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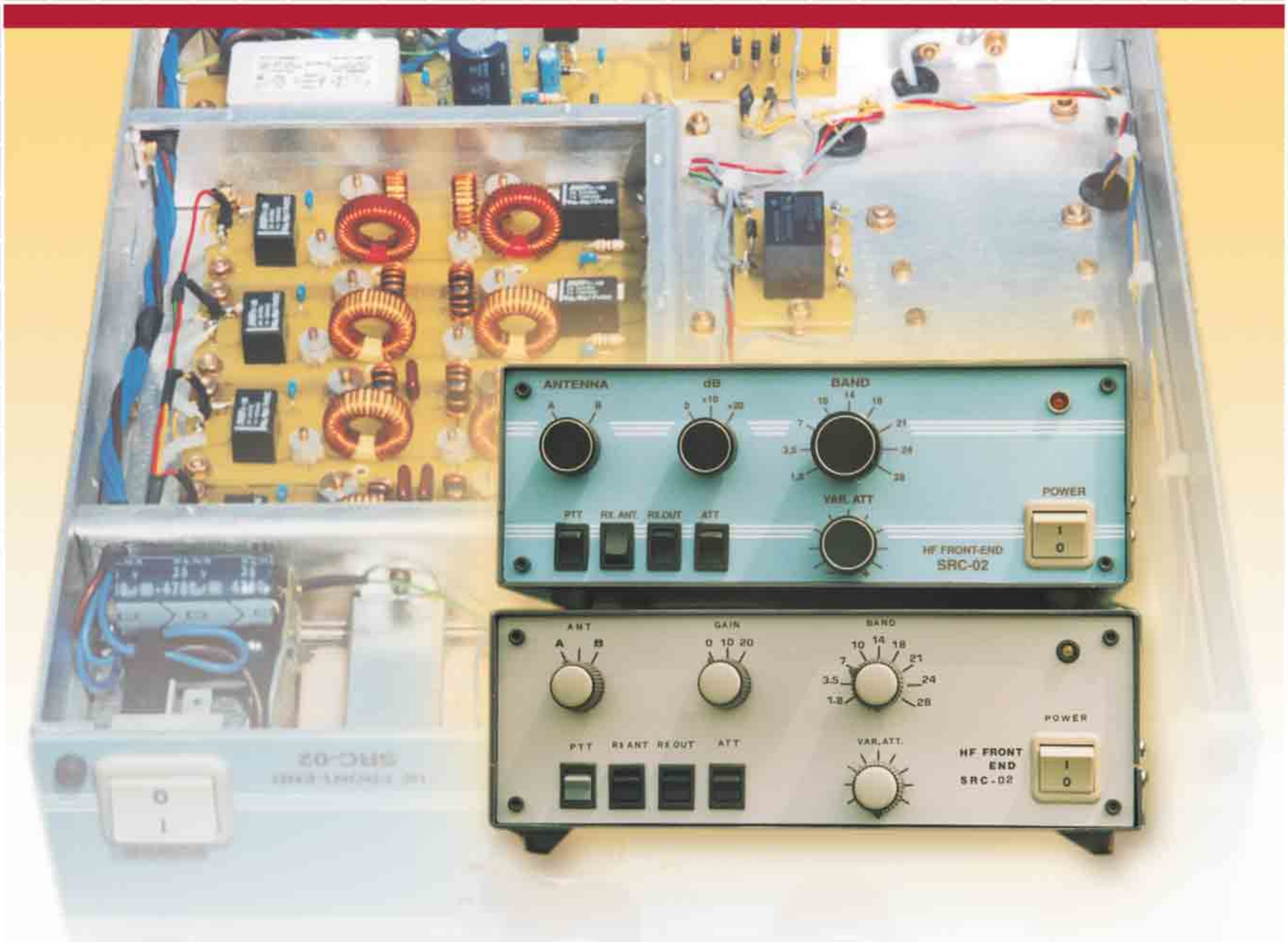
QEX

INCLUDING: COMMUNICATIONS QUARTERLY

Forum for Communications Experimenters

March/April 2003
Issue No. 217



IK4AUY/I4FAF Preselector/Preamplifier/Variable Attenuator—an aid for beleaguered HF receivers.

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INCLUDING: COMMUNICATIONS QUARTERLY

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About the Cover

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Mar/Apr 2003 *QEX* Advertising Index

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1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,
2) document advanced technical work in the Amateur Radio field, and
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Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted on IBM or Mac format 3.5-inch diskette in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX*. Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Empirical Outlook

Technical Standards in Amateur Radio

Every inspired designer eventually thinks he has the best way of doing something. Although it has been said that great minds think alike, history has shown ample evidence of a tendency to diverge, especially when we operate from so-called ivory towers. The need for technical standards to limit that divergence becomes plain as we organize ourselves.

Though as one wag observed, "The nice thing about standards is that there are so many of them to choose from"—*Andrew S. Tanenbaum*. You can declare some particular arrangement of things a standard, but it really becomes that only after it achieves popularity through actual use. Subsequently, someone always comes along with a better way. With the rapid pace of technological development today, it is harder than ever to create rigid definitions of things and stick with them for long.

In a sense, standardization is anathema to a spirit of innovation and experimentation; but it is necessary where interoperability of equipment is required. AM was the standard phone mode for many decades in Amateur Radio and elsewhere. SSB bumped it out of first place in our service only after commercial interests used it for quite some time. Hams were involved in the early development of SSB; but its popularity did not make it a standard until much later. AM is still fairly popular because its baseband response is easily extended to dc and because it is so easily demodulated.

Now an impatient few are evidently surprised that the next big transition from SSB to digital phone has not already occurred. They say that all we need is a standard, and soon; but we have to ask in reply, "What is the data communications standard in Amateur Radio?" Well, there is not one standard, but many. In fact, the growing variety of digital modes represents a very healthy part of our avocation. The lack of a single standard has not hindered those who enjoy Baudot RTTY, Pactor, PSK31 and the rest.

What is the slow-scan television standard? Again, there are several. What is the standard for control of rigs using external computers? What is the standard microphone pin assignment in Amateur Radio? Power connector? To each his own.

What we are discovering is that it is very difficult—if not impossible—to establish standards *after* designers have done their separate things. The time to get organized is *before* the designs mature. Even so, it takes vision, a certain measure of cooperation nurtured among interested parties *and time*. In addition, the goals of those who want solely to better our service are often at cross-purposes with those who want to obtain technical advantages over their competition.

It is easy to sit back and take pot shots at those who are wrestling with these issues, but why fire in the dark? Reach out to your fellow hams directly and discuss with them where we are going and why. Write about it—for *QEX*!

In This Issue

We have two segments of series on software radio in this issue. Jim Scarlett, KD7O, presents the second installation on his digital receiver design. Jim was kind enough to arrange for a donation of state-of-the-art hardware to the ARRL Software Radio Working Group. Gerald Youngblood, AC5OG, concludes his series on the SDR-1000. Gerald is thinking about releasing his design as a kit.

Dan Handelsman, N2DT, and I put our heads together to produce an introduction to intellectual-property law for you. Many thanks to Jack Stone of *antenneX* for permission to use some material from Dan's earlier work.

Dan Maguire, AC6LA, describes how spreadsheets and antenna-modeling software can communicate. This allows far more design exploration than simple frequency sweeps.

James Lawrence, Sr, WB5HVH, brings us a spreadsheet solution for RF path analysis. It is a simple and efficient way to determine your chances of success on VHF and above over any given terrain.

Sergio Cartoceti, IK4AUY, writes about his experience in difficult HF DXing situations and what he did to improve his lot. One focus is on second-order distortion in receivers. In *RF*, Contributing Editor Zack Lau, W1VT, describes a portable two-element 6-meter Yagi—*Doug Smith, KF6DX, **kf6dx@arrl.org***.

A High-Performance Digital-Transceiver Design, Part 2

Part 1 of this series looked at some of the system-level design and tradeoffs for a high-performance transceiver that translates directly from RF to digital. In this installment, we'll look at the details of the receiver front end.

By Jim Scarlett, KD7O

The first installment of this series gave an overview of a high-performance transceiver using an "almost all digital" approach.[1] While still using analog filters and amplifiers, this approach translates directly between RF and digital domains. Design goals were presented that would meet or exceed the performance of commercially manufactured radios.

Here in Part 2, we'll look at the design details for the receive side of the radio. DSP will not be covered until a later installment, but we'll look at the receive signal processor (RSP) hardware design here. Considerations for using the ADC will be covered, including the use of dither and interfacing the ADC to the RSP. The clock source will get special attention as well.

[1]Notes appear on page 9.

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Analog Processing First

First I'll describe the analog front end. The RF amplifier uses ideas that have been covered before. I'll describe how those ideas come together in this design. I won't spend much time on the filters, since the basic design has already been covered in detail in *QEX*.[2]

I mentioned in Part 1 that my filter capacitor values are different from Bill Sabin's to account for my own areas of interest. I also made some other changes regarding components. In the interest of minimizing size and cost, I use NP0 ceramic-chip capacitors in the filters. These capacitors will be okay for up to a few watts of input power. See Fig 1 for the filter schematics. Capacitor values of "0" indicate where the board has pads for additional capacitors to adjust the values while tuning the filters.

The relays are inexpensive and can easily handle the required power. They also switch in about 5 ms, so they are compatible with rapid switching between transmit and receive—or between bands. Additional relays are included in the filter/amplifier module for transmit/receive switching of the module.

RF Amplifier Board

As I mentioned in Part 1, I rejected the use of a common-base amplifier configuration because of poor reverse isolation. While looking at the amplifier in John Stephensen's noise blanker,[3] I decided that the topology would meet my needs in this area. The desired *IP3* performance would be achieved through a push-pull transistor pair operating at a higher bias current.

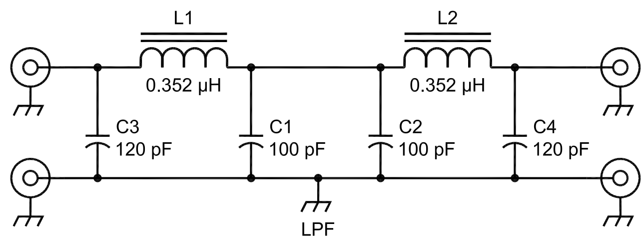
The resulting amplifier is shown in Fig 2. A Mini Circuits 1:1 transformer generates the balanced signal used by the two transistors. I used some Motorola MRF5811 transistors that I had on hand. These devices—basically an MRF581 in a SOT-143 package—provide an excellent noise figure while running at high currents. In this application, they are being operated with a bias current of about 40 mA.

