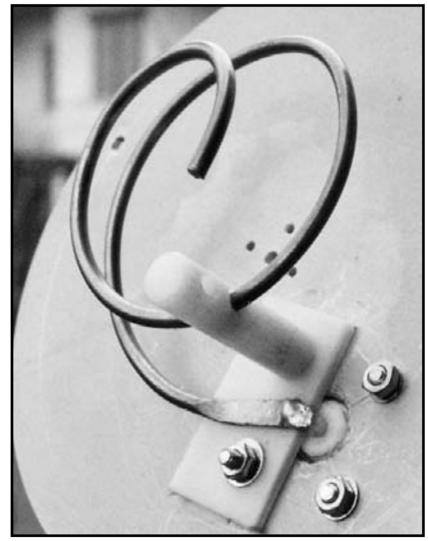

Figure 6 — A zoom on the helix feeder.

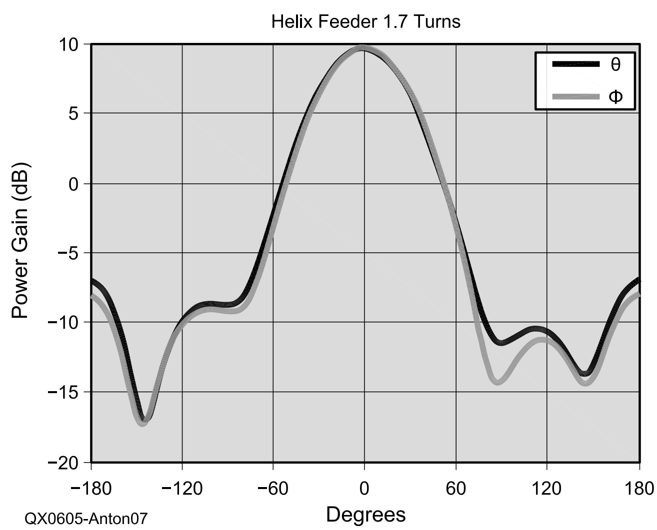

Figure 7 — Simulated radiation diagram of the low-profile feeder.

segmentation technique to simulate both the helical feeder ground plane and the surface of the disk reflector to avoid the program limitation due to the nearness of the wires in the center of round geometric figures.

The Disk Reflector

The parabolic dish antenna can provide very high power gains at microwave frequencies. At a well-defined frequency, the gain is only limited by the size and accuracy of the parabolic reflector, if the antennas are properly implemented. A dish is a parabola of rotation: a parabolic line rotated around the axis that passes through both the focus and the center of the curve. The choice of a parabola with a central focus instead of one with offset has been made considering the easiness of finding it and of exactly centering the feeder without problems.

The physical dimensions of the reflector we considered are the diameter of the disk, $D = 600$ mm, and its depth in the center, $c = 94$ mm. Using these two dimensions, we can calculate the focal length f and the f / D ratio (that is an easy way to describe our parabola), and the power gain G in dBi (referred to the isotropic):

$f = D / (16\ c) = 239.4$ mm and $f / D \approx 0.4$
$G = 10 \log_{10} [(\pi^2 D^2 \eta) / \lambda^2]$
where: λ = wavelength in millimeters
η = aerial efficiency (< 1)

This last equation comes from the capture area of an antenna and is a re-elaboration of it.

It is a simple geometric computation to consider a dish antenna pointed at an isotropic antenna that is transmitting from some distance away. The assumed isotropic antenna is a point source that radiates equally in every direction.

The efficiency is the ratio between the power received at the electrical connection to the feed, and the power that actually arrives. By hypothesis, considering the unrealistic case of a perfect lossless antenna ($\eta = 1$), the power gain can grow until $G_{MAX} = 23.6$ dBi. A more practical value of the gain should be $G = 20.6$ dBi, assuming that its efficiency drops to 50% ($\eta = 0.5$). This point will be analyzed better again, after the evaluation of the simulation results of the complete system.

As already stated in the introduction, to avoid the error problems related to the model included in *NEC-Win Synth* software, we decided to use *AutoCAD* to prepare the file for the simulation program.[7] The equation used was the classic one of the parabola of rotation — easy to find in every manual of analytic geometry:

$y = x^2 / (4\ f) + z^2 / (4\ f)$

where: x, y and z are the three coordinates in the space of the reflector expressed in millimeters.

At this point we used the *NEC-Win Synth* software for a double purpose: to translate the .DXF file to a .NEC file and to remove the possible segmentation errors caused by *AutoCAD*. It is mandatory to perform this kind of translation because .NEC is the only file format acknowledged by *NEC-Win Pro*.

To have a good trade-off between the simulation accuracy and the computation time (the number of segments of the complete system is 2102) it was decided to use a mesh reflector pitch a = 20 mm combined with a wire diameter d = 6 mm. This matching, with the aid of the chart in Figure 3, allows us to calculate the gain loss caused by the use of a non-solid reflector.[8] The values to be entered in the graph are $a / \lambda = 0.16$ and $a / d = 3.33$, which is equivalent a very low loss of gain: less than 0.05 dB (off the scale of Figure 3).

For your information, *Microsoft Excel* can also be used, but the procedure to obtain the .NEC file is more complicated because some operations must be done manually. The first step consists in the preparation of a table, using a format compatible to *NEC-Win Pro*, but limited to one half of the parabola. The resulting file is to be converted by *Excel* to text and then to .NEC. To generate the entire disk, *NEC-Win Pro* allows us to perform a rotation by 90° and a three times duplication. Another inconvenience of this procedure is the impossibility to use *NEC-Win Synth* to remove possible segmentation errors because this software does not acknowledge the "GM" instruction.

Low-Profile Helix Feeder

As a feeder for the 60-cm diameter disk we need an RF source, placed at the parabola focus, having a beam width given by the following:

$\theta = 2 \arctan [\, D / 2 (f - c) \,] \approx 128°$

Almost all feeders will provide less energy at the edge than in the center of the disk.[9] The difference in power at the edge of the disk is defined as the "edge taper." Changing the shape of the feeder radiation pattern, we can change the edge taper illuminating the disk. Different edge tapers will produce different amounts of illumination and spillover losses: a small edge taper results in a larger spillover loss, while a large edge taper reduces the spillover loss at the expense of an increase of the illumination loss. A good tradeoff, to have the maximum efficiency of the antenna, is obtained with an edge or illumination taper of about 10 dB.

Looking closely at the parabolic surface, we find that the focus distance (f = 239.4 mm) at the center is smaller than the distance d between the focus and the disk edge:

$d = \mathrm{sqrt}\,[(D / 2)^2 + (f - c)^2] = 333$ mm

So, the wave that follows the edge way has been submitted to an additional "ΔA" attenuation:

$\Delta A = 20 \log_{10} (d / f) = 2.9$ dB

This equation comes from the free space attenuation formula where the variables, frequency and gain, are constant (also defined as inverse square law) and so do not appear in the formula.

Considering the additional attenuation just analyzed, the optimum illumination taper becomes: 7.1 dB (10 dB – 2.9 dB). The feeder comprises every microwave device able to concentrate the beam, on the reflector, within the already calculated beamwidth θ: dipole, horn, waveguide, helix and so on. We decided to use the helix, considering our previous experience in simulating and manufacturing this kind of antenna.[10, 11, 12] In Figure 4 we can see the spillover and illumination losses in two different situations (3 dB taper and 10 dB taper).

The behavior of the current versus length of a typical helix shows three different regions:

1) Near the feed point, where the current decay is exponential.

2) Near the open end, with visible standing waves.

3) Between the two helix ends, where there is a relatively uniform current and small SWR (transmission line).

There are two ways to obtain a good circular polarization helix:

1) Tapering the helical turns near the open end, to reduce the reflected current from the arm end, and

2) Using only the first helical turns where the decaying current travels from the feed point to the first minimum point (see Figure 5).

Starting from these considerations our final low-profile helix uses a pitch $a = 0.16\ \lambda$ (20 mm at 2.4 GHz) and is both conically wound with conic 62 / 41 mm diameters and very short (only 1.7 turns — as shown in the photo of Figure 6).

Let us summarize the mechanical dimensions of the feeder chosen to supply our 60 cm disk:

1) Number of turns: 1.7 (10 segments per turn).

2) Pitch between contiguous turns: 20 mm.

3) Starting diameter: 62 mm.

4) Ending diameter: 41 mm.

5) Helix wire diameter 3 mm.

Table 1
Different Simulated Low-Profile Helices (1.7 Turns, Pitch=20 mm)

Helix Diameter and Shape (mm)	*Reflector Type and Size (mm)*	*Gain (dB) f = **2.2** GHz*	*Gain (dB) f = **2.4** GHz*	*Gain (dB) f = **2.6** GHz*
67.5 / 45.0 Conical	Square 125 × 125	9.49	9.64	8.80
62.0 / 41.0 Conical	Square 125 × 125	9.35	9.73	9.84
62.0 / 41.0 Conical	**Circular** (dia = 124)	**9.31**	**9.59**	**9.55**
56.0 Cylindrical	Square 125 × 125	9.38	9.68	8.52
67.5 / 45.0 Conical	Square 75 × 75	9.15	9.28	8.36
62.0 / 41.0 Conical	Square 75 × 75	8.84	9.29	9.39
56.0 Cylindrical	Square 75 × 75	8.90	9.19	8.38

Table 2
Simulation of Gain and F/B @ 2.4 GHz of the Complete System vs Feeder Positioning

Distance between Feeder Reflector and Parabola Surface (mm)	*Power Gain (dB)*	*–3db Radiation Angle (degrees)*	*Front to Back Ratio (dB)*	*Notes*
220	20.33	16	32.24	High Illumination Loss and Very Low Spillover Loss
230	20.23	16	30.98	
240	20.19	15	30.34	
250	20.57	15	30.08	
260	21.25	16	28.6	
265	**21.38**	**16**	**27.87**	Best power Gain
270	21.35	16	27.45	
280	21.05	15	26.85	
290	20.90	16	26.30	

Feeder Reflector Diameter = 125 mm

6) Stub: 6 mm long, 2 mm wire diameter, only for simulation purposes to contain the current in the screen.

7) Screen diameter: 125 mm, wire diameter 1 mm.

The just-defined feeder has been accomplished using 674 segments: 18 for the helix and the stub and the remaining 654 to define the screen with a 6.25 mm segmentation pitch.

The simulation of the feeder, using *NEC Win Pro* software (see Note 10 and Table 1), gave the following results at the frequency of 2.4 GHz:

Power gain: G = 9.6 dB

Radiation angle: 59° at –3 dB, 91° at –7.1 dB, 107° at –10 dB, 128° at –14 dB, 146° at –17 dB.

Front to back ratio: 16.6 dB.

The power gain versus the angles θ and ϕ that are in two planes perpendicular between them and parallel to the wave propagation direction is plotted in Figure 7.

As you can see from the simulation results, our 21-dB-gain disk has a high illumination loss, but this is an advantage from the F/B, secondary lobes and spillover point of view.

The Complete System

During the simulation phase, we made some changes in the original file to optimize the performance of our antenna. The first one involved the diameter of the feeder reflector. We reduced it from 125 to 100 mm in order to reduce its masking effect on the beam reflected by the disk surface by 36%. The advantage of this operation was insignificant both for power gain (–0.05 dB) and for front-to-back ratio (–1 dB) so we decided to come back to the original screen: 125 mm diameter.

The second action involved increasing the distance between the feeder ground plane and the parabola surface from 220 to 290 mm. The best distance, from the power gain point of view, which is the most important parameter for the amateur, was 265 mm (see Table 2).

Theoretically, the disk receiving antenna works as follows: the electromagnetic waves coming on a parallel path from a distant source are reflected by the parabola to a common point: the parabola focus point. A transmitting antenna reverses the path: the radio wave generated by a point source placed at the focus point is reflected into a beam of rays parallel to the axis of the parabola.

The differences between the theory and the practice are related to the following:

1) The focus is only a geometric point and the feeder is bigger than that.

2) It is not so easy to find the electrical center of a helix having a truncated cone shape. Considering the simulation results, and the calculated focal length, it can be put at about ¾ of the total feeder length (34 mm).

After the implementation of the whole optimization described, the simulation results at the frequency of 2.4GHz are summarized here:

Power Gain: 21.4 dBi

Beamwidth: 16° at –3 dB

Front to back ratio: 29.7 dB

Figure 8 shows the power gain graphs, as a function both of the azimuthal and elevation angles ϕ and θ.

Now we are able to calculate the antenna efficiency η that is the ratio between the gain, obtained from the simulation, and the maximum possible gain G_{MAX}, already computed using the equation of aerial capture area:

$$\eta = (10^{G/10}) / [(\pi D) / \lambda]^2 \approx 60\%$$

The efficiency loss of our complete system (~2.2 dB) can be referred to the disk illumination (see Table 2). We consider a uniform illumination of the disk and so the increased attenuation caused by the different path that the electromagnetic wave covers at the periphery of the reflector with respect to the center. The spillover and side-lobe losses can be neglected.

Measuring the Disk

The complete system is shown in Figure 9. In our experience, it's not very difficult to design and make helices for different working frequencies and gains; but more difficult, for the serious experimenter, is a way of precise measurements.

Before starting with measurements, it is important to define the polarization sense of the helices used in the test setup. The IEEE definition that has become the standard is that in viewing the antenna from the feed-point end, a clockwise wind results in right-hand circular polarization (RHC), and a counterclockwise wind results in left-hand circular polarization (LHC). This is important, because when two stations use helical antennas *over a nonreflective path*, both must use antennas with the same polarization sense. If antennas of opposite sense are used, a signal loss of at least 20 to 30 dB results from cross polarization alone. Remember that the complete disk produces circular polarization with a sense *opposite* that of the feed helix

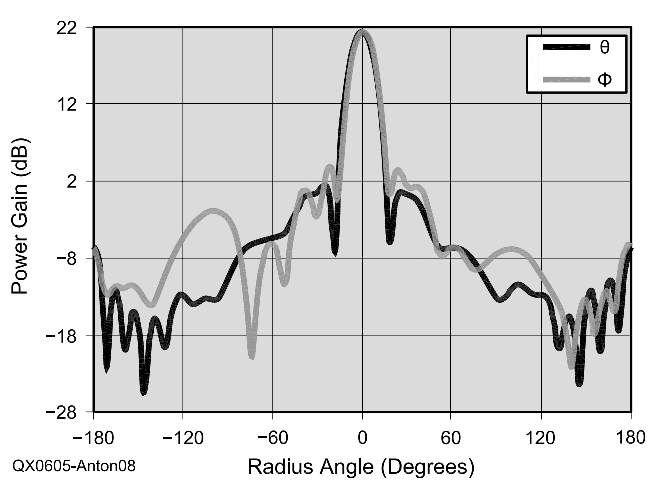

Figure 8 — Simulated radiation diagram of the complete system.

Figure 9 — The complete 60 cm disk antenna.

(because of the reflection of the wave from the parabolic surface).

For almost all tests, we used a generator from 2.2 to 2.7 GHz composed of a VCO, JTOS-3000 by Mini-Circuits, followed by the MNA-6 broadband amplifier (3 × 3 mm), or the Keps 0535AN2. The output level from the oscillator is very high (+10 dBm), but some attenuation must be included for stability (the wide-bandwidth amplifiers oscillate easily with loads not exactly 50 Ω). Using a Boonton RF Millivoltmeter (mod 92B) as a detector, we have a sensitivity loss of about 10 dB at 2.4 GHz (referred to the maximum suggested operating frequency). More suited to the measurements is the classic HP 431B power meter.

To limit the measurement errors, the distance between transmitting and receiving antennas has to be considered. To determine this distance, you need to be able to measure the signal level easily with a filtered RF voltmeter having a 20 to 30 dB dynamic range. Also, the wave reaching the receiving antenna should be as planar as possible.

The first condition can be easily established, starting with the received power and calculating the attenuation experienced by the wave in open space:

$A = 32.4 + 20 \log (f) + 20 \log (d) \quad G_t - G_r$

where A is the attenuation in decibels, f is the frequency in megahertz, d is the distance in km, G_t is the gain of the transmitting antenna in dBi and G_r is the gain of the receiving antenna, also in dBi, obtained by simulation.

There is also a simple, easy to remember method of calculating the free-space attenuation by considering the distance between the two antennas in terms of wavelengths. When the distance between two isotropic antennas is $d = \lambda$, the free-space attenuation is always 22 dB. This distance is 12.5 cm at 2400 MHz. The attenuation increases by 6 dB for each doubling of the path distance. This means that the free-space attenuation is 22 dB at 0.125 m, 28 dB at 0.25 m, 34 dB at 0.5 m, and so on. To make the wave reaching the receiving antenna as planar as possible, the capture area in square meters of the receiving antenna is:

$$A_c = G_r \lambda^2 / 4 \pi$$

This expression is valid for an antenna with no thermal losses and was certainly useful for our experiments. With a circular capture area, the minimum distance in meters between the antennas will be:

$$d > n D^2 / \lambda$$

where D: disk diameter in meters.

A maximum acceptable phase error will also be considered. For a phase error of 22.5°, which is usually enough, $n = 2$. If a phase error of only 5° is required, $n = 9$.[13, 14] See Figure 10 for free space attenuation at 2.4 GHz. The site (in the garden) we selected is particularly useful for all our helix measurements (d = 4 m = 32 λ at 2400 MHz).

Table 3 compares the simulation results with our measurements.

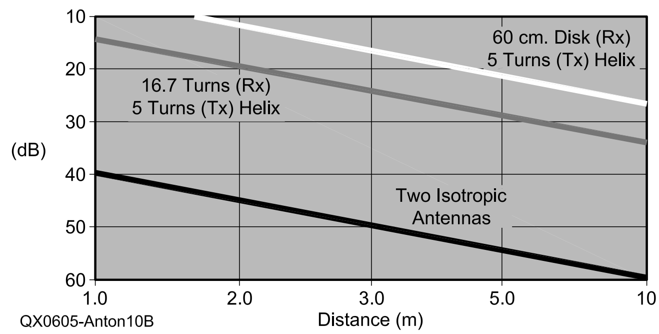

Figure10 — Free space attenuation @ 2.4 GHz.

Table 3
Comparison Between 60 cm Diameter Disk and Standard 16.7 Turns Helix

	Simulation Results			*Measurements*	
Type of Antenna	*Power gain (*dBi)	*Radiation Angle –3* dB	*Front to back Ratio (dB)*	*Radiation Angle –3 dB*	*Relative Gain* (dB)
60 *cm disk*	21.4	16	27.9	14	+7
16.7 turn helix	14.5	26	18.5	30	0

Notes

[1] "P3-Express — An Intermediate Report," *AMSAT-DL Journal*, Dec/Feb 2003/2004.
[2] "AMSAT-Phase 3 EXPRESS," *AMSAT-DL Journal*, Sep/Nov 2003.
[3] "P3E News," *AMSAT-DL Journal*, Mar/May 2004.
[4] P. Antoniazzi and M. Arecco, "Chaparral Style 10 GHz Feed Simulation," *DUBUS*, N.2/2005, pp 38 to 50.
[5] John D.Kraus, The Helical Antenna, pp 265 to 339, *Antennas*, McGraw-Hill, 2nd Edition, 1988.
[6] H. Nakano,Y. Samada and J. Yamauchi, "Axial Mode Helical Antennas," *IEEE Transactions on Antennas and Propagation*, 1986, pp 1143 to 1148.
[7] **www.nittany-scientific.com** (NEC-Win Pro Rel 1.4 and NEC-Win Synth Rel 1.0).
[8] *Microwave Handbook*, Volume 1, RSGB, 1989, p 4.21
[9] W1GHz, *Microwave Antenna Handbook Online*, **www.w1ghz.org**.
[10] P. Antoniazzi and M. Arecco, "Simulating and Making Low Profile WiFi Antennae," *Electronics World*, Dec 2004, pp 22 to 25.
[11] P. Antoniazzi and M. Arecco, "Measuring 2.4 GHz Helix Antennas using Slotted Lines," *Electronics World*, Jul 2003, pp 47 to 52.
[12] P. Antoniazzi and M. Arecco, "Measuring 2.4 GHz Helix Antennas," *QEX*, May/Jun 2004, pp 14 to 22.
[13] P. Antoniazzi and M. Arecco, "Measuring Yagis," *Electronics World*, Dec 1998, pp 1002 to 1006.
[14] D. R. J. White, *Electromagnetic Interference and Compatibility*, Vol 2, D.White Consult, Inc, 1980, pp 2.11 to 2.13.

Paolo Antoniazzi was licensed as IW2ACD and became a member of the Associazione Radiamatori Italiani (ARI) in Italy in 1961. He worked for Siemens, GTE, Sprague and ST Microelectronics in the fields of RF and telecommunications marketing and applications until his retirement in November 2003. Paolo is particularly interested in simulation and measurements (and writing papers) about narrow-band LF, 2.4 GHz, and more recently, the 10 GHz band. Space communications with the future P3E and P5A satellites is a priority interest for the future.

Marco Arecco, IK2WAQ, matured his professional experience in semiconductor manufacturing, working for STMicroelectronics for more than 35 years. His last position, before retirement in early 2004, was as the Nonvolatile Memory Product Engineering Team leader. Marco is a member of ARI. He has been licensed 12 years and his Amateur Radio interests include LF operation and VHF, UHF and microwave antenna simulations and measurements. □□

A Talking Logbook with Rig Control

The artificial speech synthesis capabilities of Windows *PCs assist the visually impaired with logging and basic rig control.*

Steve Gradijan, WB5KIA

A free, general purpose logging program called *TalkingLogBook* (Figure 1) keeps track of contacts and critical transceiver functions. It announces transceiver readings, settings and reviews logbook information. Its voice is that of Microsoft's Sam, the artificial speech voice included with *Windows XP* and available for earlier versions of *Windows* including *Windows 98*.

An op-ed article written in a late 2004 issue of *QST* magazine bemoaned the problems of handicapped radio amateurs and the lack of inexpensive solutions. *TalkingLogBook* was written to help close that gap by providing speech facilities for computer logging and to support operation of legacy transceivers.

The Microsoft SAPI 5.1, an activeX programming control, *TTSX*, developed by WAØTTN, a "universal" Amateur Radio control program called *OmniRig* by VE3NEA, *Halcyon*, a *Delphi* compatible database control by Griffin Solutions, Inc and a Borland *Delphi* compiler were used to create *TalkingLogBook.* This logging program allows the visually impaired to listen to the contents of an electronic logbook and review basic transceiver frequency and mode information using a *Windows* based PC.

Software applications oriented towards handicapped radio amateurs are needed that automate the logging process, have large print fonts and provide aural information to the user. The tools need to work with both PC controllable transceivers and those that cannot be controlled. The *TalkingLogBook* program supports most modern Kenwood, Yaesu, ICOM and a few other transceivers. The program can be used with automatic or manual frequency and mode control. Users without PC controllable radios can use most of the program's features. ADIF file transfer protocol is supported to permit "transfer" of log information from an existing program into *TalkingLogBook* or the other way around, if necessary.

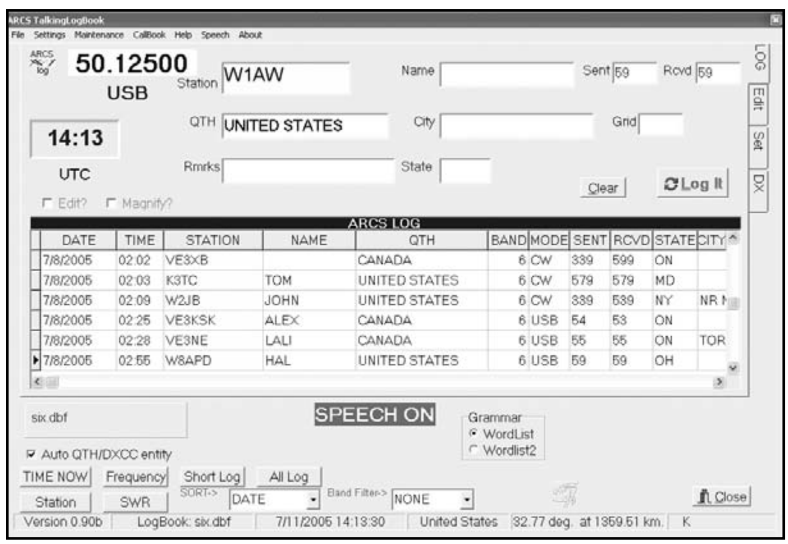

Figure 1 — The *TalkingLogBook* main screen allows review of logbook contents, announcements of current time, radio frequency and mode and other features using the synthetic speech capabilities of Microsoft *Windows*.

Three pieces of software work together to provide the logbook functions as shown diagrammatically in Figure 2. The umbrella program is the *TalkingLogBook* executable file compiled with the Borland *Delphi* compiler. It provides the logging and sorting functions and interfaces with the *TTSX* activeX control and the *OmniRig* server to provide artificial speech capabilities and automatic frequency and mode control respectively. A second version of *TalkingLogBook* is programmed without *TTSX* and can be used with *Windows XP* (or versions of *Windows 98* or higher that have SAPI 5 installed).

OmniRig is a server program that provides a software interface between the user's transceiver and PC. It is the engine that provides serial communication links for numerous transceivers as described in the *OmniRig* documentation. Many different transceivers can be linked to the umbrella program. *TTSX* is a program that streamlines the process of coding speech capabilities into *TalkingLogBook*. It also loads the required Microsoft SAPI onto operating systems other than Windows XP. It contains the routines necessary to trigger text-to-speech actions through control of the Microsoft SAPI. A *dBase* database engine

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called *Halcyon* provides Structured Query Language (SQL) capabilities to sort and retrieve information from individual logbooks and to create the actual logbook files. An editable text file provides grammar rules that allow enunciation of strange or unusual words or abbreviations unique to Amateur Radio with the program version using *TTSX*.

The mouse, short-cut keys and the up/down arrows on the keyboard control the program functions. Rig mode and frequency changes are made using the radio or the self contained, small frequency database.

Audible cues pass logging and rig control information from features that are accessed by both short-cut keys and/or mouse clicks. Error messages may be either audible or visual. Keyboard characters may be monitored and read back using the PC's sound card as individual characters are typed. The contents of logbook fields can be queried prior to writing them to the electronic logbook database. Because users can control basic functions with the short-cut keys and audible cues are provided at several levels, anyone who is completely blind but is computer literate should be able to use the program.

A default logbook is loaded automatically when the program is loaded. Alternative logbooks can be loaded or new logs created just as in a normal logbook program, however, audible cues are available during the process.

TalkingLogBook is easy to use but, because of the complexity of the software routines, initial program installation, configuration and the settings for a few features require the support of a sighted or partially sighted person.

TalkingLogBook was programmed using Delphi. However, the activeX controls that make voice announcements and a wide range of radios possible can also work with *Visual Basic*, *Visual C++* or *C++*.

Does Your Transceiver Support PC Control?

Most modern and some older transceivers can be PC controlled but the radios might require a little sleight of hand to accomplish this magic. How PC-to-radio communication is accomplished depends on the radio in use and the software that will control the radio or other peripheral equipment. This is why it is not possible to provide a do-it-all logging program using software alone.

All modern radios are not created equal, nor are they necessarily capable of communicating with your computer right out of the box. Several popular, modern radios require accessories not supplied by the manufacturer with the basic radio. A "level" converter may be necessary to establish PC logging or CAT capabilities (notably ICOM and Yaesu transceivers — the IC-7800 is an exception). The latest Kenwood, Yaesu, Ten-Tec and Elecraft transceivers generally are equipped with RS-232 capable ports, ready to accept a cable from your PC's COM port.

Some radios are more amenable to PC controls than others, as shown in Table 1. If a radio has an RS-232 port, it probably can be controlled without a level converter.

Level converters are usually available from the radio manufacturer and cost in the range of \$100. It is possible to build level converters from plans that are contained in some versions of the *ARRL Handbook* or from plans on the Internet with a parts outlay of around \$10.

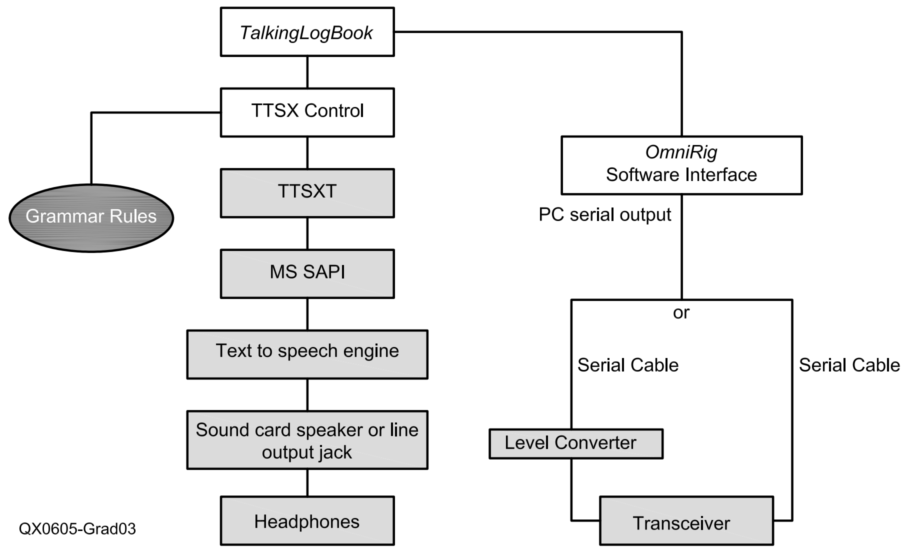

Figure 2 — A flow chart of the *TalkingLogBook* hardware and software modules.

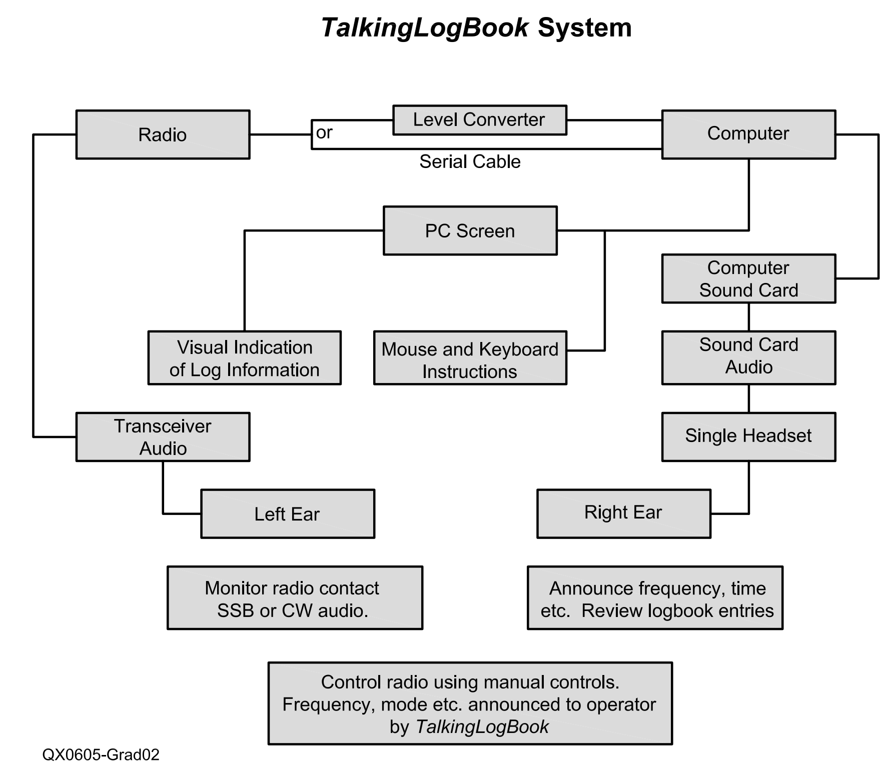

Figure 3 — *TalkingLogBook* might best be used in conjunction with a single-headset headphone to listen to the audio cues as outlined in the text.

Table 1
Computer Control Capabilities of Representative Modern Amateur Transceivers

Transceiver	*RS232 Port*	*Level Converter*	*Control Basic Functions*	*Advanced Functions*	*PC Keyer Control*
Kenwood TS-2000	yes	Not needed	yes	yes, lots	yes
Kenwood TS-570	yes	Not needed	yes	yes	yes
Icom-756 Pro II	no	Required	yes	yes, limited	limited
Icom-746 Pro	no	Required	yes	yes, limited	limited
Icom-706 MkIIG	no	Required	yes	no	no
Yaesu (recent)	unknown	Some	yes		unknown
Ten-Tec Pegasus	yes	Not needed	yes	yes	unknown

Table 2
Radios Known to be Compatible with *OmniRig* and *TalkingLogBook*

ALINCO	*ELECRAFT*	*ICOM*	*JST*	*KENWOOD*	*TEN-TEC*	*YAESU*
DX-77	K2	IC-706II	JST-245	TS-440	Paragon II	FT-100
		IC-706MKIIG		TS-570	Orion	FT-817
		IC-718		TS-850		FT-847
		IC-735		TS-870		FT-857
		IC-737		TS-2000		FT-897
		IC-746 PRO				FT-900
		IC-756 PRO II				FT-920
		IC-761				FT-990
		IC-765				FT-1000
		IC-7800				FT-1000MP

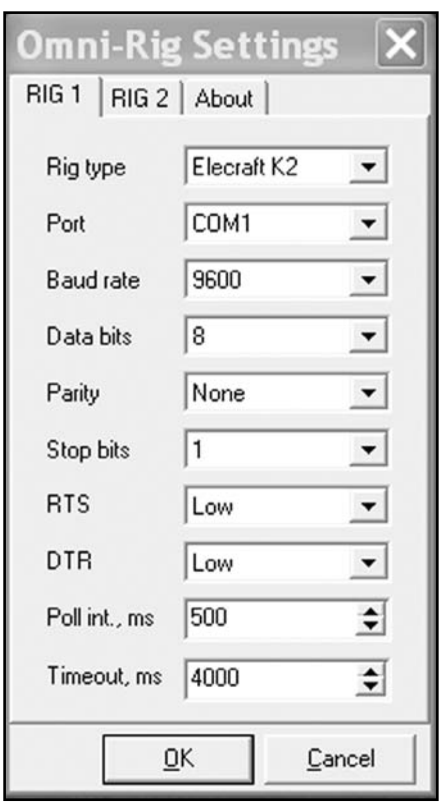

Figure 4 — *OmniRig* configuration screen set for use with the Elecraft K2 radio.

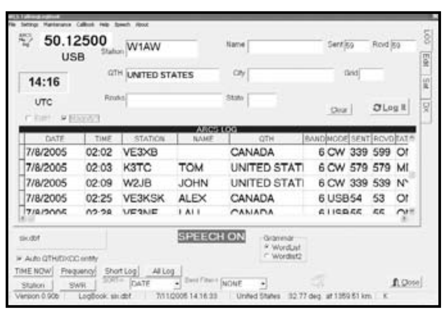

Figure 5 — *TalkingLogBook* adjustable fonts improve readability of the logbook entries for some individuals.

Radios Supported

TalkingLogBook is usable today for automatic frequency and mode logging with the radios listed in Table 2. All other radios can be used with the logging software in manual mode (no PC connection). All the program's functions, except the automatic frequency and mode logging and the program's internal frequency memory channels, which are used to control rig frequency and mode, are functional in the manual mode.

TalkingLogBook was tested with the Kenwood TS-2000 and the ICOM IC-706 MKIIG. Read the *OmniRig* documentation to learn how to set up and interface the other radios listed. VE3NEA continually updates the *OmniRig* database of compatible radios; check Alex's Web site for new additions. If your PC controllable radio is not listed, it may be possible to program the OmniRig.ini file to provide the appropriate data links/commands as described on the *OmniRig* Web site.

Installation

You must have a copy of *Windows 98* or higher to use the artificial speech feature. *Windows 95* and earlier versions are not supported. A Pentium II or equivalent processor at 233 MHz with 128 MB of RAM or better is recommended. Microsoft reports that not all soundcards are supported by SAPI 5 although I have loaded the software on three PCs and one laptop and all the installations worked fine.