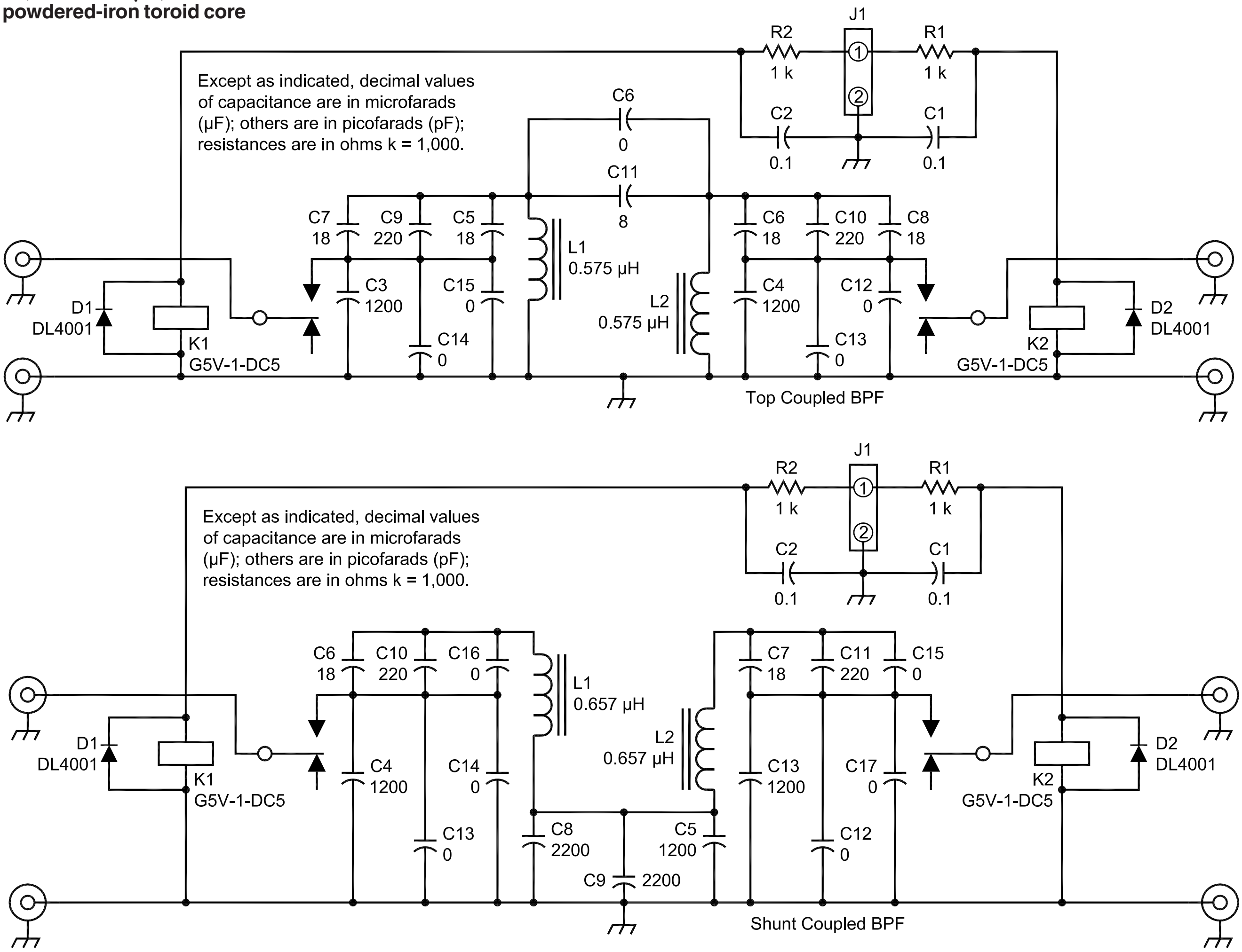
The 20-Ω resistors at the emitters set the bias currents. Along with the feedback transformers, they determine the input impedance of the amplifier. The transformers are wound on binocular ferrite cores. The current levels are too high for typically available RF transformers, which should not be substituted. The transformer impedance ratio also helps set the gain of the amplifier.

The transistors can be powered with +9.5 to +12 V. The higher supply voltage will improve the *IP3* by a couple of decibels, but will force the transistors to dissipate more power. I am operating the prototype at about +10 V. Later versions will probably have a regulator on board to generate this voltage.

The 475-Ω resistors provide a proper output match for the filters. Since this resistor is unnecessary in the biasing scheme, it is ac coupled. Had it been dc coupled, a ½-W resis-

Fig 1—Filter schematics: (A) low-pass; (B) top-coupled band-pass; (C) shunt-coupled band-pass. Values shown are for the 20-meter filters. Component values for the other HF bands are available on the ARRL Web site. Resistors and capacitors are 0805 SMT unless noted. 0 pF capacitors are pads only.
LPF
L1, L2—0.352 µH, 8 turns #26 enameled wire on a T50-6 powdered-iron toroid core
Top-coupled BPF
D1, D2—DL4001
K1, K2—PC mount SPDT relay (Digi-Key #G5V-1DC5)
L1, L2—0.575 µH, 10 turns #26 enameled wire on a T50-6 powdered-iron toroid core
Shunt-coupled BPF
D1, D2—DL4001
K1, K2—PC mount SPDT relay (Digi-Key #G5V-1DC5)
L1, L2—0.657 µH, 12 turns #26 enameled wire on a T50-6 powdered-iron toroid core





tor (expensive in chip form) would have been required. As it is, a 100-mW 0805 resistor is fine (and cheap). The output transformer matches the amplifier to 50 Ω. The impedance ratio also helps determine the amplifier gain. The output transformer is a conventional broadband transformer with a 16:1 impedance ratio (center tapped), wound on a BN3312-43 core.

The gain of the amplifier was measured to be flat within less than 2 dB across the HF spectrum, with the least gain at the high end. The input and output match also showed some degradation at the high end. Still, the SWR is 2:1 or better across the HF spectrum. One might get better performance using transmission-line transformers, but I did not try this. As is, the total gain of the front end at 20 meters was measured to be just over 6.8 dB, which is close enough to the 7 dB used in the spreadsheet in Part 1.

As shown in the block diagram (Fig 8 of Note 1), there are two RF amplifiers. One is used for the lower HF bands and both are used on the higher bands. The preamplifier used on the higher bands can be switched in or out using the relays on the board. I opted to leave the second RF amplifier in line at all times and therefore did not populate the relays or control circuitry. Jumpers were installed across the relay pads.

I have not yet measured the *IP3* of this amplifier, but my simulation in *Serenade SV*[4] gave better than +44 dBm. Because the reverse isolation is very high, this value should not suffer much even when terminated by the band-pass filters.

RF to Digital

At the heart of this receiver architecture is the AD6645 analog-to-digital converter. The design for the ADC board is quite straightforward, as shown in Fig 3. Notice that all inputs to the ADC are differential. This is necessary for all devices of this type if you are to get maximum performance. Even with much less-precise parts, the noise and distortion performance above a few megahertz will degrade if you use a single-ended input.

Local regulation is provided to reduce noise-pickup issues. Decoupling

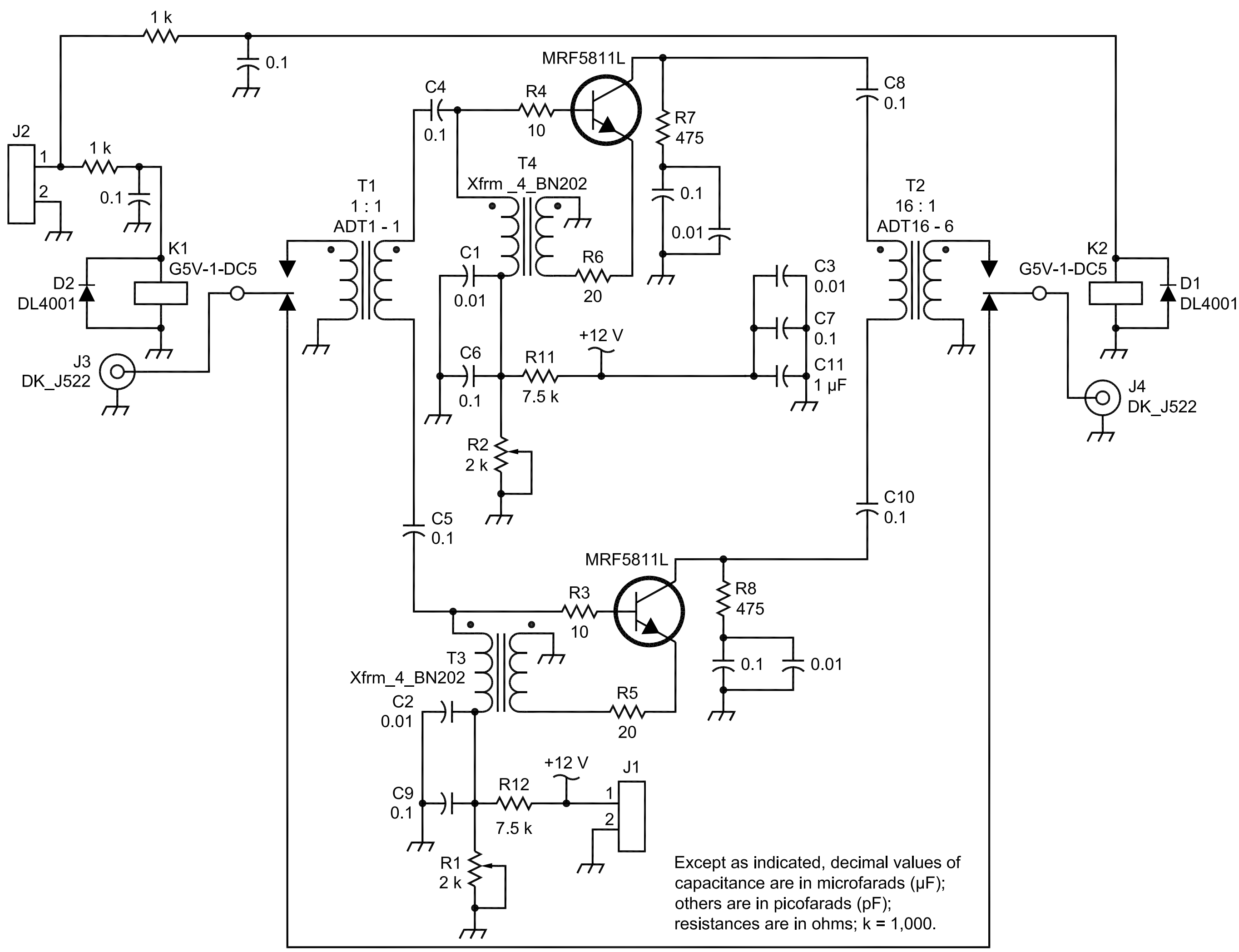


Fig 2—RF amplifier schematic diagram. Resistors and capacitors are 0805 SMT unless noted.

C11—1 µF, 25-V, X7R, 1206 SMT capacitor
D1, D2—DL4001
J1, J2—2-pin header
J3, J4—PC mount SMB bulkhead jack (Digi-Key #J522)
K1, K2—PC mount SPDT relay (Digi-Key #G5V-1DC5)
Q1, Q2—MRF5811 NPN transistor
R1, R2—2 kΩ, 10-turn trim pot (Digi-Key #3214W-202ECT)
R9, R10—56.2 Ω, 1210 SMT resistor
T1, T2—6 t primary, 3 t secondary on a BN202-61 binocular ferrite core
T3—1:1 transformer (Minicircuits ADT1-1)
T4—8 t center-tapped primary, 2 t secondary on a BN3312-43 binocular ferrite core

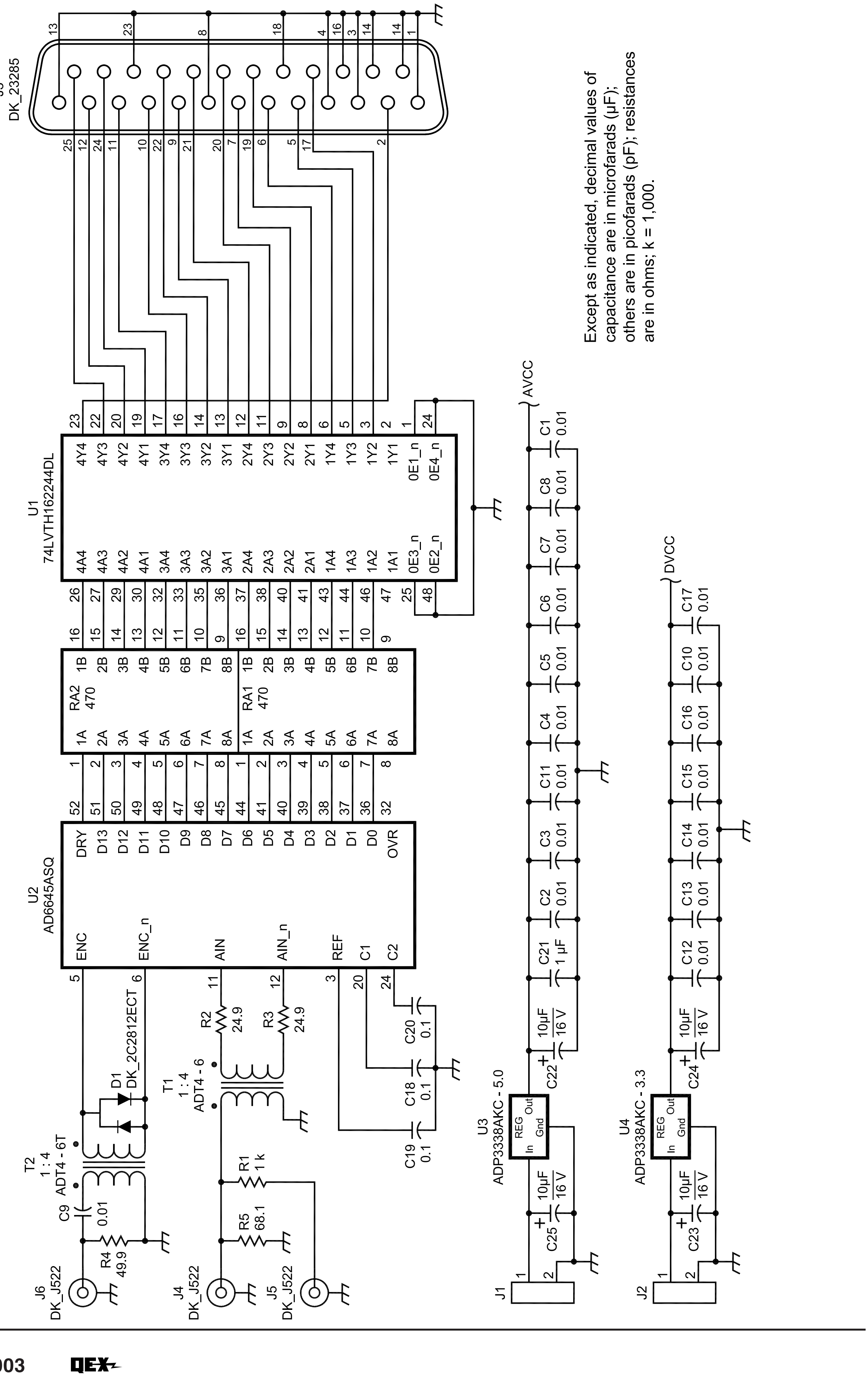

J3
DK_23285
U1
74LVTH162244DL
U2
AD6645ASQ
RA2
470
RA1
470
T1
1 : 4
ADT4 - 6
T2
1 : 4
ADT4 - 6T
D1
DK_2C2812ECT
R1
1 k
R2
24.9
R3
24.9
R4
49.9
R5
68.1
C9
0.01
C18
0.1
C19
0.1
C20
0.1
J4
DK_J522
J5
DK_J522
J6
DK_J522
J1
J2
U3
ADP3338AKC - 5.0
U4
ADP3338AKC - 3.3
AVCC
DVCC
C22
C23
C24
C25
10µF
16 V
C21
1 µF
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.

Fig 3—ADC schematic diagram. Resistors are 0805 SMT unless noted. Capacitors are 0603 SMT unless noted.
C9—0.01 µF X7R 0805 SMT capacitor
C21—1 µF, 16-V X7R 1206 SMT capacitor
C22-25—10 µF, 16-V tantalum capacitor
D1—Dual Schottky diode, series configuration, SOT23 (ZC2812E, BAS70-04)
J1, J2—2-pin header
J3—DB25 right-angle, PC mount (Digi-Key #A23285)
J4-6—PC mount SMB bulkhead jack (Digi-Key #J522)
RA1, RA2—Resistor array, 8×470 Ω (Digi-Key #742C163471)
T1—1:4 transformer (Mini Circuits ADT4-6)
T2—1:4 transformer, CT (Mini Circuits ADT4-6T)
U1—16-bit buffer with integrated termination resistors, SSOP, 74LVTH162244DL
U2—High-speed ADC, AD6645ASQ-80
U3—3.3-V low-dropout voltage regulator, Analog Devices ADP3338AKC-3.3
U4—5.0-V low-dropout voltage regulator, Analog Devices ADP3338AKC-5.0

capacitors are placed as closely as possible to each power pin. Small (0603) capacitors are used to allow placement on the same side of the board as the ADC. At this resolution and speed, even using vias to decoupling capacitors on the bottom of the board can degrade converter performance, according to the applications engineers at ADI.

The clock input is terminated to provide a 50-Ω match to the synthesizer. A 1:4 transformer converts the input to a differential signal, which is then clipped by the diode pair and applied to the ADC. As described in the AD6645 datasheet, the diode clipping prevents large voltage swings that can feed through to other parts of the device. It also helps reduce noise susceptibility.

Similarly, the analog inputs to the ADC are provided differentially via a transformer. The 24.9-Ω series resistors isolate the transformer from the ADC inputs. Notice that the 1-kΩ differential input impedance is not fully matched. Instead, the input signal is transformed up to 200 Ω. The result is that the full-scale input is about +4.8 dBm. Remember that the ADC is a voltage device—we don't need to worry about mismatch power losses. A termination resistor provides the proper match to the 50-Ω source.

The dither signal is also summed at the input node, via a 1-kΩ resistor. This impedance minimizes loading of the input and divides the signal down to the appropriate level. The datasheet recommends a dither power level of about –19 dBm. I thought about a simpler-looking idea where the dither would be injected at a center tap on the transformer secondary. However, as Brad Brannon, N4RGI, pointed out, this results in a balanced noise signal that negates the effectiveness of dither.

Output signals are series terminated to minimize dynamic currents at the output pins, which can reflect back and disrupt part performance. The termination resistors also help improve the quality of the signals at the buffer input. These termination resistors should be as close to the output pins as possible, which is why I used small surface-mount resistor arrays.

The output buffer was chosen for its timing characteristics: in particular, the skew between the bits (0.5 ns). The timing window between the clock and data signals at the RSP is very tight. Excessive variation between the RSP clock and the incoming data can result in bad data due to timing violations. The DRY signal from the ADC meets the timing requirements for the RSP clock, as long as additional variation is not introduced. By sending this signal through the same buffer as the data lines (since they only take up 14 of 16 bits), the variation is minimized and proper timing is guaranteed.

An important aspect of ADC performance is the quality of the encode clock and the aperture jitter of the converter itself. If the level of jitter on the clock or in the sample-and-hold circuit of the ADC is too high, it can directly affect the noise performance of the converter. This topic was discussed briefly in Part 1; but after further reflection, I felt that the explanation was somewhat confusing. Some of the feedback I received confirmed this. Therefore, a discussion is included in the sidebar "Jitter and ADC Performance."

A Low-Noise Clock

The ADC used in this architecture is subject to the same problems with a noisy local oscillator as a traditional analog design. In the traditional design, the problem is known as reciprocal mixing. Here, we have a degradation of the S/N caused by clock jitter. The result is the same, with noise sidebands appearing on the incoming signals. If the oscillator is noisy enough, these sidebands may mask a desired weak signal. This is discussed more thoroughly in the jitter sidebar.

One advantage of this architecture is that the tuning is done digitally within the RSP. Therefore, the "local oscillator" can actually be a low-noise crystal oscillator instead of the typical wide bandwidth VCO.

The oscillator design (Fig 4) is based closely on the design John Stephensen presented in *QEX*.[5] For the most part, I just altered components for the frequency difference, or for ease of procurement. The operating frequency is 64.96 MHz, which is phase-locked to a 12.992-MHz reference. The ADC frequency was chosen to allow integer decimation to both 40 kHz (for FM) and 16 kHz (for SSB and CW).