The installation of the software is more complicated than most. Two support programs are required and, at present, they must be installed independently from the logging program. All three programs install easily, however. Download *OmniRig* from the Internet at **www.dxatlas.com/omnirig** and *TTSX* at **www.netdave.com/wa0ttn/**

(go to the section describing PSK31). *TalkingLogBook* may be downloaded at **www.qsl.net/wb5kia**.[1]

Follow the installation instructions for the *helper* programs. *OmniRig* by VE3NEA is a server program used to communicate between a large number of transceivers and *TalkingLogBook*. *TTSX* is the activeX control developed by WAØTTN to communicate with Microsoft's artificial speech system. Microsoft's *Speech System* must also be installed on your PC. Microsoft *Windows XP* includes the speech system; however, all users must install the WAØTTN *TTSX* interface. *TTSX* will install the SAPI modules necessary (if you do not use *Windows XP*) and establishes the program links to run artificial speech with the *TalkingLogBook*. Next, install *OmniRig* to provide the radio to PC link. Install *TalkingLogBook* last. Either unzip *TalkingLogBook* into a file folder of your choice and run the executable or use the install program (depending on which version of the software you download). Follow the setup instructions.

I recommend you download the program and its support programs from the home web sites as described above. If you do not have an Internet connection, *TalkingLogBook* is available on a CD ROM. The disk is free, though I request $10 per copy to cover packaging, reproduction and mailing expenses.

Setup

TalkingLogBook requires a compatible transceiver and a PC running *Microsoft Windows* (It has been tested only with *Windows XP* so far but will work on *Windows 98* and higher, too). It will not work with *Windows 95* or earlier *Windows* versions. A *straight through* serial cable is required to connect the PC to the radio or the radio's level adapter. Do not use a null-modem cable; it will not work! A single ear headphone is recommended to listen to announcements while in QSO. Plug the headset into your soundcard's line output or speaker jack. A typical use of the software is diagrammed in Figure 3.

Run the configuration part of the program to set up the radio type, communication port and data transfer speeds (see Figure 4) from the *TalkingLogBook* SETTINGS menu or from *OmniRig*. Kenwood radios need the RTS set to HANDSHAKE and the data rate to 57,600 kbps. The RTS on ICOM radios should be set to LOW. The data rate for ICOM radios is limited to a maximum of 19,200 kbps. The data rate should be set to identical values in both the program and the radio. Follow the instructions from your radio's manual to set the data rate of the radio. DATA BITS should be set to 8 and PARITY to NONE for all radios. POLL INTERVAL and TIMEOUT are partially dependent on your PC speed, length of serial cable and whether a true serial connection or a USB to serial converter is used.

Features

- Microsoft's artificial speech voice known as "Sam" reads the electronic logbook information on request. Other voices may be substituted, see the text. Limited CAT capabilities are provided. *TalkingLogBook* (*TLB*) reads the rig frequency and mode, also SWR with some transceivers. A small "external" memory data base is provided to control simplex frequency input.
- Audible cues and short-cut keys are used.
- Key strokes are announced if the Audible Typing box is checked.
- Current time and rig frequency announcements are available.
- An editable "grammar" file allows the "Sam" voice to pronounce abbreviations of states or radio shorthand as the full words with one *TLB* version.
- Logbooks may be queried in "standard" and "magnified" type fonts as shown in Figure 5.
- The program automatically recognizes, lists and announces previous contacts with a logged station when the call sign is entered into the STATION log field.
- The program provides band and license restricted frequency-limit cues.
- SWR is announced with some transceivers.
- Accesses the *Radio Amateur Callbook* if this callbook disk is available. *TLB* can vocalize the information.

[1]The qsl.net Web site is quite slow at times. You can download the program version that is current at publication time from the ARRL Web site at **www.arrl.org/qexfiles/**. Look for 5x06Gradijan.zip. You may want to check the author's Web site periodically for updated versions.

Table 3
Recommended *OmniRig* Configuration Settings. Selections in Bold Text are Preferred.

Equipment	*Data Rate*	*Stop bits*	*RTS*	*DTR*	*Poll Int.*	*Timeout*
Kenwood	4800-**57600**	**1** or 2	Handshake	Low	100	2000
Icom	1200-**19200**	2	Low	Low	100	2000
Yaesu	4800	2	Low	Low	100	2000
Ten-Tec	4800	2	Low	Low	100	2000

TalkingLogBook and ARRL's *LogBook of the World*

ARRL's *LogBook of the World* (*LoTW*) software provides a secure method of confirming contacts electronically. If you have been thinking about registering with *LoTW*, or already have registered, then *TalkingLogBook* will help you submit your electronic logs to *LoTW*. If you want to learn more about *LoTW*, or want to register with the program, you should read H. Ward Silver's (NØAX) September 2005 *QST* article, "The LogBook of the World — 75 Million QSOs Can't Be Wrong!" on pages 50 to 53. You can also find more information on the ARRL Web site at **www.arrl.org/lotw/**. The text of Ward's *QST* article is also available on that site.

Once you have downloaded and installed the *LoTW* software and obtained your software "certificate," you are ready to upload your logs to the ARRL *LogBook* Web site. *TalkingLogBook* will create the appropriate file, by converting the log file to the Amateur Digital Interchange Format (ADIF).

To prepare the log that you have created in *TalkingLogBook*, simply go to the "Maintenance" menu and select "Create/Convert ADIF." Select the "TLB to ADIF" button and then select the log file you wish to convert, and select "Open." The program will write a new file into the *TalkingLogBook* main folder, with the same name as the log file you wanted to convert, but using the .ADI file extension.

Now you are ready to "sign" the ADIF file with the *TQSL* program that you downloaded with the *LoTW* software. The *TQSL* program adds a digital signature to each QSO in the log. This is the method *LoTW* uses to ensure the log data really came from you. Look for the new ADIF file you created with *TalkingLogBook*. It will have a suffix of .adi. In the *TQSL* program you will select "Sign Existing ADIF File" and then select the file you created. *TQSL* will ask you for a file name to save the signed log document and will create that file with a file extension of .tq8.

Finally, send the .tq8 file as an e-mail attachment to **lotw-logs@arrl.org**, or upload it via the LoTW Web site, and you are done. You will receive a confirmation e-mail after your log has been uploaded and processed by *LogBook of the World*. — *Larry Wolfgang, WR1B*

- A module converts ADIF files from other programs to import logbook information.
- A module exports native *TalkingLogBook* files into ADIF format to export logbook information.

Who Can Use *TalkingLogBook* Effectively?

The program is intentionally easy to use. *TLB* is an electronic logbook plus basic CAT program. There are only a few frills. Operating it is straightforward but effective. *TalkingLogBook* was initially designed for the visually impaired. Blind radio amateurs may find the program useful. Large standard fonts can be "magnified" for other users. The speech capability may be useful to either blind or visually impaired individuals.

The Future

A keyboard text to speech capability is in development using a text buffer. Anything typed using your PCs keyboard terminal might be vocalized and sent to the transceiver audio circuits. Radio amateurs with speech impairments might be able to effectively use voice communications by typing text rather than having to avoid SSB or FM modes. Unfortunately, the unnatural sounding characteristics of Microsoft Sam's artificial voice may limit this use. Commercial voice modules do provide more naturally sounding voices but are not free. A female counterpart to "Sam" and an alternative male voice are available free from Microsoft. Both of these have slightly better characteristics. Recorded speech in wave files in conjunction with a "look-up" dictionary might be used to provide a limited, but highly intelligible vocabulary as an alternative to the current synthetic speech system in a future version of the program. Wave file audio could provide correct diction. An extensive Amateur Radio vocabulary of *.wav audio files would be necessary and might be practical with high end PCs.

Microsoft's speech system also has limited speech recognition capabilities not presently utilized in the *TalkingLogBook*. A data input by voice command version for individuals who find it difficult or impossible to use a keyboard is possible. A prototype is in development and may be the topic of a future article. Commercial software is available that converts speech into text. A challenge is to convert SSB audio to text. FM audio might be easier. My early tests with text conversion of streaming audio is that the Microsoft *recognizer* requires training (providing a speech sample). My voice and others that have been trained work reasonably well; the difficulty is in getting the recognizer to understand untrained voices from your radio.

A contest version of the program is possible if there is sufficient user interest and additional developers become involved in the project.

The program currently generates beam headings and makes call sign prefix identification for the various DXCC entities. Rotator control could be linked to the current beam heading capabilities for DX stations using Dynamic Date Exchange (DDE) or newer technologies. Some rotators are software friendly, others require modifications to control box circuitry to allow software control. Rotator implementation will be a challenge from both the hardware and software perspectives. Several commercial logging programs already have the capability to control several rotators. The rotator control hardware is expensive in most instances. What is needed is affordable hardware solutions that can be linked to programs like *TLB*.

A facility to read PSK and RTTY text is possible using the technology employed here and the PSK Core DLL developed by AE4JY (**www.qsl.net/ae4jy/pskcoredll.htm**), the MMTTY DLL by JE3HHT (**mmhamsoft.ham-radio.ch/#MMVari**) and/or the PSK ActiveX control developed by WAØTTN (**www.netdave.com/wa0ttn/PSK31.asp**). It is realistic to develop simple software to allow the visually impaired use of PSK and RTTY.

TalkingLogBook includes only the Microsoft "Sam" artificial voice provided by the *TTSX* control. Microsoft's *Mary* and *Mike* voices are also free. The additional voices are in a file called **Sp5TTIntXP.msm** that is used by developers to distribute the artificial "voices." Special developer software I do not have is required to make the file usable on a target PC. However, these voices are added to a user's PC when the entire free Microsoft SDK is installed on a PC. The SDK is over 68 MB and is available on disk for a nominal fee. Download the Speech SDK 5.1 from **www.microsoft.com/speech/download/SDK51**. Install the SDK. The install places an icon called SPEECH in the *Windows*' Control Panel. Select and double-click the SPEECH icon from the Control Panel. Go to the TEXT TO SPEECH tab and make a VOICE selection. Finally, click the APPLY button at the bottom of the window. Now, when you run the *TalkingLogBook*, the voice you select will be used. I personally prefer the MIKE and MARY voices.

Other Aids

Windows XP and some other flavors of *Windows* already contain features that might be of use to handicapped radio amateurs. Tools like *Magnifier*, *Narrator* and *On-screen Keyboard* may be accessed from the *Windows* menu (Accessories->Accessibility-> followed by the named feature). These helper utilities may be used in conjunction with the *TalkingLogBook* or other Amateur Radio oriented software. *Windows XP* users may also access *Magnifier* and *On-screen Keyboard* using *TalkingLogBook's* ACCESSIBILITY menu toolbar.

The *Amateur Radio and the Blind* series by Butch Bussen, WAØVJR, was published in *QST* between October 1987 and January 1988. These articles contain alternative solutions for the visually handicapped using hardware speech synthesis to pass logging information to the user. Most of the described solutions rely on expensive hardware but may still be appropriate for some visually impaired individuals. The entire series is available from the ARRL *QEX* Web page (see Note 1).

Conclusion

TalkingLogBook is an ongoing project to develop free software for persons with disabilities to facilitate Amateur Radio activities. Artificial speech makes logging software easier to use for some radio amateurs. Comments and suggestions for improvement of the software are appreciated. Drop WB5KIA a line if you can help with the development of the program, especially if you have Delphi programming skills. Check **www.qsl.net/wb5kia** for program updates and additional information.

Acknowledgments

Significant contributions to this project came from the work of Alex Shovkoplyas, VE3NEA; Dave Cook, WAØTTN, and Griffin Solutions, Inc (**www.grifsolu.com**). Their helper software and programming tools facilitated development. Tom Repstad, K1VG, provided coding for the program's WAS and DXCC lists and provided lots of ideas as to how to use the *Halcyon* database programming tool. Joe White, K4RYH, provided suggestions for an early version of the logbook.

Steve Gradijan is a geoscience consultant in the Dallas area. He holds an Amateur Extra class license, WB5KIA, and has BS and MS degrees in geology.

Uniform Current Loop Radiators

Here is a small loop antenna you can build using 300-Ω twin-lead for the antenna wire.

Robert K. Zimmerman, Jr, NP4B

Uniform current loop radiators (magnetic loops) have been used by amateurs in the small-loop limit, where the circumference is less than λ/3. This requires great care in construction and use: Copper pipe must be used along with vacuum variable capacitors and motorized tuning (due to high Q).

The loops described in this paper bear little resemblance to the above because:

1. Special resonant cable is used so that the loop has no reactance at its design frequency and no need for a lumped resonating capacitor.
2. The impedance is 50 Ω without a matching transformer.
3. The Q is low, about 46, so that the bandwidth is sufficient for an entire ham band.
4. The gain of the loop, operating as a two-element array, is better than that of a half-wave dipole. The loop radiator is horizontally polarized with no radiation at zenith.

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Canada
zimmo2@juno.com

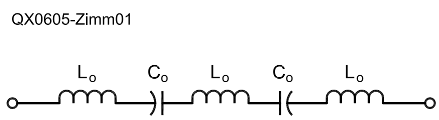


Figure 1 — Wire with periodic capacitive loading to cancel the inductive reactance of the wire at a given frequency.

We will first discuss the resonant cable, since it makes such loops possible. Then a 50-Ω loop plan is outlined. Finally, detailed construction plans are presented by means of photographs. All the theory used in this paper is contained in References 1 and 2, available free from the author.[1]

Resonant Cable

All wire has inductance. It is this natural inductance that causes a sinusoidal current distribution on regular antennas, such as dipoles. For wire antennas longer than half a wavelength, there will be current reversals (nodes) every half wavelength.

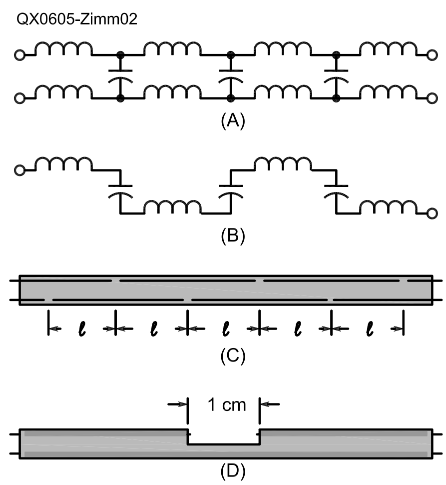


Figure 2 — (A) Ladder network representation of balanced transmission line (ribbon cable). (B) Network representation of transmission line modified by cutting opposite conductors every length *ℓ*. (C) Pictorial view of modified line. (D) Photograph of a modified line, where one wire of the twin-lead has been severed.

Table 1
Antenna Dimensions in Meters Versus the Operating Frequency in MHz for an Effective Velocity Factor of v_f = 0.9364.

Frequency (MHz)	*Wavelength (m)*	*Section (m)*	*Circumference (m)*	*Side (m)*	*Strut (m)*
14.15	21.187	2.232	15.623	3.906	2.762
21.2	14.141	1.490	10.428	2.607	1.843
28.2	10.631	1.120	7.839	1.960	1.386
50.2	5.972	0.629	4.404	1.101	0.778
144.2	2.079	0.219	1.533	0.383	0.271

To obtain a uniform current distribution on a wire, the natural inductive reactance of the wire must be cancelled with periodic capacitive loading, that is, series loading. This concept is shown in Figure 1. Such loading must be used more frequently than every quarter wavelength to perform its reactance canceling job effectively.

Here we make zero-reactance cable from 300-Ω twin-lead. As an equivalent circuit, twin-lead cable may be modeled as in Figure 2A. If we sever a single conductor at every length ℓ (on opposite sides of the twin-lead), we have effectively produced the cable in Figure 2B. This is exactly the desired circuit in Figure 1. Of course we must choose the length ℓ in accordance with the operating wavelength λ we desire to use. The formula is:

$$\ell = v_f \cdot 0.1125 \cdot \lambda \qquad \text{(Eq 1)}$$

where v_f is the cable velocity factor and λ is the wavelength. For the RadioShack foam core twinlead,[2] $v_f = 0.9364$. The wavelength in meters is:

$$\lambda\,(\text{meters}) = \frac{299.7925}{f\ (MHz)} \qquad \text{(Eq 2)}$$

The plan for cutting the cable is shown in Figure 2C. The photograph in Figure 2D shows a cable modified for zero reactance.

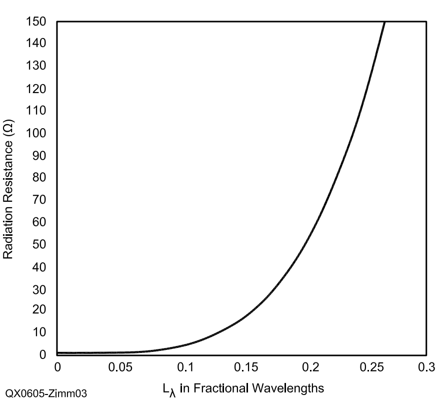


Figure 3 — Radiation resistance of a square uniform current loop versus the normalized side dimension L_λ in wavelengths.

Uniform Current Loops

Suppose we make square loops. We express the side length of the loop in fractional wavelengths as:

$$L_\lambda = \frac{Side\ (\text{meters})}{\lambda\ (\text{meters})} \qquad \text{(Eq 3)}$$

Figure 3 gives the radiation resistance of a uniform current loop versus L_λ. We see that the resistance is near 50 Ω for $L_\lambda \approx 0.2$. We already know the length of a resonant section from Equation 1. If we use 7 sections to construct our loop, then L_λ will be close to 0.2 and the loop will provide a radiation resistance of 50 Ω. Table 1 lists the dimensions (meters) for RadioShack 300 Ω foam-core twin-lead for the 14-MHz through the 144-MHz amateur bands. Table 2 lists the same information in feet. The Appendix contains the equations used to prepare Tables 1 and 2 so that one may program a spreadsheet for design work.