The key to John's design is the input level in the JFET. In this configuration, the impedance of the FET is fairly high (50-100 Ω). Therefore, since the same RF current flows in the resonator and FET, the oscillator input power is higher than the power dissipated in the resonator. For a given resonator power, most common oscillator designs extract much less. The higher power leads directly to lower phase noise, per Leeson's equation.

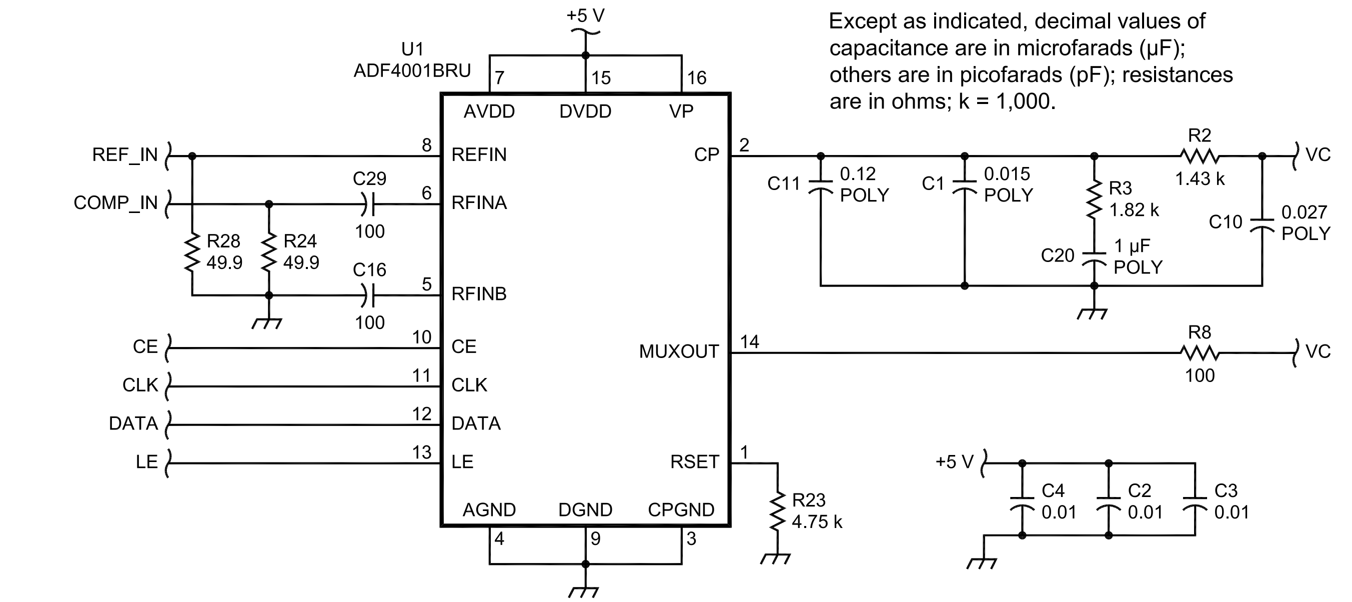
Approximately +8 dBm is available to the ADC after the 3-dB attenuator. A buffered output from the emitter follower is also attenuated to provide the correct level to the PLL chip. The PLL used is the ADF4001, which is designed specifically for clock applications below 200 MHz. There is no prescaler and there are no minimum division ratios for either the RF or reference frequency.

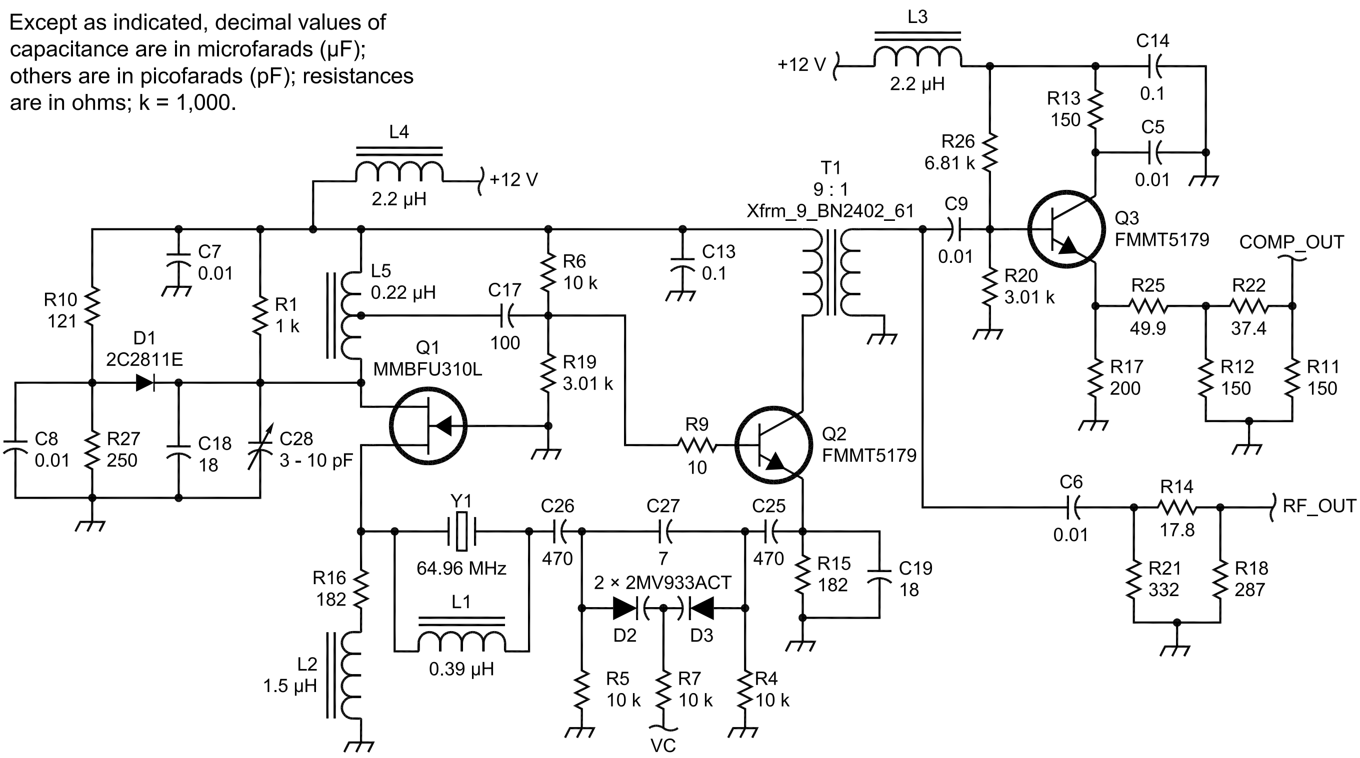
Other advantages of this device are the very low phase-noise floor and that the phase/frequency discriminator (PFD) can be used up to 100 MHz. In this application, we'll use it at 12.992 MHz with no reference division. The higher frequency in the PFD helps lower the noise floor within the loop bandwidth. For example, doubling the reference frequency increases the PFD noise by 3 dB, but the multiplier noise is reduced by 6 dB. Here, the PLL noise floor is about –128 dBc/Hz, and the VCXO noise will dominate outside the loop bandwidth of about 230 Hz. Close-in (inside about 100 Hz), the noise will be determined by the reference oscillator. The jitter sidebar provides additional details for the predicted clock performance.

Cheap and Easy Dither

There are two main ways to generate large-scale dither for the ADC. One that is commonly used in commercial applications—such as test equipment—involves digitally generating pseudorandom noise. This wide-band noise is then injected into the front end of the ADC, and subsequently subtracted from the digital result. The second alternative is to inject narrow-band noise that is outside the bands of interest. This method is used here.

The dither circuit is shown in Fig 5. It is based on a circuit presented in the manual for Analog Devices' *High Speed Design Techniques* seminar. I opted to use op amps throughout instead of the variable-gain amplifiers presented in that discussion.[6] I also used an op amp with a higher input impedance (the AD8055), which allows me to use a larger resistor for my noise





source. This requires less total gain and reduces the variation in the noise levels, as the resistor noise dominates. Additional variation in generated noise can be caused by part-to-part differences in the op amp's current noise. The third amplifier stage has adjustable gain to set the proper level.

The noise generator (first three stages) is followed by an active low-pass filter. This filter has a bandwidth of about 450 kHz; with six poles, it reduces the injected noise below the noise floor at 160 meters. Within that constraint, the bandwidth is enough to minimize the gain required to achieve the correct noise power, and makes sure the noise is random rather than sinusoidal. The output of the final filter stage drives a buffer that has a gain of two.

The AD8055 amplifier is a high-speed device, with a 3-dB bandwidth of over 300 MHz. This is helpful in two ways. First, with the high gains involved, the actual bandwidth of the amplifiers ends up being only a few megahertz. Second, the output impedance of an op amp increases with frequency. This allows the noise signal a forward path through the filter feedback capacitors, since the op amp output is no longer an ideal, zero-impedance voltage source. The result is that the ultimate rejection of the filters is limited. The AD8055 output impedance remains very low into the VHF range, thus minimizing this effect.

Receive Processing Hardware

The hardware implementation of the Receive Signal Processor (RSP) is fairly simple. See Fig 6. At this point, the sig-

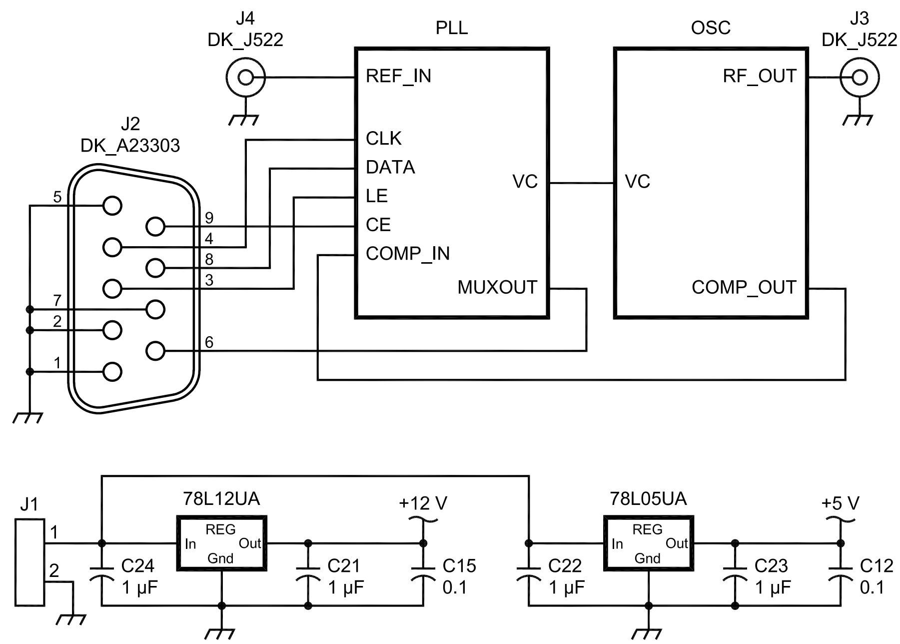


Fig 4—Receive PLL schematic diagram. Resistors and capacitors are 0805 SMT unless noted.

C1—0.015 µF polyester capacitor (Digi-Key #P10957)
C2-4—0.01 µF 0603 SMT capacitor
C10—0.027 µF polyester capacitor (Digi-Key #P10960)
C11—0.12 µF polyester capacitor (Digi-Key #P10967)
C20—1 µF polyester capacitor (Digi-Key #P10979)
C21, C24—1 µF, 25-V, X7R, 1206 SMT capacitor
C22, C23—1 µF, 16-V, X7R, 1206 SMT capacitor
C28—3-10 pF trimmer capacitor, SMT (Digi-Key #SG2002)
J1—2-pin header
J2—DB9 right angle, PC mount (Digi-Key #A23303)
J3, J4—PC mount SMB bulkhead jack (Digi-Key #J522)
L1—0.39 µH, 1210 SMT (Digi-Key #M6100)
L2—1.5 µH, 1210 SMT (Digi-Key #M6107)
L3, L4—2.2 µH
L5—12 turns #28 enameled wire on a T37-12 powdered-iron core, tap 3 turns from cold end
T1—primary 6 turns #28, secondary 2 turns #28 on a BN2402-61 binocular ferrite core
U1—PLL, ADF4001BRU
U2—5-V voltage regulator, SOT89, 78L05UA
U3—12-V voltage regulator, SOT89, 78L12UA
Y1—64.96 MHz, third-overtone (International Crystal)

nal has been digitized, so we only need be concerned with signal integrity, primarily on the input side. At the output, the serial interface only needs to be fast enough to get the I/Q words out at a maximum 40-kHz rate for DSP processing by another computer. Signal integrity is easy at this speed, but we still use proper buffering and a series termination to keep everything clean.

The RSP is initialized and controlled using the parallel *Microport* interface. The serial interface can be used for control after setup, but cannot be used for initialization. Since the parallel interface is necessary anyway, I thought it would be simpler to use it exclusively and to use the serial port only for the signal path. I also wanted to use the DSP for signal processing only, not for housekeeping. The Microport interface is set to mode 0 in hardware by grounding the **MODE** pin.

Grounding the **PAR/SER** pin selects the serial interface for output data. The serial word length is set to 24 bits, which transfers all of the data but minimizes overhead. The RSP is the serial master, so control signals are generated in the RSP and sent to the DSP serial port. The serial clock is set to its minimum value, which meets the bandwidth requirements for a 40-kHz output using 24-bit I/Q words.

The actual use of the RSP functions is more closely related to the DSP than to the analog front end. Therefore, I will spend more time in the DSP installment discussing RSP operation than I will here. The RSP will be set up to provide I/Q data at a sampling rate of 16 kHz for SSB/CW and 40 kHz for the wider-bandwidth modes (AM/FM). The goal was to get better than 100-dB out-of-band rejection. The high decimation rates make this a challenge in the RSP, but it can be done with the chosen output rates. I'll go into my thoughts in this area more in the DSP installment.

Summary

In this installment, we looked at the design details for an analog front end that should meet the requirements laid down in Part 1. Some of the ideas have been seen in these pages before. The key, as always, is to ensure that the tradeoffs are managed to allow maximum performance.

The AD6645 ADC is at the heart of the receiver architecture. This part has excellent noise and distortion performance, which allows this architecture to work. Low aperture jitter helps to minimize the ADC equivalent of reciprocal mixing, further enhancing performance. Further improvements of ADCs in the future will allow even better system performance, and the modular design will allow us to take advantage of this. For now, however, the performance available from the AD6645 is the state of the art.

In the next installment, we will look at the transmit side of the transceiver. Just as the receiver has been helped by improvements in ADCs, newer high-speed DACs allow excellent transmitter performance. The details of integrating them into the transmitter will be covered in Part 3.

Acknowledgements

I would like to thank Doug Smith, KF6DX; Paul Smith and Gary Hendrickson for their input on topics related to this article.

Notes

[1]J. Scarlett, KD7O, "A High-Performance Digital Transceiver Design, Part 1," *QEX*, Jul/Aug 2002, pp 35-44.
[2]W. Sabin, WØIYH, "Narrow Band-Pass Filters for HF," *QEX*, Sep/Oct 2001, pp 13-17.
[3]J. Stephensen, KD6OZH, "The ATR-2000: A Homemade, High-Performance HF Transceiver, Part 2," *QEX*, May/Jun 2000, pp 39-51.
[4]Available at **www.ansoft.com**.
[5]J. Stephensen, KD6OZH, "A Stable, Low-Noise Crystal Oscillator for Microwave and Millimeter-Wave Transverters," *QEX*, Nov/Dec 1999, pp 11-17.
[6]Analog Devices, *High Speed Design Techniques*, 1996, pp 5.41-5.48.

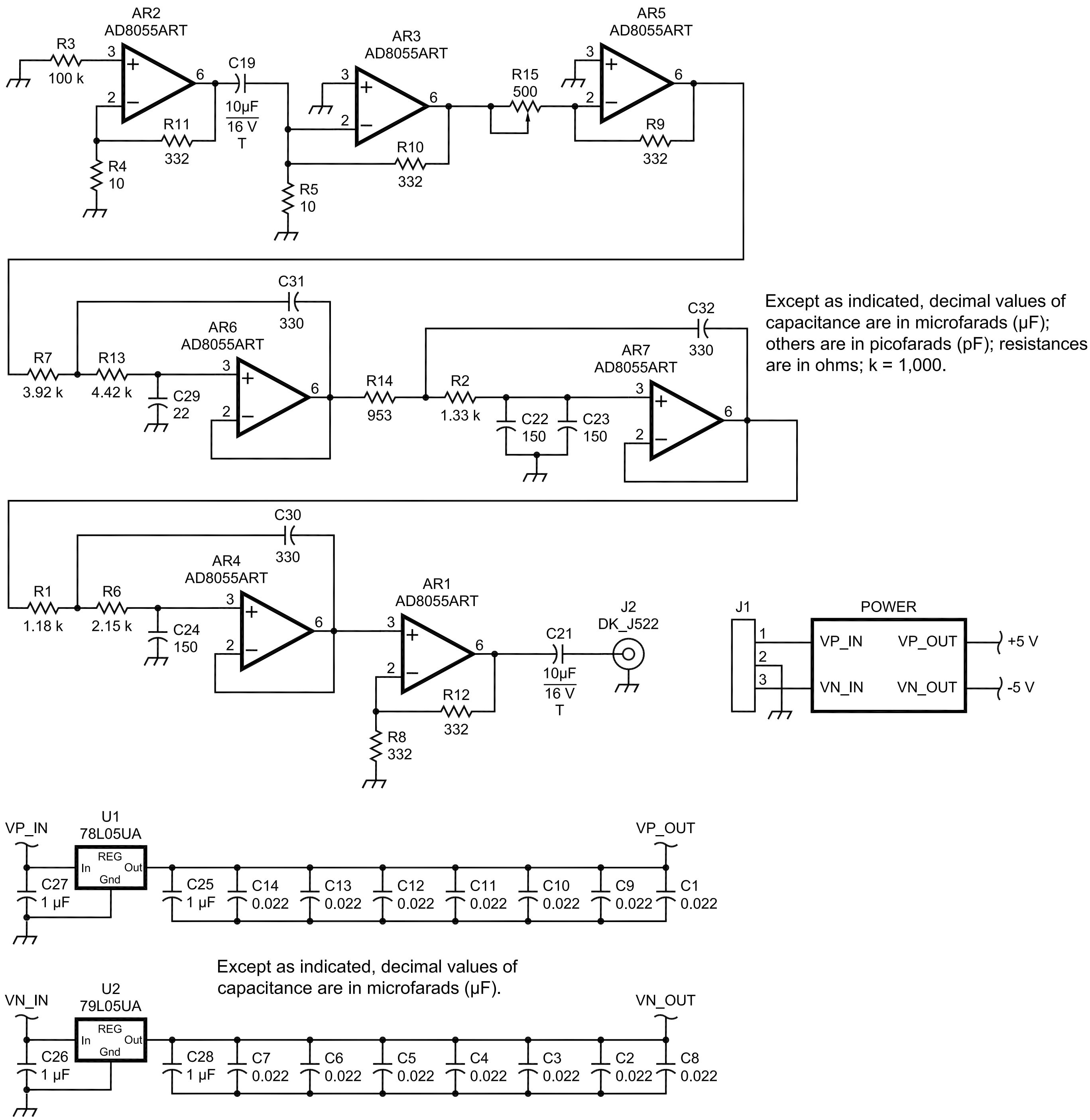
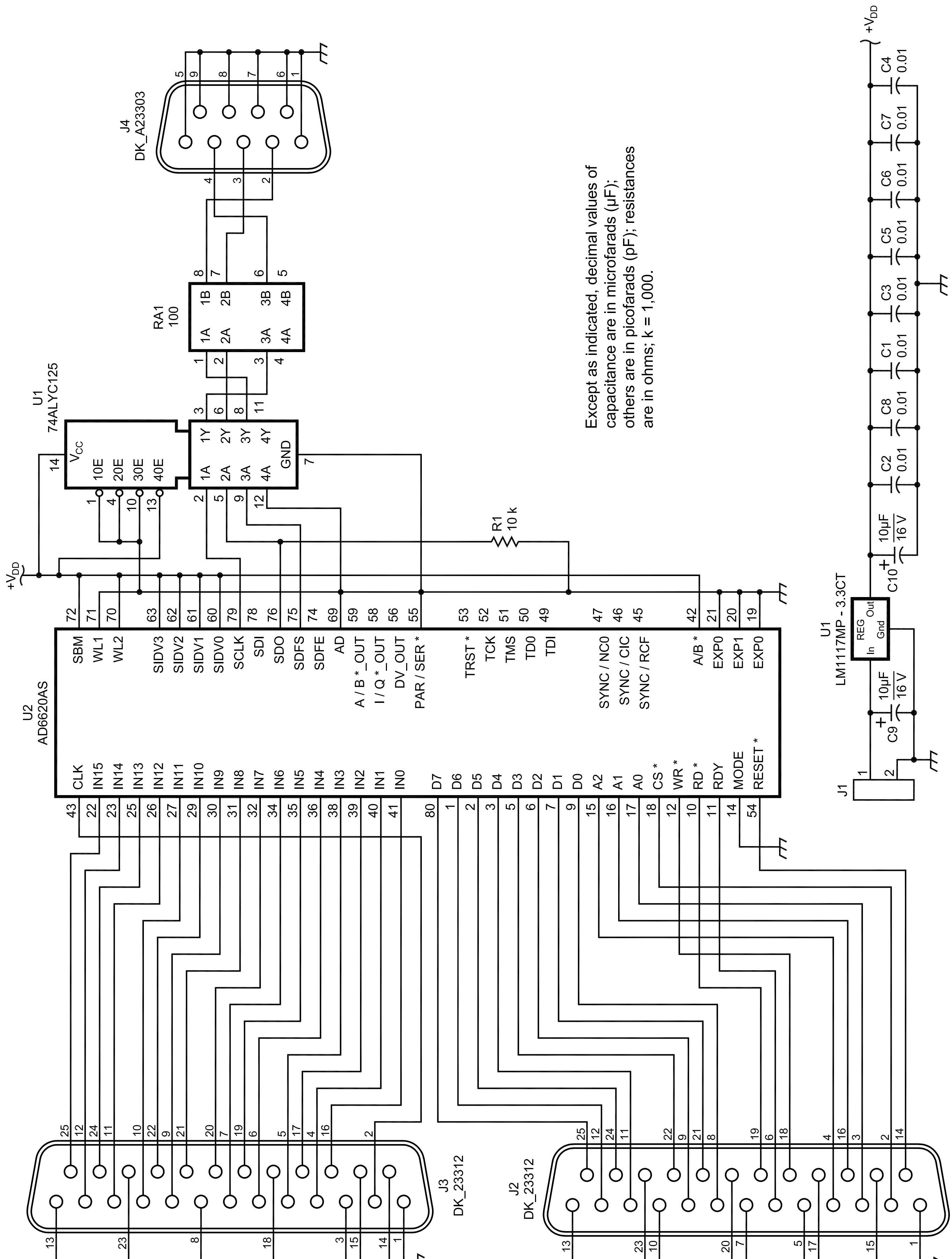