Figure 4 shows the loop gain modeled on EZNEC with the loop mounted horizontally ¼ λ above real earth (high accuracy, conductivity = 0.005 S/m, relative permittivity = 13) for $L_\lambda = 0.2$. The main lobe of the loop is +4.46 dBi at a takeoff angle of 37.0°. The EZNEC model is a symbolic model, because it contains 4 resonating capacitors, one on

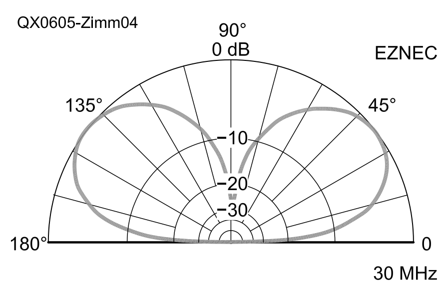
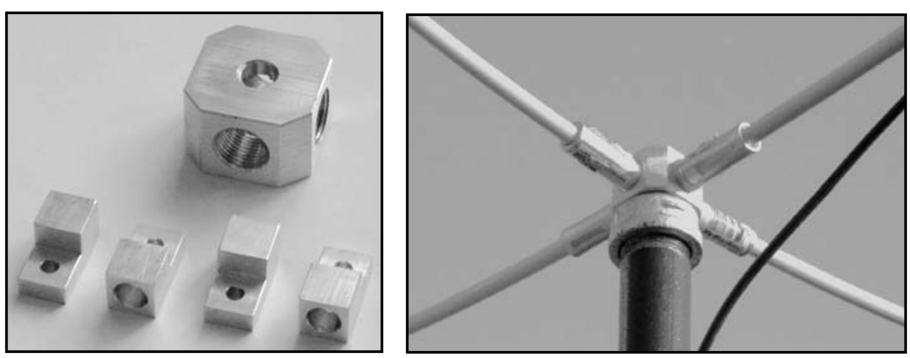


Figure 4 — EZNEC modeled antenna pattern for uniform current loop radiator with L_λ = 0.200 over real earth (high accuracy, conductivity = 0.005 S/m, ε_r = 13). The main lobe gain is 4.9 dBi at a takeoff angle of 37.0°.

Figure 5 — Completed 15-m loop radiator.



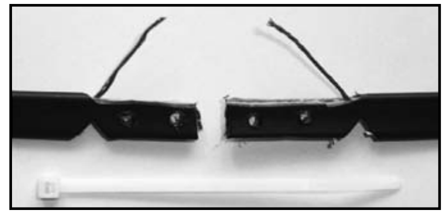
(A) **(B)**

Figure 6 — (A) Threaded aluminum-hub bracket and loop-support brackets as supplied by the author. (B) Hub bracket as installed on the antenna. The coaxial feed line crosses the photo at the right.

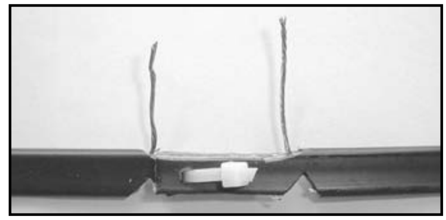
Table 2
Antenna Dimensions in Feet Versus Operating Frequency in MHz for an Effective Velocity Factor of v_f = 0.9364.

Frequency (MHz)	*Wavelength (ft)*	*Section (ft)*	*Circumference (ft)*	*Side (ft)*	*Strut (ft)*
14.15	69.514	7.323	51.260	12.815	9.062
21.2	46.397	4.888	34.214	8.553	6.048
28.2	34.880	3.674	25.721	6.430	4.547
50.2	19.594	2.064	14.449	3.612	2.554
144.2	6.821	0.719	5.030	1.258	0.889

Figure 7 — This photograph shows the connection of the resonant cable loop to the end of the strut.



(A)



(B)

(C)

Figure 8 — (A) Preparation of the cable ends for attachment. (B) Cable ends after attachment, with tie wrap. (C) Connection of loop to 50-Ω coaxial feed line (any length).

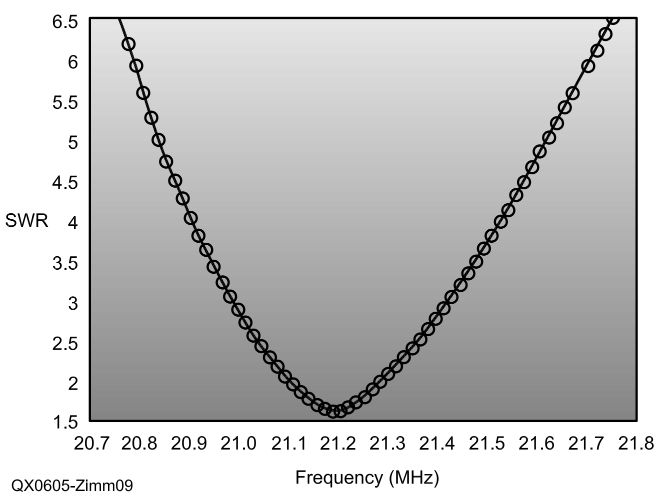


Figure 9 — Experimentally measured SWR curve for 15-m loop.

each side of the loop. The real antenna contains 7 resonant sections, and will have an even smoother uniform current.[3]

Construction

Fiberglass rods that screw together are available in major hardware stores as chimney sweep rods.[4] These rods are excellent for use as antenna struts. Figure 5 is a photograph of a 15-m loop made with eight fiberglass rods as four spreader struts. The center hub, shown in Figure 6, is a threaded aluminum bracket available from the author.[5] The twin-lead is attached to the ends of the struts as detailed in Figure 7, using nylon tie wraps and four aluminum brackets provided by the author along with the hub bracket. Figure 8 shows the ends of the resonant cable fastened together with a tie wrap at the feed point. Since the resonant cable maintains a uniform current throughout its length (even at the ends), there is no opportunity for coaxial sheath currents and a balun is *not* required. Figure 9 is the measured SWR curve for the 15-m loop.

Making It Work

With any luck, and careful measurement of the resonant sections, your antenna will come out just where you want it: at the design frequency. Because of very minor variations in the velocity factor, v_f, from batch to batch, you might end up being only close (±200 kHz) instead of at your desired frequency. What to do? Fortunately you have several options to tweak your loop.

For the purist: You may calculate your own velocity factor and prepare another cable, at a cost of about $5.00 for the new wire. Here is how: Take the actual center frequency and calculate the operating wavelength with Equation 2. Then from Equation 1,

$$v_f = \frac{\ell}{0.1125\,\lambda} \qquad \text{(Eq 4)}$$

where ℓ is the resonant section length previously used. Use this new velocity factor to redesign your antenna. The second iteration must work.

For the more practical ham, you may move the center frequency in the following manner:

1. If you are below your desired center frequency, widen the gaps in your cable cuts. Go slowly, widening one gap at a time by one inch. The author's 15-m loop initially came out centered at 21.012 MHz, even though the design frequency was 21.200 MHz. Widening 4 gaps to 3 inches each brought the loop to the desired frequency.

2. If you are above your desired center frequency, place a small inductor in series with your feed line (at the feed point). Using the 15-m loop as an example, being 200 kHz high means you must cancel $-j50\ \Omega$

reactance at the desired frequency, which would be a series inductor of 0.37 µH.

After a rainstorm, the author noticed his 15-m loop was resonating 1 MHz low. After thinking about it, the reason seemed clear: the water (having a high permittivity) on the twin-lead had increased the capacitance of the cable and lowered the frequency. I wiped down the twin-lead with WD40 water displacement spray, and that seems to help. [A coating of automotive polish has also been reported to reduce the effects of rain and possible dirt build-up on twin-lead — *Ed.*]

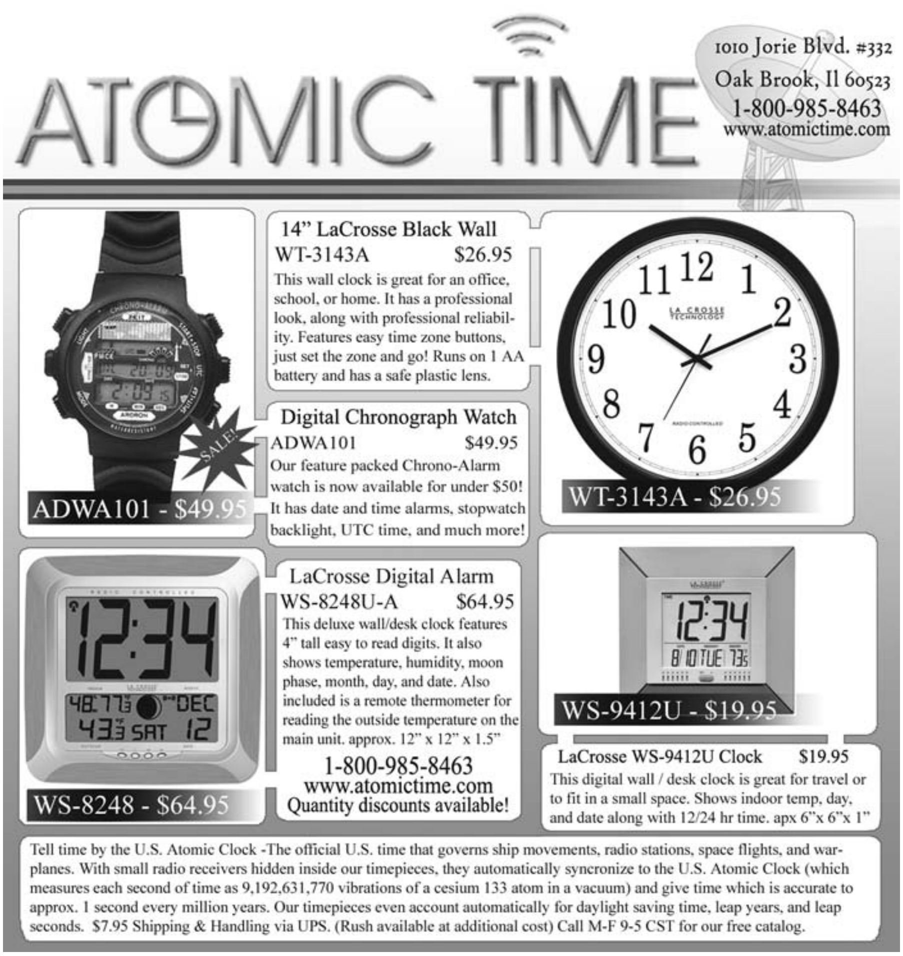
Summary

These antennas are easy to build, and will give satisfying performance well beyond that of a dipole or ¼-λ vertical with ground plane. The uniform current loop has a gain higher than a dipole and is smaller than a ¼-λ vertical. There are only two spots where you might make a mistake: (1) when you cut the cable every length ℓ, be sure it is on *opposite sides* of the twin-lead; (2) when you measure the cable sections, measure the length from the center of one cut to the center of the next cut — be accurate to within 2 mm. If you can handle a tape measure, you can successfully make this antenna.

The first contact with the 15-m beam was HK3JJH in Bogota, Colombia. Pedro gave me a 59+ constant signal (no QSB) for my 20-W output power from an SG-2020 mini-rig. Much good DX has followed.

Dedication

This paper is dedicated to the memory of radio engineer Jay D. Gooch, W9YRV (1922-2003), the University of Illinois research engineer who mentored the author during graduate school, and later, a lifelong friend.



References

[1]Robert K. Zimmerman, Jr, "Uniform Current Dipoles and Loops, Part 1 — Theory", *AntenneX*, April 2006. See **www.antennex.com**.

[2]Robert K. Zimmerman, Jr, "Uniform Current Dipoles and Loops, Part 2 — Practical Realization," *AntenneX*, May 2006.

Appendix

The formulas used to prepare Tables 1 and 2 are as follows:

$$\lambda \text{ (meters)} = \frac{299.7925}{f\,(\text{MHz})}$$

$$\text{Resonant Section Length} = \lambda \bullet 0.1125 \bullet v_f$$

$$\text{Side} = \frac{\text{Circumference}}{4}$$

$$\text{Strut Length} = \frac{\text{Side}}{\sqrt{2}}$$

To obtain values in feet, multiply all lengths by 3.28084.

Notes

[1]Contact the author at **np4b@arrl.net** and ask him to e-mail the article files to you.

[2]RadioShack "NEXTECH 300-Ω super-low-loss foam TV twin-lead cable," part number 150-1174.

[3]You can download the author's EZNEC model file from the ARRL Web at **www.arrl.org/qexfiles/**. Look for 5x06_Zimmerman.zip.

[4]Sold in Canada at Canadian Tire. Note that these rods are too flexible for use on a 20-m loop. Wooden struts (spruce) must be used instead.

[5]An aluminum center hub is available from the author (at cost) for $25 US postage paid.

Robert K. Zimmerman, Jr, NP4B, was born in 1951 in Dupo, Illinois where he grew up. He graduated from Southern Illinois University with degrees in physics (BS, MS) (1973, 1975) and then attended the University of Illinois (MSEE, 1980). He has spent his entire career in radio science, working for Cornell University (Arecibo Observatory), NASA Goddard Spaceflight Center, Los Alamos National Laboratory (accelerator division), and most recently as a radar engineer on Kwajalein Atoll. He is an avid radio amateur and microwave experimenter. □□

Tech Notes

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A simple method for measuring the heights of objects using the trigonometry discovered by the ancients.

Measuring Height With a Poor Man's Gizmo

How high is that pole? Do I need to rent a boom hoist? Do I need to contact the FAA because the new property that I purchased has an antenna tower that may exceed 200 feet in height?

Those are not very common situations, but they do occur. A poor man's height-measuring gizmo was constructed to satisfy this need.[1] The principle of operation utilizes a law of plane geometry that we studied in secondary school: corresponding sides of similar triangles have the same ratios.

While many historians believe that the ancient Egyptians knew about the principle, Thales of Melitus (*circa* 645 to 534 BC) is universally credited with its first practical use in measuring heights and distances.

Similar Triangles

In Figure 1, note that $a/b = A/B$. If we know a, b and A, then:

$$B = A\frac{b}{a} \quad \text{(Eq 1)}$$

For example, if $a = 3$, $b = 2$ and $A = 60$, then $B = 40$.

Isosceles Triangles: Refer to Figure 2. For the instance where $a = b$, then:

$$B = A \quad \text{(Eq 2)}$$

The interior angles of the triangle are 45°. For example, suppose that we made a small triangle out of cardboard where $a = b = 1$ foot. If we moved the triangle away from the base of the object whose height we desire, and sight along the triangle's hypotenuse such that the top of the object whose height we desired to measure were aligned with our triangle's hypotenuse, then the horizontal distance to our sample 1-foot triangle would be the height of the object. If our sample triangle were 60 feet from the object's base, then the unknown object's height would be 60 feet.

Of course, placing the cardboard triangle on the ground and then sighting along its hypotenuse may be an impossible task, or may require us to dig a hole in the ground to place our eyes at ground level. For this reason, I designed and constructed the gizmo shown in Figures 3 and 4. We must add the gizmo height to the object height measured by sighting with the gizmo, as shown by Figure 5.

[1]Notes appear on page 53.

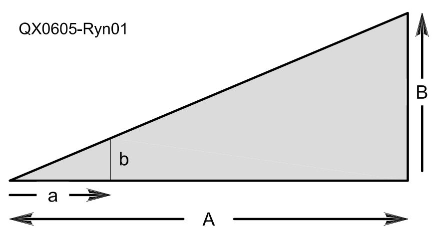

Figure 1 — Similar triangles.

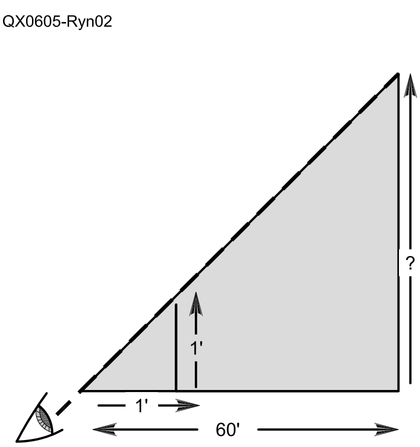

Figure 2 — Sighting along small and large triangles.

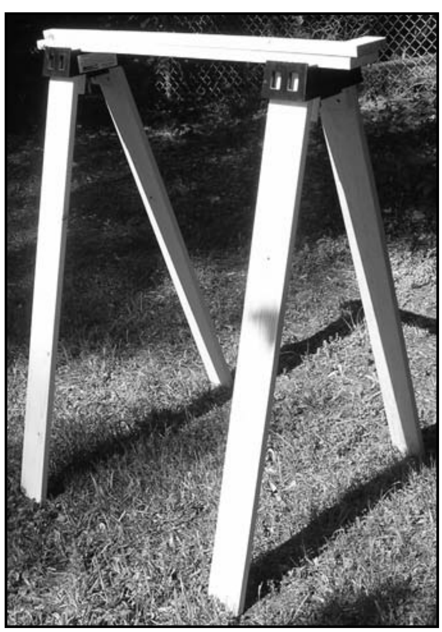

Figure 3 — This photo shows the gizmo folded flat against the base for storage or transport.

Another correction to the measured height must include any difference in ground level between the object location and gizmo location. See Figure 6.

If for some reason, the gizmo must be located farther away from the object than where Equation 2 is relevant (that is, when the object height is equal to the distance between the gizmo and the object), then we may use Equation 1 to determine the object height.

Suppose we wish to measure the height of an antenna that is located on a ground rise. See Figure 7. Move the gizmo away from the antenna base until we view the top of the gizmo aligned with the top of the antenna tower. We can now use our 100-foot tape measure to determine the value of A. Let's assume that we measured 208 feet. From Equation 2, we now know that $B = A$ = 208 feet.

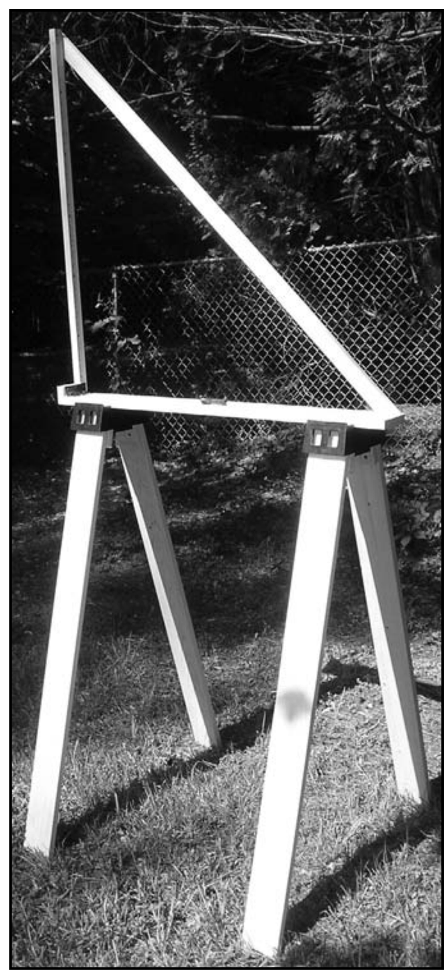

Figure 4 — This photo shows the gizmo opened ready to measure the height of a tree, tower or other object.

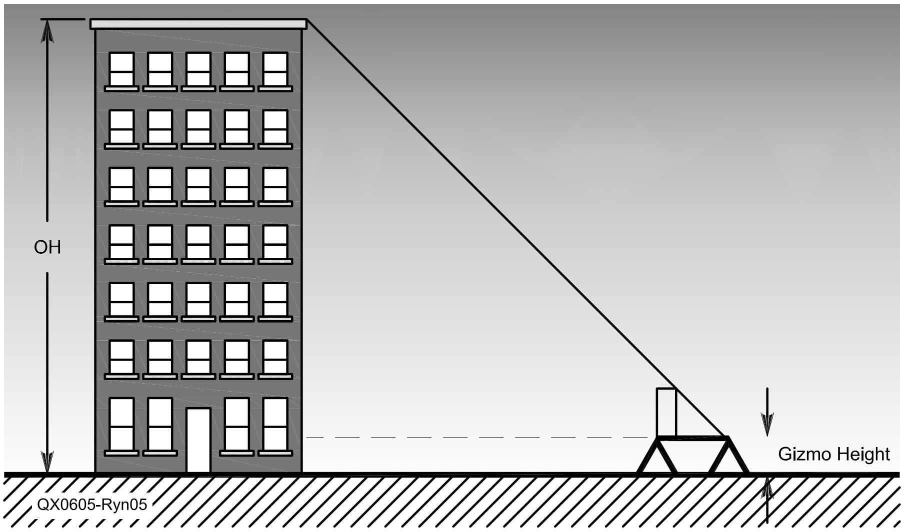

Figure 5 — This drawing illustrates how the gizmo can be used to measure the height of a building.

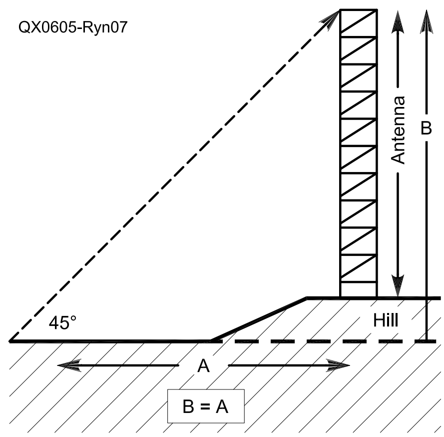

Figure 7 — This drawing shows the set-up to measure the height of a tower or antenna that is on a hill.

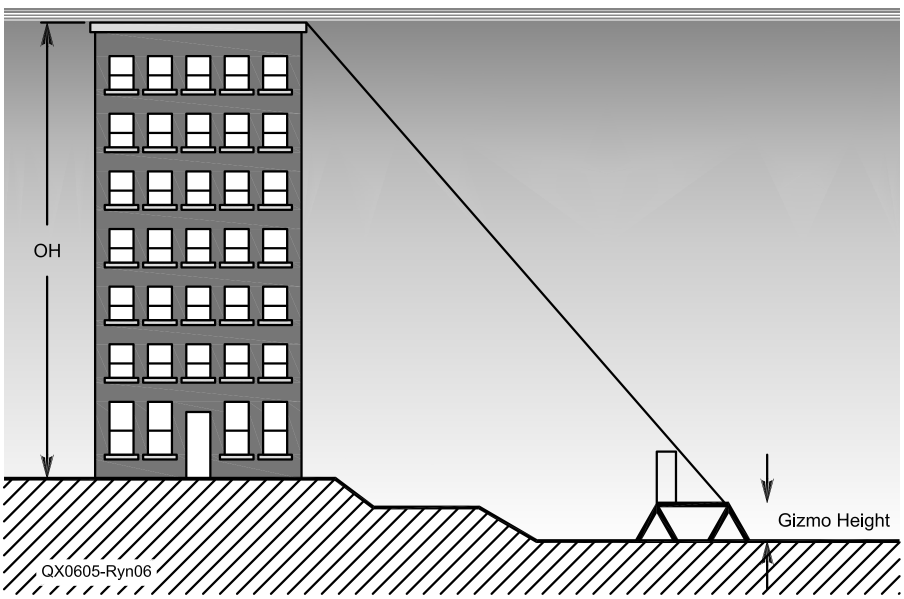

Figure 6 — Here, the gizmo is being used to measure the height of a building on uneven ground.

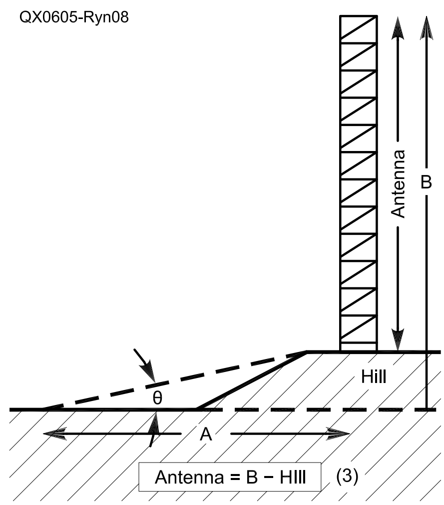

Figure 8 — After you measure the height of the tower, you will also have to measure the rise of the hill, and add that value to the measured height.

We now sight the top of the hill and read that value from the tape measure on the vertical member of the gizmo and note that value to be 0.173 feet. Employing Equation 1, we determine the height of the hill:

$$Hill = A\frac{b}{a} = 208 \text{ feet}\frac{0.173 \text{ feet}}{3 \text{ feet}} = 12 \text{ feet}$$

We may then use this value to determine the height of the antenna above ground level (Figure 8):

$Antenna = B - Hill = 208$ feet $- 12$ feet $= 196$ feet

Nighttime Height Measurement

If performing the object height measurement during daylight hours is impossible, as in a working environment where heavy equipment is transiting the area, then a laser with an adjustable angle between the laser base and laser tube may be used in place of sighting along the small triangle hypotenuse. Harbor Freight Tools sells one of these items for $14.95.[2]

The laser level is adjusted such that the laser beam skims one inch (space between the laser level surface to laser lens) over the top of the vertical member. This level employs a red laser. Readily available red lasers are not as viewable as green lasers, but are considerably less expensive.

Note: *Never look at a laser beam when pointed at you! The FAA is concerned with the effects of high-power lasers potentially blinding pilots. Pointing a powerful laser up into the night sky could be seen as an attempt to affect the flight of an airplane. Use caution with this technique!* Manufacturers of laser levels used in construction work supply lasers that have power outputs of less than 5 mW. The manufacturers of these devices are typically controlled by OSHA and FDA restrictions. Most laser levels have a limited range of 1,000 feet or less.

Parts and Construction

See Figures 9 and 10. The 10-foot measuring tape, Starrett Company model C12-10T is divided into 1/10 foot increments, and must be cut to a desired length (3 feet or 4 feet). If your particular needs may be satisfied by always employing a 45° sighting elevation angle, such that $a = b = 3$ feet (or 4 feet), the Starrett measuring tape may be omitted from the gizmo.

Horizontal member, 1 × 4 × 36 inches (actual dimensions = $^3/_4 \times 3^1/_2 \times 36$ inches).

Vertical member, 1 × 4 × $36^3/_4$ inches (actual dimensions = $^3/_4 \times 3^1/_2 \times 36^3/_4$ inches).

Diagonal member, 1 × 2 × 52 inches (actual dimensions = $^3/_4 \times ^1/_2 \times 52$ inches).

Three-inch-long hinge (minimum play), Stanley 81-9102.

Sawhorse brackets/leg braces, 2 pairs, Harbor Freight Tools Item 46313.

Legs, 1 × 4 × 48 inches (actual dimensions: $^3/_4 \times 3^1/_2 \times 48$ inches).

Measuring Tape, Starrett C12-10T, cut to 36 inches or 48 inches (Optional).

Laser Level, Slope adjustable, Harbor Freight Tools Item 92674 (Optional).

Bubble string level, Home Depot, Empire Item 83-3.

Screws, paint, etcetera.

Ancient History Behind the Invention

Thales of Melitus, one of the so-called "Seven Sages of Ancient Greece," lived over 2500 years ago (*circa* 600 BC). His wide-ranging theories of science and philosophy come to us mainly through the writings of others, since virtually none of his original documents survive. Some even say he never wrote anything at all, but what we do know about him clearly places him at the leading edge of scientific thought in his time.

Among other things, Thales was a geometer. He developed the basic theory behind Dr. Rynone's gizmo, although some historians give the credit for its discovery to the Egyptians (*Rhind papyrus*, *circa* 1700 BC, problem 57). It is that pairs of corresponding sides of similar triangles have a common (fixed) ratio.

The Egyptians evidently knew that when the Sun's elevation was 45°, the shadow cast by an object onto level ground was equal to the object's height. Thus, it was easy for them to measure the heights of all those huge pyramids and obelisks they erected. As reported by Eudemus (*circa* 300 BC), Thales used the principal to measure the distance to ships at sea. He allegedly built an adjustable wooden triangle on a tall hill. With a knowledge of its altitude, a sighting along one of the legs of the triangle and a measurement of its length yielded the distance to the ship.

Thales also gets credit for discovering other basic trigonometric theorems, such as that a circle is bisected by its diameter. That kind of thing may seem obvious to us today; but Thales used them formally to develop a lot of useful applications — *Doug Smith, KF6DX,* QEX *Editor*.

Figure 9 — Here a small laser level is attached to the sighting angle of the gizmo for possible nighttime measurements.

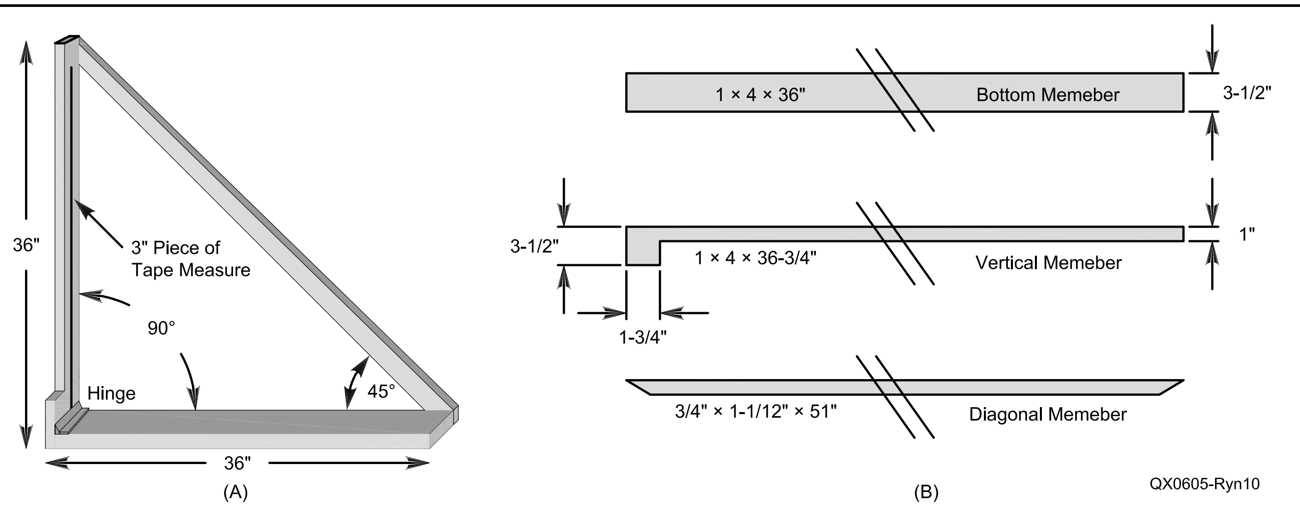

Figure 10 — This drawing shows the parts as well as the construction of the gizmo, with a foldable sighting mechanism.

On Measuring Height Like a Boy Scout

An article in *QST's* The Doctor is In column as well as this *QEX* article have described methods of measuring the height of a tree, tower or building. The "gizmo" described in this article will certainly do the job quite nicely. Still, there are quite a few things that can affect the accuracy of your measurements. For example, while no mention was made of setting the gizmo so the base is level, that is an important consideration. The construction of the gizmo is also critical. Even a degree or two of error in constructing the 45° right triangle can adversely affect your results.

When I was a young Boy Scout I learned a couple of methods of estimating the height of a tree. Scouts still learn those methods today. They are simple, requiring no special equipment except a short straight stick, and are reasonably accurate if performed with care. They can be used on uneven ground without any additional measurements or calculations. In fact, the only calculations that are required involve a simple multiplication.

One method the Scouts learn is called "The Felling Method." This method will probably result in a measurement that is at least as accurate as the gizmo described in the article.

You will need a straight stick, with a length that depends on the height of the object you want to measure, but a foot or so should be long enough for just about any measurement you would want to make. You do not have to know the length of the stick, however. A cooperative helper will make things easier, although it is certainly possible to use this method by yourself. Step back some reasonable (unmeasured) distance from the tree. (I usually like to stand approximately the same distance as the height of the tree, but it really doesn't matter.) Holding the stick at arm's length in front of you, close one eye and sight over the stick at the tree. Position the top of the stick so it appears to touch the top of the tree and then position your thumb along the stick so it appears to touch the ground at the base of the tree. See Figure A. Now rotate your wrist so the stick is horizontal, along the ground. Keeping your thumb at the base of the tree, have your helper move so he or she is standing where the top of the stick now touches the ground. Mark this spot. The most critical part of this measurement technique is to ensure that the line from you to the tree and then to your helper forms a 90° angle along the ground. A Scout would now pace off the distance between the mark and the tree, counting his steps. Then he would multiply by the length of his stride and give you the height of the tree with a fair degree of accuracy. (Hint: If your step is two feet and it takes 50 steps to cover the distance between the mark and the tree, you have a 100-foot tall tree!)

If you want to determine the length of your stride, measure off a 100-foot course and walk it several times, counting your steps. Take normal steps. Don't think about trying to make your step a certain length, like a football referee stepping off a 5-yard penalty. Just walk normally. If you get the same number of steps after walking the 100-foot course 3 or 4 times, you have probably walked normally. Now just divide 100 by the number of steps you took. Again, if you took 50 steps, your pace is 2 feet per step.

If you need even better accuracy for your measurement, stretch a tape measure along the ground and measure the distance. If you were reasonably careful to hold the stick at arm's length, mark the top and bottom of the tree with the length of the stick in your sight line, and maintained a 90° angle between you, the tree and your helper marking the spot on the ground, you should have the height of the tree to within a few feet. I've used this method on several occasions when cutting down a tree. I wanted to be sure there was enough space between the base of the tree and a fence or other obstacle that I didn't want the tree to hit. In every case, I was within a foot or two of the spot where the top of the tree hit when it landed.

Another method the Scouts might use is called "The Stick Method." Start with someone of known height. (It is easiest if you can find someone who is five feet tall, although four footers or six footers will work equally well.) Have this person stand at the base of the tree. Well, okay, if you can't find anyone to help with the task, measure 5 feet up the trunk of the tree and tie a ribbon or piece of rope around the tree. You could also just stand a board, post or other straight object of known length at the base of the tree. Then back up some reasonable distance (that does not have to be measured) and hold a short, straight stick at arm's length in front of you. Close one eye and sight along the top of the stick, moving your arm so the top of the stick is even with the top of your helper's head or aligned with the ribbon or other marker on the tree. Place your thumb at a spot on the stick that aligns with the base

QX0605-RynA

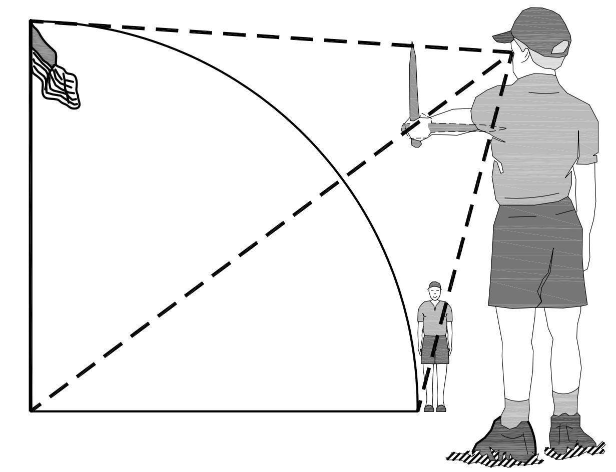

Figure A — The Felling Method of estimating heights involves transferring the height to a distance along the ground and then measuring that distance.

QX0605-RynB

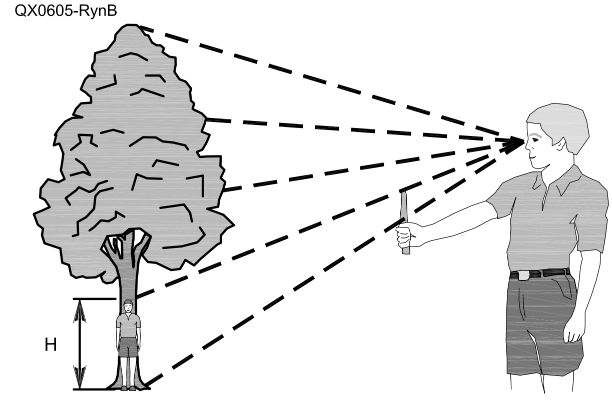

Figure B — The Stick Method of estimating heights involves counting the number of times a known height will fit into the total height of the object.

of the tree. Now simply move your arm with the stick up until your thumb aligns with the top of the helper's head and note where the top of the stick seems to touch the tree. Move your arm up again until your thumb touches the new spot and again note where the top of the stick seems to touch. Continue this procedure until you reach the top of the tree. See Figure B. To estimate the height of the tree, simply multiply your helper's height by the number of times you moved the stick upward. If you started with a 5-foot helper and measured the tree to be ten "stick lengths" then you have a 50-foot tree.