Fig 5—Dither-generator schematic diagram. Resistors and capacitors are 0805 SMT unless noted otherwise.

AR1-7—High-speed op amp, AD8055ART
C19, 21—10 µF 16-V tantalum capacitor
C25-28—1 µF, 16-V, X7R, 1206 SMT capacitor
J1—3-pin header
J2—PC mount SMB bulkhead jack (Digi-Key #J522)
R15—500 Ω, 10-turn trim pot (Digi-Key #3214W-501ECT)
U1—5-V voltage regulator (positive), SOT89, 78L05UA
U2—5-V voltage regulator (negative), SOT89, 79L05UA

Fig 6(right) —Receive Signal Processor schematic diagram. Resistors and capacitors are 0805 SMT unless noted.
C9, C10—10 µF, 16-V tantalum capacitor
J1—2-pin header
J2, J3—DB25 right angle, PC mount (Digi-Key #A23312)
J4—DB9 right angle, PC mount (Digi-Key (#A23303)
RA1—Resistor array, 4×100 Ω (Digi-Key #742C083101)
U1—Quad buffer, 74ALVC125PW
U2—RSP, AD6620AS
U3—3.3-V low-dropout voltage regulator, Analog Devices ADP3338AKC-3.3


Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
U2 AD6620AS
U1 74ALYC125
RA1 100
R1 10 k
J4 DK_A23303
J3 DK_23312
J2 DK_23312
U1 LM1117MP - 3.3CT
J1
C9 10µF 16 V
C10 10µF 16 V
C2 0.01
C8 0.01
C1 0.01
C3 0.01
C5 0.01
C6 0.01
C7 0.01
C4 0.01

Phase Noise and ADC Performance

There was some confusion from the discussion of clock jitter in Part 1 of this series. This led me to reevaluate how the material was presented and is the reason for this sidebar. One of the problems was making a connection between phase noise and clock jitter. Therefore, this discussion will be primarily focused on phase-noise requirements, and I will only discuss the relationship with clock jitter as necessary.

Rather than generating arbitrary jitter specifications, let's look at actual system requirements and translate this to a clock phase-noise requirement. Measurements are usually termed "noise-limited" when reciprocal mixing has increased the system noise figure by 1 dB. In our case, this would happen if the SNR of the converter increases by 1 dB.

Initial measurements with an evaluation board for the AD6645 showed a value for the SNR at 30 MHz of about 75 dB. We'll use this number for our calculations. We'll also do the calculations for 10 meters, since Eq 2 in Part 1 shows that jitter has a greater effect on higher frequency signals. For this discussion, we'll ignore that jitter on the clock or input signal during the SNR measurement may have had a significant contribution to the 75 dB SNR and assume this measurement was "jitterless."

In order to keep the SNR from increasing by 1 dB, the SNR contribution from jitter must be better than 81 dB. Thus, over the Nyquist bandwidth ($^1/_2$ of the sampling rate), the integrated noise due to jitter must be less than –81 dBc. For the purposes of this discussion, the carrier is assumed to be full-scale (approximately –13 dBm at the receiver input on 10 meters).

Using Eq 1 from Part 1 (processing gain), the required noise applied by the clock to the incoming signal must be less than –156.1 dBc/Hz. The clock rate is 64.96 MHz in this calculation. However, this is not the required clock noise performance, because of the relationship

$$SNR = 20 \log[\omega_{clk} / (\sigma_{clk} \omega_{sig})] \qquad \text{(Eq 1)}$$

where σ_{clk} is the frequency jitter of the clock and equals $\sigma_t \omega_{clk}$. This relationship shows that if the potential interfering signal is lower in frequency than the clock, the noise sidebands are improved based on the ratio. Likewise, noise sidebands applied to a signal that is higher in frequency than the clock are worse than the clock itself.

Thus, for a 64.96-MHz clock and a 28.5-MHz incoming signal at full scale, the clock noise requirement is –149 dBc/Hz to meet our noise specification. That is, the average noise level integrated over our desired bandwidth at a given offset from the interfering signal must be better than –149 dBc/Hz. Assuming a constant sideband slope (not true at PLL corner frequencies), that would mean that the noise must meet this specification at the center of the desired passband. We can use this to determine how our clock's predicted performance measures up.

Also, notice that the above requirements are for 10 meters. The requirements would be less stringent at 20 meters (about 6 dB), because jitter has a smaller effect at lower frequencies. Likewise, the requirements would be much more difficult at 2 meters (about 14 dB). This is offset by the fact that close-in signals tend to be much stronger at HF than on 2 meters. This is important as I consider whether to adapt this radio for 2 meters (undersample the third Nyquist zone) or to simply use a transverter.

On a separate note, by solving Eq 1 for jitter, we find that the jitter requirement is 0.5 ps at 10 meters, not the 0.1 ps example used in Part 1. At 2 meters, meeting the same specifications would require 0.1 ps jitter.

Clock Performance

Table 1 shows the expected performance of the receive PLL. The VCXO and reference were modeled using 'typical' expected values for crystal parameters (except the maximum *R*s of 40 Ω for the VCXO crystal and 25 Ω for the reference crystal) and the maximum specification for FET resistance. This should give a reasonably conservative model. I did the model using Leeson's equation on an *Excel* spreadsheet.

The noise floor within the PLL was determined using the ADF4001 datasheet. The specified noise floor for the PFD is –153 dBc/Hz at a comparison frequency of 1 MHz. Using a reference frequency of 12.992 MHz (with a 10 log relationship) and a multiplication of 5 (with a 20 log relationship) results in a PLL noise floor of approximately –128 dBc/Hz.

At offset values above the loop frequency of 230 Hz, the VCXO noise dominates, and below about 100 Hz, reference noise does. Between these offsets, the PLL is the primary noise source.

The table shows that we meet the required noise specifications at an offset of about 1200 Hz. This means that a *full-scale* signal outside of the 2400 Hz SSB passband will have no significant reciprocal mixing effects. For a 500-Hz CW bandwidth, a full-scale signal offset less than 1200 Hz from the center of the passband can cause reciprocal mixing, but an S9+40 signal outside the passband would not cause noticeable degradation. For the lower HF bands, these results are even better, while at 2 meters, a 5 kHz offset would be required for no significant effects from a full-scale signal.

This example shows a distinct advantage to this receiver architecture with respect to reciprocal mixing. Since the tuning is done digitally, we can use a low-noise crystal oscillator that drastically reduces reciprocal mixing effects. Between this and the outstanding noise and jitter performance of the AD6645 converter, we can virtually eliminate reciprocal mixing as a problem in our HF receiver. This example also demonstrates why reciprocal mixing from close-in signals is so difficult to tackle in receivers with wide-band, general coverage VCOs.

Table 1—Predicted Receive PLL Phase-Noise Performance

Offset (Hz)	*Phase Noise (dBc/Hz)*
10	–95
20	–104
50	–115
100	–125
200	–128
500	–138
1 k	–147
2 k	–155
5 k	–163
10 k	–167
20 k	–169
50 k	–170
100 k	–171

An Introduction to Intellectual Property Law

Here is a primer on patents, copyrights and trademarks: constitutionally provided protection for the products of your inventive mind.

By Dan Handelsman, N2DT, and Doug Smith, KF6DX

Having been through the process of obtaining patents, copy rights and trademarks, we wish to pass along some information about obtaining protection for your intellectual property. Since there appears to be a great deal of misunderstanding about your intellectual-property rights under law, we would like to dispel some of the myths surrounding this subject and explain it for the neophyte as clearly as we can.

Neither of us has applied for as many patents as some *QEX* readers have, so we welcome your feedback and suggestions for those who may be diving into the process for the first time. You beginners are the special target of the following discussion.

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What Is This Intellectual Property Business, Anyway?

Intellectual property lacks the tangible characteristics of other property such as real estate, stocks, bonds and cash. It does consist of one's ideas that have been reduced to some tangible form, which, after satisfying certain legal criteria, may be protected. Its definition is therefore somewhat evanescent since what is intellectual property and what is not may be difficult to determine in many cases. Additionally, we propose to discuss three classes of such property: patents, copyrights and trademarks. Those classes have been traditionally discussed together; but they are actually very different, though they share a common basis in law.