This method will not be as accurate as the tree felling method. As you move your sight stick up the tree, you will probably not hit the exact spot where the top of the stick came the last time. Your measurement could be several feet off in the height of a tree that is 50 feet or so high. If you want to know if the tree is tall enough to hang your 20-meter dipole at least ¼ wavelength above ground, it is perfectly adequate, though.

Either of these methods work at any time, whether or not the sun is shining. You may need a powerful flashlight to make the measurements at night, but certainly they will work on rainy, snowy and cloudy days when there is no sunlight to cast a shadow.

There are similar methods for estimating distances, such as the width of a stream or river, although those become a bit more elaborate. Ask your friendly Scouts to demonstrate that technique.

There are all kinds of "field measurements" that you can make with simple equipment. Often that simple equipment will provide results that are more accurate than you can make with some more sophisticated equipment. For example, if you have a short level, such as a 6-inch or 12-inch level and want to ensure that a post is set vertically (plumb) the level may not be the best tool to use. Tie a weight onto the end of a string to make a plumb line. Step back a few feet and hold the string at arm's length. Use 3 or 4 feet of string, and sight along the string at the edge of the post. You will quickly be able to see when the post is plumb because the string will match the edge of the post.

— Larry Wolfgang, WR1B

Accuracy

The object height measurement errors will be determined by length errors for the gizmo parts and sighting errors. A one-degree misalignment in a level will result in an approximate height measuring error of 3.5%. Thus, while measuring the height of a 200-foot tall object, a height error of ± 7 feet may occur.

Acknowledgments

Mr Oscar Ramsey constructed the gizmo and assisted with the testing. Ms Michele Danoff of Graphics By Design, assisted with sketches. Thanks to Mr Duard Wilson, VP Engineering, C S T Berger Company for technical data.

Notes

[1]Two articles dedicated to height measurement using elementary plane geometry and plane trigonometry appeared in the *QST* "The Doctor is In" column, Jun 2004, p 56 and Aug 2004, p 53.

[2]**www.harborfreight.com**, ITEM 92674-0VGA.

Upcoming Conferences

40th Anniversary CSVHFS Conference July 28-29, 2006
Bloomington, MN
Ramada Mall of America
(formerly the Thunderbird, across from Mall of America)

Call for Papers

The Central States VHF Society is soliciting papers, presentations, and Poster displays for the 40th Annual CSVHFS Conference on July 28 to 29, 2006. Papers, presentations, and posters on all aspects of weak-signal VHF and above Amateur Radio are requested. You do not need to attend the conference, nor present your paper, to have it published in the Proceedings. Posters will be displayed during the two days of the Conference.

Topics of interest include (but are not limited to)**:**

Antennas including modeling/design, arrays and control

Test Equipment including homebrew, using and making measurements

Construction of equipment, such as transmitters, receivers and transverters

Operating including contesting, roving & DXpeditions

RF power amps including single band and multiband vacuum tube and solid-state

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

Pre-amplifiers (low noise)

Digitial Modes WSJT, JT65, etc

Regulatory topics

EME

Software-defined Radio (SDR)

Digital Signal Processing (DSP)

Generally, topics not related to weak signal VHF, such as FM repeaters and packet radio, are not accepted for presentation or publication. However, there are always exceptions. Please contact either the Technical Program Chairman, Jon Platt, WØZQ, or the Proceedings Chairman, Donn Baker, WA2VOI/Ø at the email addresses below.

Deadline for submissions: For the Proceedings: Monday, 1 May 2006.

For Presentations to be delivered at the conference: Monday, 3 July 2006.

For notifying us that you have posters to display at the conference: Monday, 3 July 2006. Bring your poster with you on July 27.

Further information is available at the CSVHFS web site (**www.csvhfs.org**). See "The 2006 Conference," "Guidance for Proceedings Authors," "Guidance for Presenters" and "Guidance for Table-top/Poster Displays."

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Antenna Options

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This month we'll look at the bandwidth of several Yagi and Quad antenna designs.

How Wide is Wide?

Unlike the question, "How high is up?" our question has an answer. In fact, the question has many answers. The first step is deciding what we are referring to by the word "wide." Since our subject is antennas, there are two possibilities: *beamwidth* and *bandwidth*. Let's choose the latter as the more intriguing. Let's further refine the expression with a qualifier and call the operative term *operating bandwidth*. That expression should give us room to get into trouble.

The Width of the US Amateur Bands

As background, we may look at the US amateur bands from 160 meters through 70 cm and define the bandwidth of each. One common way to arrive at the width of each band is to divide the width of the band from the lowest frequency to the highest by the band's center frequency — and multiply by 100 to arrive at a percentage. 20 meters is 350 kHz wide, with a center frequency of 14.175 MHz (or 14175 kHz to use constant units of measure). The result is 2.47%. Table 1 provides bandwidth values for each of the amateur bands within the indicated scope of our survey.

The bandwidth numbers are useful in some contexts. For example, a precisely scaled antenna from one band would have the same coverage in terms of the bandwidth on the new band. Precision scaling means scaling the element length(s), diameter(s), and spacing (if relevant). If we scale an antenna known to cover one of the wider bands for a narrower band, we can be sure that the antenna will cover the new band. If we begin with an antenna for a narrower band, we cannot be certain that a scaled version will cover a wider band.

Next, with a bit more trepidation, let's subdivide the range of bandwidths that emerged for the US amateur bands. The following categories have no validity in general antenna literature, but will be useful for discussion within these notes. If the bandwidth is less than 1%, we may refer to a narrow bandwidth. If the bandwidth is between 1% and 4%, we may refer to a medium bandwidth. In the table, we may call bandwidths greater than 4% wide bandwidths.

Our concern over bandwidth derives from a subset of the antennas that we typically use. For example, someone who uses a center-fed (or off-center-fed or end-fed) wire cares very little about bandwidth. He or she simply "dials in" the correct settings of a tuner for maximum effectiveness (usually meaning a good SWR match at the transceiver). The operator measures bandwidth in terms of how many times the tuner requires resetting as the frequency moves up or down one of the bands.

The bandwidth categories evolve from general expectations that we have for relatively standard Yagi antenna designs, where the elements are aluminum tubes in the upper HF range. A 0.5-inch element for a 10 meter Yagi would require a 2.0-inch diameter on 40 meters. The normal tapered-diameter schedules for full-size 40-meter elements in amateur installations virtually never approach this value. So practical antenna scaling may not meet the table's presumption of perfect antenna scaling. As a result, horizontal 40-meter antennas based on designs that easily cover 20 or 15 meters usually suffer declining performance or full failure to perform at one or the other 40-meter band edge.

A combination of practical material limitations and total bandwidth has established conventions for band utilization. We find the 80-meter band subdivided into the 80-meter CW band and the 75-meter phone band, each requiring separate antennas (or antennas with switched element lengths, loads, or other means of changing the range). On 160 meters, we find a relatively narrow DX window, with antennas designed to cover only that region. On the 10-meter band, antennas tend to cover the first 1 MHz of the band — and sometimes, only the first 800 kHz of the band. In the VHF and UHF region, we find many high performance narrow-band antennas for use only within specific small parts of the bands. FM repeater users generally expect to use relatively simple, omni-directional, vertical antennas. Hence, high performance horizontal antennas on 6 meters tend to cover only a half MHz at the low end of the band. The upper 3 MHz of the band call for ground-plane monopoles, J-poles, and a few collinear vertical designs.

How we handle the wider amateur bands thus has at least two dimensions: operator decision or preference on the one hand and design capability on the other. We shall be interested in what various antenna designs — especially parasitic beams — can achieve by way of operating bandwidth.

Operating Bandwidth

Before we look at antenna types that will cover the various amateur bands, let's probe the idea of operating bandwidth for a

Table 1
The Bandwidth of Many US Amateur Bands as a Percentage of the Center Frequency

Band	*Lowest Frequency (MHz)*	*Highest Frequency (MHz)*	*Width (MHz)*	*Center Frequency [2] (MHz)*	*Bandwidth (%)*
160 m	1.8	2.0	0.2	1.9	10.53
80 m	3.5	4.0	0.5	3.75	13.33
60 m[1]	5.33	5.408	0.078	5.369	1.45
40 m	7.0	7.3	0.3	7.15	4.20
30 m	10.1	10.15	0.05	10.125	0.49
20 m	14.0	14.35	0.35	14.175	2.47
17 m	18.068	18.168	0.1	18.118	0.55
15 m	21.0	21.45	0.45	21.225	2.12
12 m	24.89	24.99	0.1	24.94	0.40
10 m	28.0	29.7	1.7	28.85	5.89
6 m	50.0	54.0	4.0	52.0	7.69
2 m	144.0	148.0	4.0	146.0	2.74
1.25 m	222.0	225.0	3.0	223.5	1.34
70 cm	420.0	450.0	30.0	435.0	6.90

[1]The 60 meter band is channelized, so the limits shown are approximate, though useful for antenna planning activities.
[2]Center frequencies are the arithmetic band centers; geometric center frequencies are slightly lower for the wider bands.

parasitic beam. For many amateurs, the 50-Ω SWR seems to be the only factor involved in setting the operating bandwidth. For general-purpose communications, a 2:1 SWR maximum serves as a usable standard. Avid DXers and contest operators tend to prefer a 1.5:1 limit to prevent fold-back circuits in high power amplifiers from reducing or cutting off the power output.

There are numerous other considerations that go into a final judgment of an antenna's operating bandwidth. We can focus on just two aspects of antenna performance and enlarge our thinking considerably. How well the front-to-back ratio holds up across a given band contributes to many decisions about whether to use a particular antenna design. Amateurs tend to use a 20-dB standard for the front-to-back ratio of Yagis with at least three elements. The figure represents a minimum value that we expect to achieve at all frequencies within the band. (I shall pass over the question of which front-to-back figure to use: 180°, worst-case, or average front-to-rear ratio.)

Matters of gain can be even more complex. Much commercial antenna literature cites only a single number. It does not matter whether the number is the peak gain or the average gain, because neither figure tells us anything about the antenna's gain behavior across the intended passband. Only a detailed table of samples or a graph for the entire band will give us an adequate portrait of the gain performance. Very likely, these ideas will grow more meaningful as we look at some sample designs. The samples will emerge from my stock of antenna models. All make use of very standard techniques, even though the models themselves may not be in final form, ready for the home workshop. For example, some HF Yagis will use uniform-diameter elements rather than tapered-element schedules. All present very reasonable pictures of actual antenna performance, however.[1]

Example 1: A Four-Element Yagi and a Three-Element Quad Beam for 20 Meters

We may begin with a pair of contrasting parasitic beams for the medium width 20-meter band. The four-element Yagi uses a 437-inch boom, while the quad's boom is about 387 inches long. The Yagi uses 1-inch-diameter elements, while the quad uses AWG no. 12 wire elements. We expect most extant beam designs to cover all of 20 meters. Both the standard four-element Yagi and the three-element quad provide SWR values of less than 2:1 across the band, as shown in Figure 1. The quad uses a 75-Ω standard, while the Yagi uses a beta match to arrive at a 50-Ω SWR reference. Table 2 provides sampled data from the models across the band.

At mid-band, both antennas show a free-space forward gain of more than 8.55 dBi. The two antennas have very different gain curves, however, as revealed in Figure 2. Like most standard-design Yagis with at least one director, the four-element beam shows a rising gain characteristic as we increase frequency within the passband. In contrast, the quad shows a decreasing gain value as we increase frequency. I chose these particular models because they pass at the middle of the band. Hence, neither has any particular average gain advantage.

Still, we may concern ourselves with two aspects of these curves. First, we would like to know the total change in gain across the band. For the Yagi, the gain difference is just above 0.4 dB, while for the quad, the differ-

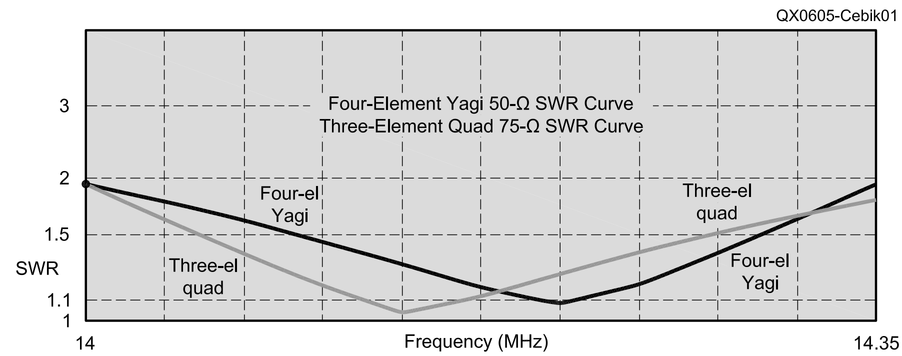


Figure 1 — SWR curves for a four-element Yagi and a three-element quad beam for 20 meters.

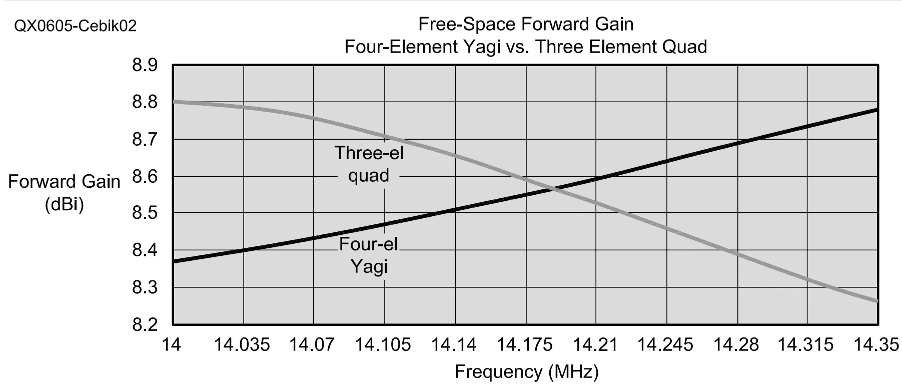


Figure 2 — Modeled gain performance for a four-element Yagi and a three-element quad beam for 20 meters.

Table 2
Comparative Modeled Performance Data for a Four-Element Yagi and a Three-Element Quad for 20 Meters

Four-Element Yagi with Beta Match

Parameter	14.0 MHz	14.175 MHz	14.35 MHz
Gain dBi	8.37	8.55	8.78
180° Front-Back dB	20.27	30.09	19.06
50-Ω SWR	1.94	1.17	1.94

Three-Element Quad Beam

Parameter	14.0 MHz	14.175 MHz	14.35 MHz
Gain dBi	8.80	8.59	8.26
180° Front-Back dB	14.69	27.76	14.79
75-Ω SWR	1.93	1.12	1.79

[1]Models used in this episode of "Antenna Options" are available in EZNEC format at the ARRL Web site. See **www.arrl.org/qexfiles/5x06_AD.zip**.

ence is a little over 0.5 dB. Only a particular operator with a good sense of what the desired operation requires can decide if these numbers are acceptable or not. Second, we may note where the highest gain values occur. Apart from other reasons for selecting one of the two antennas, the quad favors the CW end of the band, while the Yagi favors phone operation. Note that the questions we may usefully pose about antenna gain performance are simple, and the answers may be easy — if only we take the time to pose them.

The front-to-back performance of the two sample antenna designs appears in Figure 3. The Yagi just about meets the front-to-back standard of 20-dB minimum, using the 180° values across the band. However, the quad meets the standard only for about half the total bandwidth. The quad design used here emerged from a series of three-element quad designs expressly aiming for maximum operating bandwidth. As with most quad beams, the SWR bandwidth — when measured against our normal amateur standard — exceeds the front-to-back bandwidth by a considerable margin. Whether or not a particular operation needs a 20-dB minimum front-to-back ratio is once more an operator decision.

Example 2: A Five-Element Standard-Design Yagi and a Six-Element OWA Yagi for 28 to 29 MHz

When we design parasitic beams for 10 meters, we usually design them only for the first MHz of the band — from 28.0 to 29.0 MHz. The reduced passband still has a bandwidth of 3.51%, making it considerably wider than 20 or 15 meters. Many highly capable designs for 20 meters strain to achieve a good set of performance values across the most active part of 10 meters. Let's compare a five-element Yagi of standard design to a six-element OWA design. Both antennas use 0.5-inch elements. The five- element array has a 333-inch boom, partly because it increases the driver-reflector spacing to produce a near-50-Ω feed point impedance. The OWA design, pioneered by NW3Z and WA3FET, packs its elements onto a 288-inch boom.

The boom lengths and the number of elements call for a small comment. Ever since the appearance of the groundbreaking work of Jim Lawson, W2PV, a sound bite has pervaded Yagi articles: gain is a function of boom length rather than the number of elements. As true as this statement may be, it is no excuse for using too few elements to achieve all of the characteristics that one needs from a given Yagi design. As we shall see in the OWA design, extra elements may not increase gain over the longer-boom five-element Yagi, but they can shape the performance curves across the operating bandwidth of the antenna.

As shown in Figure 4, the five-element design achieves a very acceptable 50-Ω SWR curve. However, by judicious spacing between the reflector, the driver and the first director, the OWA SWR curve is superior and meets the most stringent fold-back circuit requirements. Table 3 provides sample numbers for the rest of the performance categories on which we are focusing. Note that the five-element design has a very good front-to-back ratio, but tails off in this category at the upper end of the 10-meter activity region.

More important than the front-to-back behavior is the shape of the gain curves that appear in Figure 5. Nothing in the five-element standard-design Yagi controls the steep gain increase with rising operating frequency. The total gain differential across the band is nearly a full dB. In contrast, the extra OWA element and the arrangement especially of the second and third directors place the gain peak within the passband. One result is more even gain across the band. The total gain range is a mere 0.23 dB.

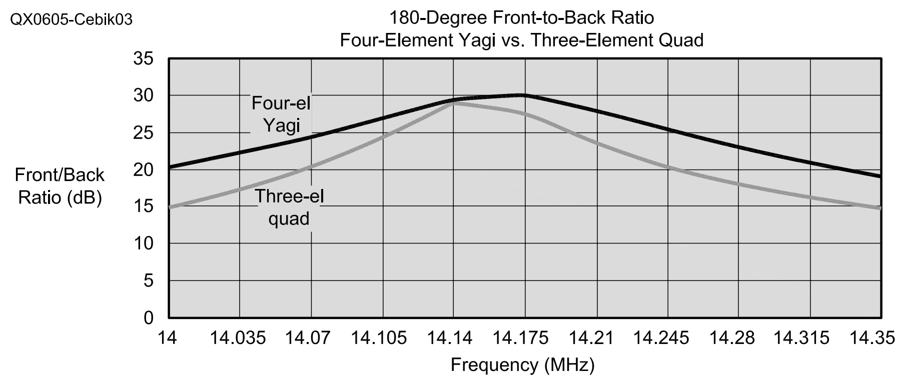


Figure 3 — Modeled 180° front-to-back performance for a four-element Yagi and a three-element quad beam for 20 meters.

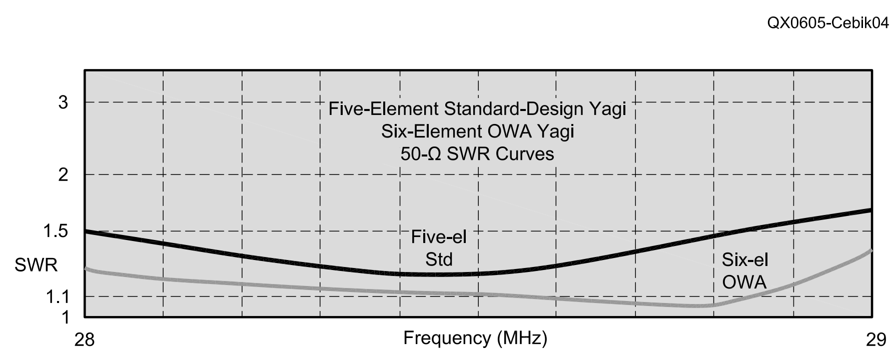


Figure 4 — SWR curves for a five-element standard-design Yagi and for a six-element OWA Yagi for 10 meters.

Table 3
Comparative Modeled Performance Data for a Five-Element Yagi and a Six-Element OWA Yagi for 28 to 29 MHz

Five-Element Standard-Design Yagi

Parameter	28.0 MHz	28.5 MHz	29.0 MHz
Gain dBi	9.80	10.32	10.74
180° Front-Back dB	25.15	23.74	14.09
50-Ω SWR	1.50	1.22	1.66

Six-Element OWA Yagi

Gain dBi	10.11	10.31	10.33
180° Front-Back dB	20.06	30.29	20.03
75-Ω SWR	1.24	1.11	1.36

Example 3: A Log-Cell Yagi for the Entire 10-Meter Band

The OWA design is only one of many techniques for increasing the operating bandwidth of a parasitic beam. Suppose that we wanted a beam that would cover the entirety of 10 meters, that is, cover nearly a 6% bandwidth. The OWA design would strain at the more than 60% increase in required bandwidth. However, a well-designed log-cell Yagi can handle the task with relative ease. Our sample uses a five-element log cell designed according to log-periodic-dipole array (LPDA) rules. It adds parasitic elements, namely, a reflector and a director. The resulting seven-element array uses a 337-inch boom with 0.75-inch diameter elements. Table 4 provides sample modeling results across the band. The band-edge 50-Ω SWR values mark the highest values for the array.

Figure 6 provides the gain and front-to-back curves for the antenna design. The peak in the front-to-back ratio may give the impression that the bulk of the band shows a poor ratio until we read the right-hand Y-axis and discover that the minimum front-to-back ratio is 27.15 dB. The use of a log cell as a driver does not generally enhance array gain relative to standard-design Yagis of the same boom length. However, it does provide strong control of the gain curve. As the graph shows, the antenna's peak gain occurs close to the band's center. As well, the gain changes by only 0.16 dB across the entire 1.7 MHz spread.

Example 4: A Short-Boom Three-Element Yagi with Phased Drivers for 6 Meters

We tend to call parasitic arrays with driver cells designed according to LPDA principles and equations "log-cell Yagis." However, we often use two or three phased driver elements that we arrive at by trial and error (or success, as the case may be). Such antennas often bear the label "phagi," in keeping with our penchant for snappy antenna names. Let's look at one example to see what a casually designed phased driver pair might be able to do to enlarge the operating bandwidth of an

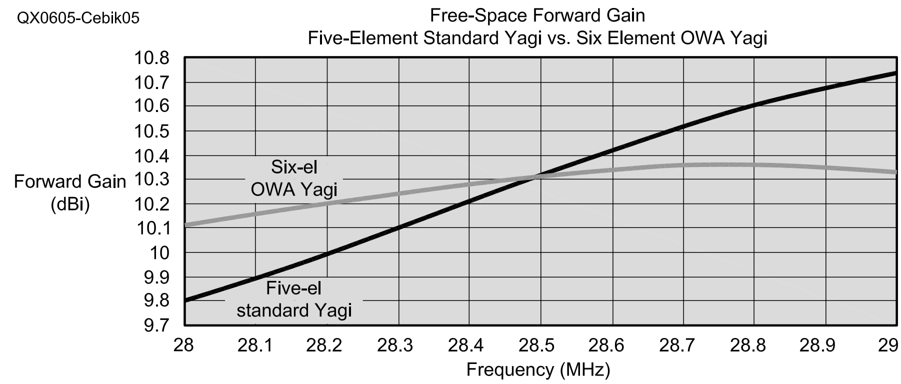


Figure 5 — Modeled gain performance for a five-element standard-design Yagi and for a six-element OWA Yagi for 10 meters.

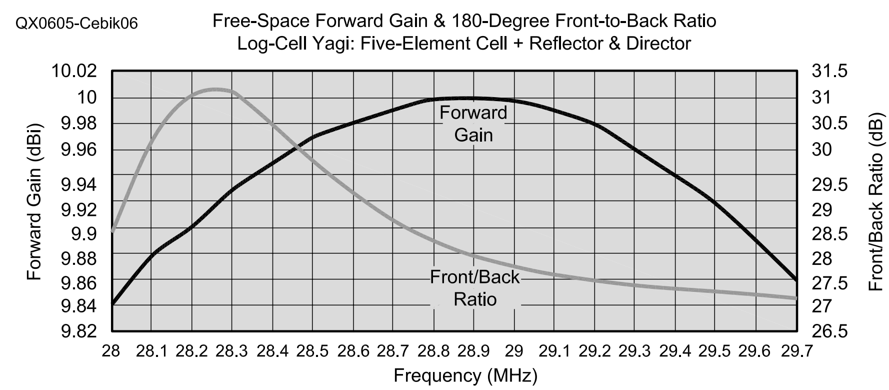


Figure 6 — Modeled gain performance for a 10-meter log-cell Yagi with a five-element log cell plus a reflector and a director.

Table 4
Modeled Performance Data for a 10-Meter Log-Cell Yagi with a Five-Element Log Cell Plus a Director and a Reflector

Parameter	28.0 MHz	28.85 MHz	29.7 MHz
Gain dBi	9.84	10.00	9.86
180° Front-Back dB	28.37	28.08	27.15
50-Ω SWR	1.52	1.28	1.49

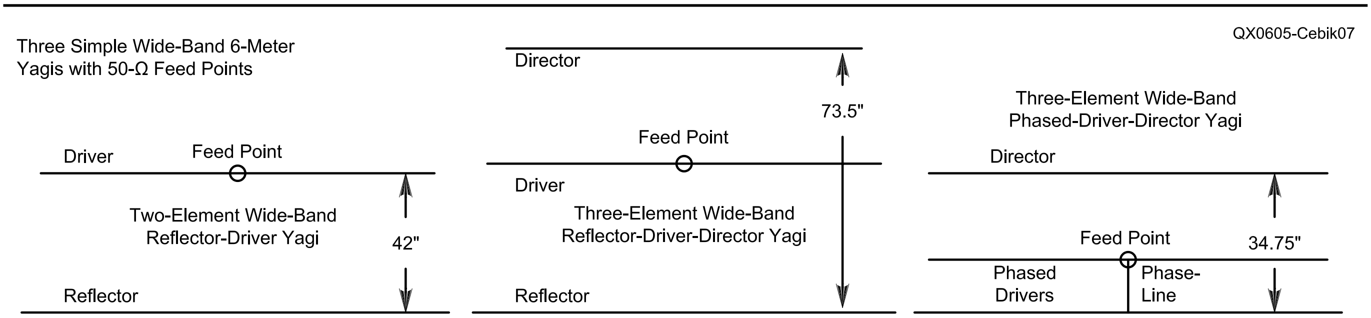


Figure 7 — Comparative outlines of three simple wide-band 6-meter Yagis with 50-Ω feed points.

antenna. Six meters is a good band for a compact, simple Yagi used vertically to provide some gain and wide directivity within the 3 MHz used for FM operations.

Standard-design wide-band Yagis that will cover all of 6 meters already exist. The two samples shown in Figure 7 are adaptations of 10-meter designs that Bill Orr, W6SAI, first presented in *Ham Radio* during the 1980s. The 3-element version is interesting because it uses a boom length that one might find in a higher-grain, narrower-bandwidth Yagi. However, as evident in the data in Table 5, the gain of the wideband Yagi is about a full dB lower — the price paid for increased operating bandwidth.

The smaller three-element Yagi with the phased pair of driver elements on the far right of Figure 7 is actually an extension of a two-element driver-director Yagi. This class of Yagi has a high peak gain but a very narrow beamwidth. I tend to recommend them for use on the narrow amateur bands, such as 30, 17, and 12 meters, although they have other specialized uses as well. The enlargement of the driver section of the antenna does not materially increase the boom length ahead of the drivers, and so the overall length is less than the boom needed by the two-element wide-band driver-reflector Yagi. All three Yagi designs use a tapered-diameter schedule consisting of 0.5-inch and 0.375-inch element sections.

The data table shows that the phased drivers do not quite equal the performance of the wide-band three-element Yagi. However, the performance curves are closer to that antenna than to the two-element curves. As well, the two-element gain shows its downward progression with rising frequency, in contrast to the nearly parallel upward gain curves of the three-element Yagis. Hence, for the intended application — FM simplex and repeater operation — the phase-fed dual driver of the sample short-boom Yagi seems a natural. Unlike either the OWA Yagi or the log-cell Yagi, the three-element phagi has too few elements to control all of the relevant operating parameters.

Example 5: An Eight-Element Utility Yagi for 70 cm

Almost as wide as 6 meters, the 70-cm band has a bandwidth of 6.9%. Very long boom Yagis, such as the DL6WU "trimming" series and other comparable designs, manage to cover the entire band with something to spare. The undulations of gain, front-to-back ratio, and SWR are fascinating to observe, but all three aspects of operation remain under relatively good control. Wide operating bandwidth is a boon to the home antenna builder, since limitations of precision tend not to void the basic performance of the antenna. The question that we might pose here is whether we can cover all of 70 cm with a relatively short utility antenna, say, using eight elements. In this frequency range, the boom length would be about 53 inches.

The project is quite feasible under two conditions. First, we should use fat elements, about 0.5 inch or so. Second, we should expect some variability of performance across the band. Other than that, we may use a fairly standard Yagi design. Figure 8 shows the 50-Ω SWR curve of our sample antenna, which reflects the application of OWA principles stretched about as far as I dare.

Sample data from across the band appear in Table 6. The gain and front-to-back curves are in Figure 9. The gain curve shows that it is possible to place the gain peak within the passband and to control the range of variation within about 0.5 dB. The gain level is close to what is standard for narrower-band Yagis with the same number of elements. However, over the wide operating bandwidth, we cannot fully control all facets of performance to the same level. The front-to-back curve dips to slightly less than 15 dB, which is adequate for the designated use as a utility beam for either horizontal or vertical installation.

Table 5
Comparative Modeled Performance Data for Three Simple Wide-Band Yagis for 6 Meters

Two-Element (standard-design) Yagi

Parameter	50 MHz	51 MHz	52 MHz	53 MHz	54 MHz
Gain dBi	6.47	6.06	5.70	5.40	5.16
180° Front-Back dB	9.65	10.66	10.30	9.49	8.67
50-Ω SWR	1.99	1.29	1.21	1.55	1.97
Three-Element (standard-design) Yagi					
Gain dBi	7.04	6.95	6.98	7.15	7.45
180° Front-Back dB	14.42	17.13	20.47	21.28	18.49
50-Ω SWR	1.55	1.22	1.05	1.36	1.98
Three-Element Yagi with Phased Drivers					
Gain dBi	5.87	6.19	6.55	6.93	7.30
180° Front-Back dB	13.13	15.84	18.59	17.35	12.66
50-Ω SWR	1.85	1.54	1.24	1.11	1.96

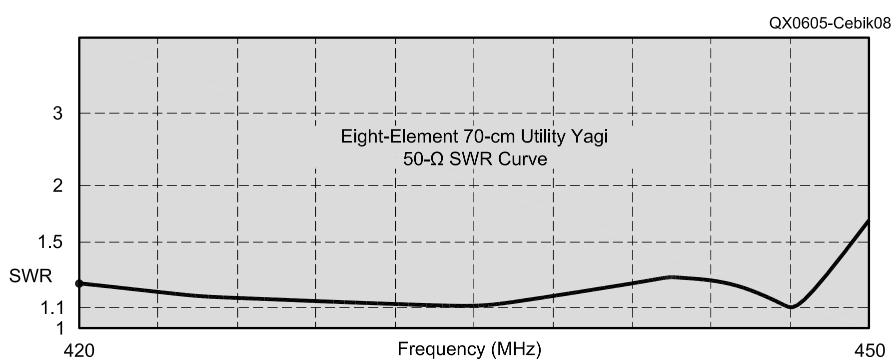


Figure 8 — 50-Ω SWR curve for a wide-band eight-element utility Yagi for 70 cm.

Table 6
Modeled Performance Data for a 70-cm Eight-Element Wide-Band Utility Yagi

Parameter	420 MHz	430 MHz	440 MHz	450 MHz
Gain dBi	12.92	13.31	13.59	12.99
180° Front-Back dB	16.27	14.95	19.61	20.91
50-Ω SWR	1.24	1.13	1.21	1.65

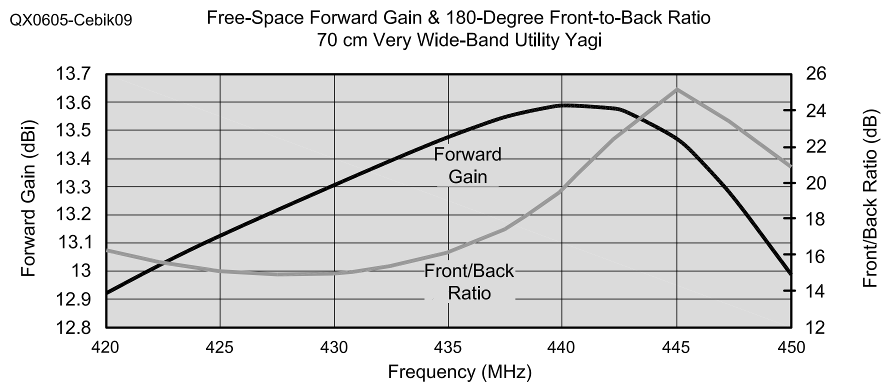


Figure 9 — Modeled free-space performance (gain and 180° front-to-back ratio) for a wide-band eight-element utility Yagi for 70 cm.

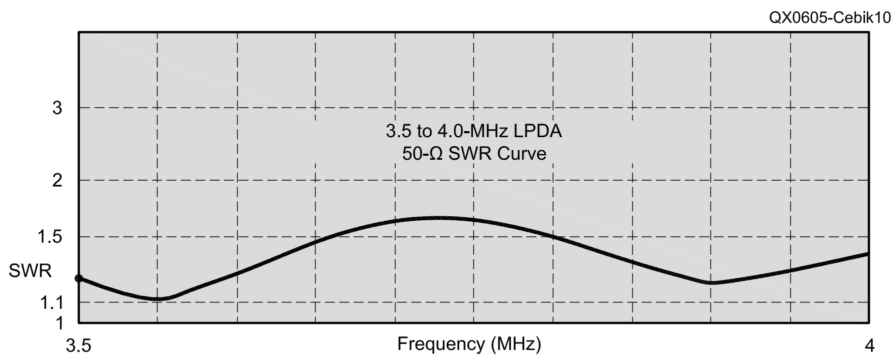


Figure 10 — 50-Ω SWR curves for an eight-element wire LPDA to cover 3.5 to 4 MHz.

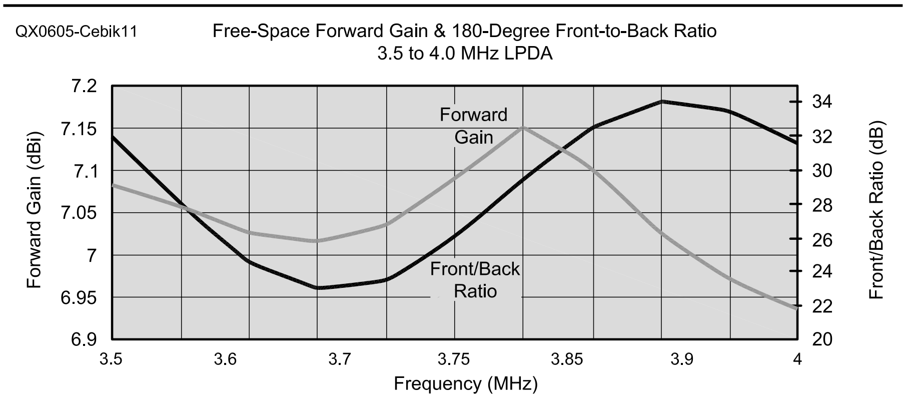


Figure 11 — Modeled free-space performance (gain and 180° front-to-back ratio) for an eight-element wire LPDA to cover 3.5 to 4 MHz.