We all use patented items, read copyrighted material and purchase trademarked brands; but we seldom have reason to think about the repercussions of using or improving on those items or materials. We shall go into those issues and try to show how the law is both distinct and ambiguous. As with most legal issues, priority is paramount. We shall show some ways to protect yourself against legal challenges to your claims.

Patents: Permissible Legal Monopolies on Technology

The US Constitution provides for the promotion of "…the progress of science and useful arts by securing for limited times to authors and inventors the exclusive right to their respective writings and discoveries…. "[1] Other countries have similar provisions.

The basis of US patent law may be traced to the codification of English Common Law in the Statute of Monopolies, adopted in England in 1623. This enabled a contract between a royal monarch and the inventor. The name of the Statute is important since it illustrates clearly the constant interplay, even up to the present, of opposing forces in the law. The law grants limited monopolies to foster specific ends while, at the same time, it prevents greater, unlimited monopolies that are thought to be harmful to society or the economy.

Under English Common Law, the royal privilege of a patent was granted in return for certain benefits to the crown.[2] Such patents eventually became burdens on free trade because of onerous and impractical requirements forced on an inventor as a precondition of the royal grant. By the time of Queen Elizabeth, common-law courts were able to declare the most egregious abuses of the royal system unlawful and their rulings were subsequently codified by legislative actions that eventually dominated the area of intellectual property law.

The Statute of Monopolies effectively terminated those monopolies that significantly affected competition in trade of staples and basic commodities. While allowing patents having an essentially monopolistic character, that law placed a limit on those that claimed overly broad scopes. Legislators also struggled with the trade-off between native inventions and those imported from other countries. The result of those legislators' reasoning was the idea that granting a monopoly on a new idea encouraged inventiveness, no matter the source.

The quote above, from Article 1, Section 8, Clause 8 of the US Constitution "enabled" Congress to enact laws in keeping with its intent. Being constitutionally derived, laws affecting patents, unlike those affecting trademarks, require no further constitutional references for enactment and are wholly within the peremptory powers of the federal government, thus superceding any state or local laws. The enabling language for trademarks is more vague and comes from a legislative interpretation of the text of Article 1, Section 8, Clause 3—the Interstate Commerce Clause—authorizing Congress to regulate commerce with other nations and among the States.

The first congressional action on patents occurred in 1790. Since then, there have been only three major revisions to the Patent Act: in 1793, 1836 and 1952. Courts have made their impact on patent law since those early days; but only Congress can change the laws.

Patent Basics

In a nutshell, the Patent Act requires that an inventor show that he or she has produced something new and different—or *novel*. That something must be useful—having some *utility*—and must be not be obvious to a person reasonably educated in the field to which the invention pertains. Such a showing must be made in a patent application.

One of the myths about patents is that you can patent something that was invented by someone else. That is wrong: Only the original inventor can obtain a valid US patent on an invention. The patent may then be assigned to someone else—an employer, for example.

The key elements of an invention are listed as claims under the preceding specification. The claims point out how the invention meets the criteria for patentability.

A downside to the application process is a result that is anathema to the intent of the law. While the central purposes of the patent system are to encourage further development by making patent documents public and to foster the expansion of knowledge and trade, the language of the patents may be used to deliberately obscure their true meaning and import. This is a product of two processes: the need for unfortunate legalisms in the language of a patent and the desire of applicants to cast a wide descriptive net that would protect against variations and improvements on their designs by others.

Once a patent application is received at the US Patent and Trademark Office (USPTO), an examination process begins. During that process, it may become apparent that other patent applications being considered interfere with the current application. Interference between patent applications results in a proceeding that tries to determine which application first reduced the invention to practice. That question of priority comes before the Patent Office Board of Interference Proceedings.

Eventually, a patent may be granted that gives the inventor the right to exclude others from making, using, selling or importing the invention for a period of 20 years from the date of application. Patents cannot be renewed, but their terms can sometimes be extended slightly. Many inventors, such as those working for pharmaceutical companies, often improve on their own inventions to lengthen the term of their protection. By the time the 20 years of an original patent is up, for example, they may have thought of a different formulation for a drug that constitutes a significant improvement on the earlier design.

After expiration, a patented invention enters the public domain. The patentee thereafter loses the exclusive right to the invention and anyone has the right to make or sell it.

Bear in mind that the original law was intended to encourage the use and promulgation of inventions and that "working" the patent was encouraged. Hiding the invention is contrary to the spirit of the original law. Although the patentee is not required to work the patent or allow others to do so, the law encourages the inventor to market his or her invention and to make its benefits available to the public.

Recently, the US has allowed the filing of a provisional patent application. The provisional filing costs less than $100 for individuals or small entities at the time of this writing. It allows inventors to make an early declaration of their inventions without making detailed claims. Provisional patent applications are not examined by the USPTO for their merits. One has 12 months after the filing of a provisional patent application to file a regular patent application to obtain the benefits of the provisional filing. A benefit of the provisional filing is the establishment of priority. Another is the ability to declare "patent pending" on the invention.

Below, Dan relates his experience with his latest patent application. You can learn from his experience. We shall also use the fact that some of the text of the next section is an excerpt from *antenneX* to illustrate the concept of copyrights, which follows.

Patent the Fruits of Your Dreams

I applied for a patent on my Prismatic Polygon[3] antenna in May, 2001, received a notification that it had been granted on January, 2002, and received the patent, number 6,400,367, on June, 2002. What was unusual about it, and a pleasant surprise to me, was that the approval came through so rapidly. Why was I surprised? It was because my reasonable expectation, based on what everyone had told me about the process, was that it would take years—waits of five years or longer are not unusual.

Doug contacted me to ask my opin-

[1]Notes appear on page 19.

ions about various strategies to use in patenting one of his inventions. He wanted to see how much he could do by himself to save money since the process can be quite expensive. Doug also mentioned that he had gotten some of his information on the Internet. That interested me and I will go into my experiences in detail a little later in this article.

At about the same time, Jack Stone, Editor and Publisher of *antenneX*, the on-line magazine devoted to experimentation with antennas, requested that I write down my thoughts and memories about the patent experience while the whole process was still fresh in my mind. The article, "Patenting One's Dreams," was published in *antenneX*.[4]

I agreed to do so, since the patent process is arcane and daunting. There are many myths about it and much misinformation. After I decided to go along with Jack's request, I did what I had not done before–and what most of you would do in the first place. I followed Doug's example and did an Internet search on how to go about getting a patent.

Internet Patent Information

I am leery about what is on the Internet. As a professional in the field of medicine, I have found that the Internet is a fountain of misinformation. The owners of many Web sites have axes to grind but frequently hide their bias under neutral-sounding domain names. The problem is that no one has yet invented a truth filter that can distinguish information from misinformation—it would definitely be a patentable and lucrative discovery!

I did a Google search and waded into many URLs to see which were helpful and reasonable. That is, I tried to correlate what I had found with the reality that I had experienced. Many sites are owned by patent attorneys and you can guess their biases. Some were actually quite truthful and reasonable. Fortunately, virtually all the information that you may need can be found at the US Patent and Trademark Office Web site.

What Intellectual Property Can Be Patented?

The USPTO allows filing for patent protection for three kinds of intellectual property. Two of these, design and plant (flora) patents, are easy to obtain. The first is to protect the physical designs, or shapes, of real objects such as lamps, sofas, telephones and various geometric shapes. Plant patents apply to new species and hybrids.

The ones that we commonly understand to be patents for new inventions are in the third category, utility patents, and these must show both novelty and utility.

Novelty

Whatever product or process you come up with, it must be novel to be patented. The term novel means that is must be new and substantially different from any *prior art*—that being the term for an earlier invention in the vernacular of patent law.

Utility

This term means that patented inventions should be useful, or, to comply with the intention of the enabling section of the Constitution, that they have some economic benefit to the nation or society as a whole. So some written justification is necessary to show the utilitarian merits such as efficiency, better performance than earlier products or lower cost of manufacture.

Patent Searches

More than six million patents have been granted in the US alone. Although you know that your idea is truly novel, it is imperative that you see what has been done before. The fact that someone else has patented something similar to yours may not be evident in textbooks or common lore. On the other hand, the basis of a conflicting patent may be in a field far removed from yours. You must do a prior-art search.

An example of the pitfalls of the (not necessarily) young and naive is a tale of what happened to me when I was a teenager. After watching the 1960 Olympics in Rome, and being a competitive swimmer at the time, I saw a fiasco in the timing of the Mens' 100-meter freestyle. The timers had given the American, Lance Larson, first place but the judges had awarded the Gold Medal to the Australian, John Devitt because they thought they saw him touch first. Devitt touched above water level and Larson had touched underwater.

Since I was already a ham and knowledgeable, or so I thought, in switching circuits, I applied for a patent on a set of simple waterproof sensors which triggered sequential latching circuits to determine order of finish. The upshot of this was that a search found a similar latching circuit patented by Bell Telephone for their own use. I was fortunate that the attorney was a friend of the family's and he only charged me his cost for the search.

When the time came to patent my antenna, the Prismatic Polygon, I was somewhat wiser. I checked all of the textbooks that I could get my hands on. I then went to the IBM patent Web site, which is no longer available, and searched the patent database using every possible description of the anatomy of the antenna, its feed and its utility—which happens to be an extremely wide bandwidth.

Fortunately, the USPTO now has a Web site where you can search its entire data base.[5] The site has a tremendous amount of helpful information about the entire patent process. It should be the starting point for anyone interested in obtaining a patent.

At any rate, I examined hundreds of patents to make sure that my idea was novel and had utility. The latter is important since there is no point in going ahead even if your idea is truly novel, but someone else has a product which does the job better or more cheaply. Also, certain libraries around the US—mentioned at the USPTO site—have the bound texts of all patents granted in this country and in some other areas of the world.

So You Have Something Novel: What Next?

After you have convinced yourself that your invention is truly new and different, you might decide to go ahead and patent it. However, suppose that you have sent your idea around to your friends and—perish the thought—you published it already. How do you deal with that?

Write down everything that you have thought of, sign it, date it and have it witnessed by others who are not family members or friends. A notary public is handy here. Mail yourself the pertinent documents and keep the unopened and post-marked envelope in a safe place. Keep copies of all e-mails referring to your invention. This is the most important thing you can do to set your priority over the invention.