Table 7
Modeled Performance Data for an Eight-Element 80 and 75-Meter LPDA with Doubled Wire Elements

Parameter	3.5 MHz	3.75 MHZ	4.0 MHz
Gain dBi	7.14	7.02	7.13
180° Front-Back dB	29.09	29.42	21.73
50-W SWR	1.23	1.64	1.39

Example 6: An Eight-Element Wire LPDA for 3.5 to 4.0 MHz

Yagi designs are capable of considerably wider bandwidth performance than we generally give them credit for — if we are willing to pay the price in terms of using more elements for a given boom length or using special driver sections. I have stretched one design to cover a 26% bandwidth with an acceptable SWR curve and modest, though usable performance. When we turn to the 80-meter amateur band, the 13% bandwidth combined with a need to use relatively thin wire presents a daunting challenge. One solution is to use a full LPDA for the band. With eight special wire elements and an 86 foot virtual boom length, such an array can provide about 7 dBi free-space gain with a front-to-back ratio that remains above 20 dB across the band. Since such an array is a major undertaking, I have not hesitated to use virtual 2-inch elements composed of dual wires (shorted at the phase line and the outer end) spaced anywhere from 8 to 12 inches apart along their length.

Figure 10 shows the modeled 50-Ω SWR curve, and sample data appears in Table 7. The gain and front-to-back curves are in Figure 11. All of the curves show a particular trait inherent to LPDAs. The curves undulate across the passband, but the peaks and nulls do not coincide with each other. Many newcomers to phased element design (including LPDAs) believe that the phase line provides a direct source of energy to the elements. While this belief is true, it is equally true that the elements exhibit mutual coupling. Hence, an LPDA (and any phased-fed collection of elements) is a form of parasitic array, at least in part. The balance among the energy sources for the elements undergoes continuous change as we change frequency. As one consequence, the operating parameters change with frequency. A good LPDA design is one that minimizes the level of change, although some change is inevitable in even the most ideal designs.

Conclusion

Our sample antennas have had two goals. One purpose was to show that it is possible by a variety of techniques to enlarge the operating bandwidth of a directional array. We have only touched a few of the many techniques available. The second aim was to expand our appreciation of the concept of operating bandwidth so that SWR becomes only one of many equal parts in the equation. We have many options in deciding which aspect of performance deserves primary attention, and equally many options in the techniques by which we achieve acceptable performance over a wide operating bandwidth. □□

Letters to the Editor

A Stable, Low-Noise Crystal Oscillator for Microwave and Millimeter-Wave Transverters (Nov/Dec 1999)

Doug,

John Stephensen, KD6OZH, reported difficulty in using a loop bandwidth wider than 1 Hz with a VCXO in the Nov/Dec 1999 QEX. Maybe the problem was in interfacing the low-pass filter to the phase detector. A modern fractional-N synthesizer such as an ADF4153 running with a 10- or 20-MHz reference should be compatible with a 100-Hz loop bandwidth and should give 35 dB less phase noise at 1 Hz.
— *William Cross, KAØJAD*, **kaØjad@yahoo.com**

Help for Oscillator Failure in the 8640B (Sep/Oct 2005)

Hi Doug,

Our company subscribes to *QEX* and I read of the problem with an HP8640B signal generator. I have two of those generators, bought off *eBay* in non-working condition and I repaired them both fairly easily.

Although one had a problem which seemed to suggest that the oscillator transistor had failed, I could not locate another transistor for replacement, so decided to dig a little further into the 8640. It should be noted that the transistor in the 8640 oscillator is very easily removed as it's a plug-in part in the side of the oscillator cavity — it's easy to blame this part. While in some cases it may well be the transistor, it's worth checking further. This involves removing the entire cavity from the HP8640B, and that's not for those mechanically challenged!

Disassembling the end of the oscillator cavity, you will find some additional components for biasing and other things. The problem turned out to be that one of the two resistors in that area had gone open-circuit. I can't remember which one it was now; but once the mechanical disassembly is completed, the parts are obvious. A quick check with an ohmmeter will determine if one of the parts is faulty.

Thanks for an excellent magazine!
— *73, Glenn Percy, VK3PE,* **glennp@bigpond.net.au**

Source Coding and Digital Voice for PSK31 (Nov/Dec 2005)

Doug,

First, I want to say how much I look forward to each issue of *QEX*. I ran across something interesting today via **slashdot.org**: a link to a link to an article describing the relationship between computer data compression algorithms and artificial intelligence. Frankly, the research is over my head but the idea makes a lot of sense. That fired off thoughts of your recent article suggesting use of computer voice recognition to produce a digital channel for voice audio. For a while now I've pondered the tremendous efficiencies of any system that allows a lot of assumptions, turning the actual real-time communication into just a footnote reference to some larger body of information (e.g., "implement standing order #44").

I'm starting to wonder if we are tackling this issue on the wrong end, putting the compression and data-channel efficiency work on the wrong end of the system.

Communication is all about the transmission of ideas, concepts and information between two systems. Much of the time the two systems that sit at the communication endpoints are persons. People are usually pretty intelligent, at least when compared to computers or data modems or radios. People also seem to be wired to find the patterns (communication) and residue of other people all around them. A recent article in the *Atlantic Monthly* presented this in the context of the current controversy concerning the theories of intelligent design. Even infants pick up on communications from other people with tremendous alacrity. A woman can communicate with her husband across a crowded room by lifting an eyebrow. What more data-compression should we be looking for?

The mechanical aspects of establishing a communication channel seem to be in the realm of machinery like computers. Once the channel is established, though, the best tool for data compression and decompression sits in the operator's chair.

So, that leads to questions about *Winlink* and other such systems. If the channel produced with such difficulty is used to send waving-smiley-face gif images via e-mail to ten of your dearest friends, is that really an efficient communication? On the other hand maybe the reception of such an image encodes a message meaning, "It's a boy" or "The new oil well is a gusher." It is up to the operator to know what words or pictures would be the most efficient means of getting the message across.

Handing a communication link over to an uninformed user is a recipe for inefficiency. As with Pony Express, telegraph and old-style long-distance phone calls, we need some way to inform the people at each end of the channel of the costs involved, so that they can adjust the level of intelligence and "compression" in their operations.
— *Chris Howard, WØEP, Estes Park, Colorado,* **chris@yipyap.com**

Hello,

I just read your article — very interesting. I actually had a similar idea a few years ago for a dictation program and format for recording. I think maybe a combination of my idea and your idea would be very useful. I am an opera singer and I use a special phonetic alphabet when learning foreign languages, or even writing down the exact sound I want to make in English.

The biggest problem in your model is the speech-to-text conversion, which is error-prone and computationally intense. Text-to-speech programs, on the other hand, are rather simple because they do not need to understand the sentence to convert the text to sounds.

My approach to this problem is to convert the recorded sounds into the IPA (International Phonetic Alphabet) equivalent, including intonation and stress symbols. This way, the computer just has to keep track of how you said something and condense it down to the IPA symbols. These can then be played back on an IPA-to-speech player, and should be more accurate than any speech-to-text and vice-versa program.

Take a look at this link for a very elaborate definition on IPA: **http://en.wikipedia.org/wiki/International_Phonetic_Alphabet.** This site gets way more complicated than what would actually be required. These symbols may look unusual but they make sense to linguists and opera singers. I think you will find that you could condense the most used sounds down to just a few bits and the less frequently used sounds to larger bits. I think there are only maybe 30 sounds in English that are used, and there are around 120 total sounds possible, not including stress and intonation.

Another advantage of this system would be its language-independence. My guess is that at PSK31 speeds, you could send just as many IPA symbols as letters and you would find that most words have more letters than actual sounds, though that isn't always the case. I just think it would be a more accurate way to convert the speech into a very compressed digital means. In my research, however, I also ran across a few universities doing this same research for cell-phone companies. I believe one or two of those groups had a patent pending. I don't know how that would effect an idea like this for ham radio but I do know that I wasn't the first to come up with the idea. That was back in 1999 or 2000. Anyway, for what it's worth, I could do some more research into this if you actually think hams would like this — I know cell-phone companies do!

I have found a few programs that will convert speech to and from IPA and I am sure there must be some source code available to look at if you are interested in pursuing this further. My guess would be that it would be better to create a separate PSK or FSK format using a similar varicode method, but not have it be PSK31, but its own unique mode.

Well, you certainly got my brain thinking about this idea again. I will have to look into this more. I never thought about applying it to Amateur Radio until I read your article.
— Brian Shircliffe, AD5OS, 2226 West 18th #46, Houston, TX 77008, **bshircliffe@houston.rr.com**

Quantifying SETI (Jan/Feb 2006)

Hi Doug,

I've been away in Australia since mid-January so it is only during the last few days that I have seen the Jan/Feb issue of *QEX*. I was shocked and saddened when I read that Bob Schetgen, KU7G, had become a silent key. I had no idea that he had been ill. I had reason to phone him on several occasions and he seemed to be a really nice guy.

On a more cheerful note I was particularly pleased to see the article on SETI. It addressed all of the questions that I had in mind when I e-mailed you on this subject last year. Having said that, however, I am still of the opinion that they will never hear anything but I hope that I'm proved wrong. My gut feeling is that this wonderful planet of ours is not unique but that the Almighty made sure that intelligent beings were sufficiently separated in distance and time so as not to be able to compare notes. I think N6TX should be congratulated on writing a superb article.
— 73, Ron Barker, G4JNH, **g4jnh@onetel.com**

Ron,

You might be right about the paucity of advanced civilizations in the galaxy. Since we're trying so hard to listen for them, let's start discussing what we'd do were we to establish communications. Is our current leadership up to the task?
— Doug Smith, KF6DX, QEX Editor, **kf6dx@arrl.org**

Doug,

There is a problem with the Drake Equation's prediction of the number of intelligent, communicative civilizations that exist in our galaxy.

The final factor in the famous Drake Equation is a measure of the length of time a civilization is communicative. The assumption is that a civilization will develop the technologies to communicate by radio, and that they would continue using radio communication indefinitely after that.

At the time that Frank Drake penned the equation that bears his name, the primary concern was that some cultures would self-destruct through war, particularly nuclear war, which would take the civilization to a more primitive, non-communicative state — or to extinction. Other cultures would get past the wars and continue to develop over very long periods of time and continue radiating in the electromagnetic spectrum.

The final factor does not take into account the direction that technology may lead a civilization. I predict that the vast majority of radio emissions from a civilization will cease fairly quickly, not because the civilization has extinguished itself or bombed itself back into the Dark Ages, but because technological advances will continue to expand rapidly and exhaust the available radio spectrum. At some point, a civilization's technology will run out of bandwidth and move to communications channels outside the electromagnetic spectrum. This is not a consideration in the Drake Equation.

We can look at ourselves and see how technology is moving us in the direction of an electromagnetically quiet civilization. We still have 100+kilowatt television and radio stations for the broadcast media, but we are rapidly moving toward cable, broadband Internet and satellite services, which either do not broadcast at all or focus all of their radiation directly and completely at the earth. It will not be long before all high-powered transmissions will cease in favor of contained (nonradiating) media. About the only high power transmitters we may have left by the end of the 21st century will be radar (and DX hounds — HI!)

Some estimates I have seen for the final factor of the Drake equation assume the life expectancy of a civilization to be in the tens of thousands of years. If our own civilization is any measure, the final factor in the Drake Equation should probably assume a civilization to broadcast its presence for only 100 to 300 years. Remember, our first commercial radio broadcasts started early in the 20th century. We may be quiet by the end of the 21st — a total span of less than 200 years.

Regardless of how small the final factor of the Drake Equation becomes, it still leaves a lot of civilizations out there to hear; however, the chances of hearing a civilization that is significantly ahead of us technologically may be grossly overestimated.
— Bill Brown, KG5AR, Little Rock, Arkansas, **kg5ar@sbcglobal.net**

A 100-W Class-D Power Amplifier for LF and MF (Mar/Apr 2006)

Greetings:

Unfortunately, there is an erratum to report. Q201 is upside-down in Figure 3, on page 11. The source of Q201 is connected to the V_{DD} supply while the drain is connected to the drain of Q202.
— With best regards, Frederick H. (Fritz) Raab, Ph.D., Green Mountain Radio Research Company, 77 Vermont Avenue, Fort Ethan Allen, Colchester, VT 05446, **f.raab@ieee.org**

Tech Notes: Quantifying Measurement Uncertainty (Jan/Feb 2006)

Doug,

That is a good article on measurement uncertainty. A useful but somewhat heavy reference, meant especially for our sort of measurements, may be found in the European Telecommunication Standards Institute Technical Report TR 100 028-1 and TR100 028-2. Two parts: Part 1 is 241 pages, Part 2 is 285 pages. They are available free from the ETSI web site at **www.etsi.org**.
— 73, Peter Chadwick, G3RZP, **peter.chadwick@Zarlink.Com**

An L-Q Meter (Mar/Apr 2006)

Doug,

I am thoroughly enjoying the Mar/Apr 2006 issue of *QEX*. I will get to every article, but I am picking and choosing and not going from front to back.

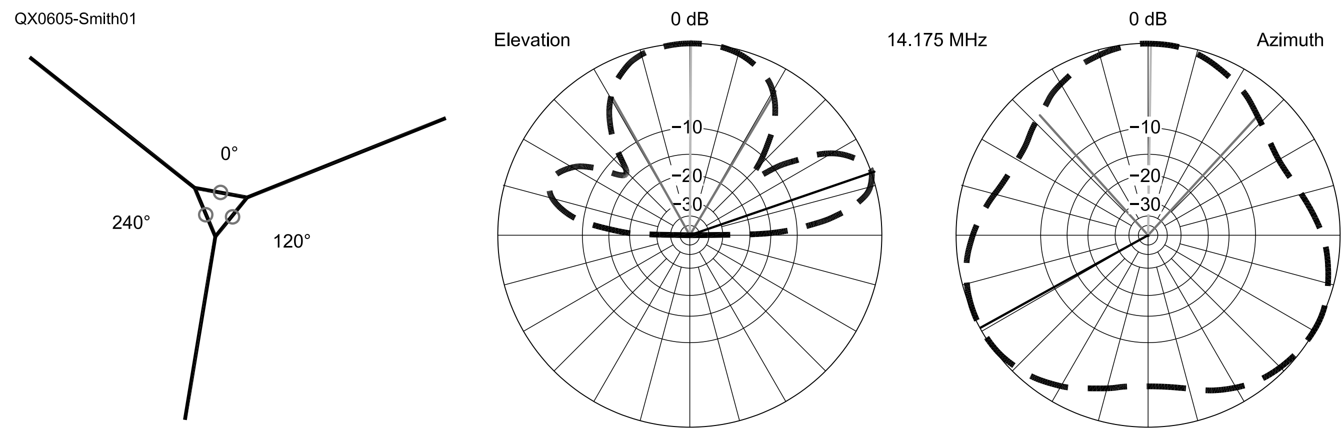
In "An L-Q Meter," on p 31, middle column, third paragraph, first sentence, appears "The LCD display…LCD display." The D in LCD stands for "display." Therefore this is saying "liquid crystal display display." I guess people want it known that this is a display. So, I think, this should be just "LCD" or "LC display."

On p 31, right column, under "Other Considerations," first sentence: "There is a potential problem when using the instrument with very high-Q inductors, and Appendix A discusses it in detail." My reading of Appendix A does not reveal this detail. Perhaps in editing Appendix A this detail was dropped. There is mention of the high-Q problem in "Conclusion."

Also in Appendix A, on p 28, reference is made to a "Figure A2," See "2. Measurement Mode," fourth and fifth lines. There is no Figure A2 shown
— 73, Larry Joy, **ljoy@kantronics.com**

Larry,

Thanks for those corrections. Because the machine measures the ratio v_m/v_c, when that ratio approaches its limit it becomes difficult



Feed system for the phase-fed 20-Meter Y

Plots of a 20-Meter Y with simultaneous feeding of all pairs at phase angles of 0°, 120° and 240°

Figure 1 — Feed system and modeled radiation pattern for three-phase-fed Y antenna.

to assess Q. I think that's the detail the author was referring to. The reference to Figure A2 should be to Figure B instead.
— Doug Smith, KF6DX, QEX *Editor,* **kf6dx@arrl.org**.

Antenna Options (Mar/Apr 2006)

Doug,

One can make W4RNL's Y antenna omnidirectional by driving the 1, 2, 3 feed points with three-phase power (where the 1, 2, 3 feed points are driven with RF power at 0°, 120° and 240° phases, respectively).
— Peter Traneus Anderson, KC1HR, **traneus@comcast.net**

Doug,

Although the idea is intriguing, I do not know if it is practical. I took the 20-meter Y model at 600 inches (50 feet) and modified it to have three short feed wires, with simultaneous 0°, 120° and 240° phased sources. The modeling plan and the elevation/azimuth plots are Figure 1.

The original switched Y had bidirectional gain of a little over 7 dBi at 19° elevation. The phase-fed Y has a maximum gain of a little over 6 dBi — straight upward. There is a moderately strong triangular pattern at 19° elevation, with a gain only a little under the upward max value.

The impedance of each source point is about 85 Ω. One would need three identical twisted pairs of about that impedance as a Z_0, and sufficiently separated so as not to interact. Back at the shack end of the line one could set up delay lines for 120° and 240°; but that is a lot of line, and it is good only for one band. One might also install delay lines at the antenna hub with a single main feed line back to the shack.

Such lines might be difficult to route to prevent undesired interactions, however. In the shack, double-shielded coax might do the trick, with a 1:1 balun for the twisted-pair runs. Coiling delay-line coax in old large metal popcorn cans would likely serve the added shielding needs, although the usual decorations on such cans might need disguising to maintain proper station dignity.

All is not lost, though. If we scale the design for 40 meters and adjust for practical height considerations, then we might have a workable combination of an NVIS and longer-distance antenna. MARS and other military-affiliate and related operations often look for antennas that would serve both short-range and long-range regional needs, and such an antenna pattern might meet their goals. Because of the need for delay lines, the antenna would be frequency- or fairly narrow-band-specific. I have not looked into what lower heights as measured in wavelengths will do to the pattern, but this may be a start in the design process if one has a need for such an antenna.

It is likely that the junction of the three lines (one direct and two delay) at the TX end of the line would need a transmission-line transformer at about a 2:1 ratio to raise the parallel connection composite impedance of a little over 25 Ω up to 50 Ω.
— *73, L.B. Cebik, W4RNL,* **cebik@cebik.com**

In the next issue of QEX

George Steber, WB9LVI, shows how to build a useful tool for your lab: a PC-based curve tracer. Curve tracing is a simple way to test and characterize solid-state devices. George explains how a sound card gets the job done even when its inputs and outputs are ac-coupled. His circuit is inexpensive — it's a straightforward construction project. The software will be available on the ARRL Web site. Check it out!