Priority

Assuming you have the proper records indicating when you created an invention, you have priority under US law. In this country, the date of the invention is the critical component. In Europe and some other countries, priority is set by the date of filing the application, regardless of the identity of the inventor or the date of the invention. In case you have published the idea, you have one year from the time of publication to file a patent in the US.

The problem is to balance the need for advice against the need for secrecy. That is the point of discretion. Your intellectual property might be able to be improved upon or you, like me, might need help in figuring out how it works. Better explanations of the utility, construction and the underlying theoretical basis of your design increase your chances in the patent process. I am eternally grateful to L. B. Cebik, W4RNL, with whom I shared my early ideas and designs. His personal antenna database and experience are much greater than mine. Only after many discussions with him did I convince myself that I had stumbled on something truly novel and worth embarking on the adventure of the patent process.

Does It Work?

Testing your invention is not strictly necessary under the new US rules. In the old days, a working model was needed for approval. Now only drawings are necessary, but reduction to practice is still a requirement. The USPTO wants to know that you have actually done what you say is possible. I found myself reluctant to go ahead and invest any money in the patent application until I was certain that the Prismatic worked. I built many prototypes, bought test equipment for HF and managed to get access to some very expensive and sophisticated equipment at some local labs for VHF/UHF testing.

Can You File a Patent Application by Yourself?

Probably not. The language of patents is beyond "legalese." I almost could not recognize the description of my antennas in my patent application. The Prismatic Triangle is described as having "… a plurality of three radiators fed by a plurality of three transmission lines...." It probably takes years to learn the jargon and make acceptable gibberish. So the issue is to see how much you can do for yourself to make the process as rapid and economical as possible.

The most expensive way is to go to a patent attorney, describe the invention to him and hand him some rudimentary drawings. He will charge you by the hour and charge for each drawing. The clock starts running with every telephone call.

What I did was describe everything in detail. I began with the historical background of the antennas on which I based my invention and how they worked. I then described exactly how the antenna was built: the dimensions, the feed arrangement and the materials. I then outlined the basis for the novelty of the structure and how the wide bandwidth was obtained. Along with the text, I added and referred to schematics generated with *NEC* and various graphs outlining how the antenna worked. The entire text that I handed to my lawyer amounted to 20 pages, single-spaced.

I subsequently found that he had taken the text as it was and then added the suitable jargon to make it a workable application. I had to do only a single edit on the final document. He then had my drawings redrawn to make them conform to the format required by the USPTO. Their requirements for drawings are quite specific (visit their Web site, address given in Note 2) and those not conforming will be rejected.

The extent and thoroughness of your prior-art search also reduces the cost of the search that a professional search company must do to satisfy the attorney. If you have a record of what you have done by yourself, what search terms you used and where you looked, that reduces the amount of time the professionals need to do their own search. They have more sophistication and facility in finding things that you may have missed. This search is not a waste of time since it can abort the application process before you put a lot of money into it—if it is destined to fail.

Speaking of money, a patent is going to cost you something. The question is how much. Patent applications can cost anywhere from $5000-$15,000 or more. Remember, you are dealing with attorneys and, although it breaks my heart since I am one also, I must warn you that every single one of them will try to pad his or her billable hours. The one thing into which you do not want to go is an open-ended contract.

Limit your liability. It is imperative that you negotiate all aspects of the application process. I asked for my attorney's retainer price based on supplying him with completely detailed text—as mentioned above—and detailed drawings. Into this total went the amount necessary for the professional search, the cost of the application and other filing fees and the estimated cost of corrections and edits.

The trick now is to prevent the attorney from adding to that cost. I have learned over the years to keep a log of all contacts—each and every single contact—with my attorneys: time in, time out, material discussed, what is promised, what next and so forth. I am not saying that you should have a tape recorder running, but it may not be a bad idea as long as you inform the other party of what you are doing. If you are dumb enough not to do so, you will be surprised as to what additional expenses are billed to you: You would have no means of refuting any of them.

I had a telephone call to me where the attorney apologized for the delay in my application. The call turned up later in a bill—believe it. You must take prudent steps to have proof to refute the charges.

The Scope of a Patent Application

By now, you are aware that future copycats may try to take a patented design and modify it sufficiently to try for their own patent. The first thing one must do is to try to determine the scope of his or her patent application and how many variations and permutations you wish to cover to prevent this from happening.

I knew that my antenna's unique properties were the result of both the coupling of identical and parallel radiators and the feed method. I therefore decided to include both items as intellectual property. In fact, the feed method, which I originally thought was not the most important element in the design, turned out to be critical, and it is the basis for the entire class of antennas. I had tested designs involving from two to four radiators but had modeled antennas with up to 12 radiators: P12s or Prismatic dodecahedrons. I had also figured out how to stack and feed each of these as individual elements in an array.

My application therefore included a discussion of all the Prismatics that I had tested and modeled but stated that the design was applicable to n radiating elements where n could be a number from two to infinity. It also stated that each of these could be used as an individual element in an array of n elements.

In addition to the above, I established the two factors regarding the utility of the designs; the wide bandwidth and the usefulness of their radiation patterns beyond a bandwidth of two octaves.

To summarize: Include in your application the specific geometric, mechanical and constructional properties of the product, possible design variations that you can think of at the moment and possible ways that you, or someone else, may modify the design in the future. In addition, carefully differentiate your invention from all prior art, not only based on its constructional or design elements but also based on its novel properties and its utility.

What Next?

File and receive a tracking number that identifies your application up to the point that the patent is approved and, if you succeed, file again, with a hefty application fee, for a patent number. Once you have filed originally, you are protected if the patent is eventually granted. At this point, it is safe to publicize it as I have done with the Prismatics[6] and share the basic ideas with investors and manufacturers. It is still a good idea to hold back some proprietary information that enables one to maximize the properties or cut the construction costs of the invention.

You may wait years. The first delay is in getting the application reviewed in the first place. The patent examiner must evaluate how similar or different your art is from prior art. The chances that there will be objections to some aspects and a resulting rejection are high—I think they are in the 70% range or more. You then have a right to refile and explain away any of the objections, if you can. This process of rejection and refiling can go on for a long time, I am told. Ultimately, about 70% of patents are granted.

Eventually, the waiting will be over. You get lucky or wear them down and the application is approved: time to cough up more money. Formal drawings must be done. You can see the exact requirements at the USPTO Web site. I am investing in a CAD program because I always wanted one and also because I can save a lot of money by doing my own drawings instead of having them done by a professional draftsman. I cannot avoid about $1500 in filing fees however. Speaking of filing fees: I now have one year from the time my application was originally filed in May, 2001 to file for a patent in both Japan and the EEC.

Lastly, expect to pay a maintenance fee to keep your patent good during its life. The fee increases as the patent matures.

Confidentiality of the Application

For the first 18 months after filing, the application is totally confidential and not subject to the provisions of the Freedom of Information Act. During that time, other parties cannot access the details of your invention. Subsequently, anyone may look up the contents of your application.

Limits of the Protection Granted by Filing a Patent Application

Filing a patent application does not totally prevent some other party from manufacturing your invention without infringing on your patent. The time frame for infringement technically begins when the patent is granted. If you publish or manufacture your design before the patent is granted, another party may manufacture it and profit by it. This could be done from reading what you have published or by "reverse engineering" a design you are marketing.

What limits the likelihood of this type of short-term "legal infringement" is that the other party does not know when the patent will be granted and, at that time, the infringement becomes legally culpable. As an example, if a party decided to build my antenna, he would have had to invest in materials, advertising and production. This would take some time and, since the patent was granted in eight months, it is probable that most, if not all, of the profits garnered from the manufacture of the antenna would have been subject to a suit for infringement.

Patent Summary

I have tried to give you an idea of what I went through in patenting my Prismatic antennas. Everything about the process is identical to what everyone else has gone through, and will go through in the future, except for my good fortune in obtaining quick approval. I am sorry that I cannot share my magic formula for it, since I haven't a clue as to why it happened the way it did.

Copyrights

A copyright is a body of legal rights that protect creative works from being reproduced, performed or disseminated by others without the permission of the holder of the right. Examples of materials that can be copyrighted are: literature, music, advertisements, paintings, photographs, graphs, movies, certain maps and, as of 1980, computer software. Notice that the last involves more than just a complete package that cannot be duplicated without the holder's permission. The guts of the code must be separately protected by a patent to prevent another party from simply rewriting some sections and creating a new product.

The classical form of copyright violation is plagiarism. Most commonly, this is seen when sections of text from one author's work are lifted by another author without proper permission. You can see similar copyright violations with music, but the issues there may be more complex. How many notes are involved? Does the identity of a chord prove plagiarism? What if an entire refrain is the same? Of course, there is seldom a problem if there is proper attribution: Paganini's "Variations on a Theme by Bach," as a hypothetical example. Nor is there a problem if the music were lifted from a centuries-old composition.

The main misconception about copyrights is that they must be registered with the government to be valid. In fact, a copyright may be declared the moment a work is complete. Copyright is available to protect both published and unpublished works. Under the 1976 Copyright Act, a copyright gives the author the right to exclude others from:

- Reproducing the work in copies or recordings;
- Preparing derivative works based upon the original work;
- Distributing copies of the work to the public, including via digital transmission;
- Performing the work publicly;
- Displaying the work publicly.

Those rights are not unlimited in scope, though.

One limitation to your rights under copyright law is called the "fair-use doctrine," given statutory basis in Section 107 of the 1976 Copyright Act. Among other things, fair use means that the literary reviewer of the New York Times, for example, can quote excerpts from your work and make commentary on them. Another limitation to copyrights is that of "compulsory license," whereby published works may be played or performed with appropriate payment of royalties or other compensation previously specified.

In the case of works made for hire, the employer and not the employee is considered to be the copyright owner *if the parties involved expressly agree in a written instrument signed by them that the work shall be considered a work for hire.*[7] Many companies make employees sign such a waiver on employment. If you did not sign such a waiver or release, you may retain the copyright to your work! The authors of a joint work are co-owners of the copyright unless there is an agreement to the contrary.

Contributions to a periodical or other collective work are considered separate from the copyright of the work as a whole. Unless there is an agreement to the contrary, the original author retains the copyright to his or her work. For example, *QEX* authors are required to assign copyrights to their work to the ARRL, but they retain the right to distribute copies of the work as they see fit without remuneration, to post it to their Web sites and so forth. In other words, they cannot sell the thing to someone else, but they can give it away at will to individuals with proper credit to *QEX*.

Certain rules apply to what is eligible for copyright and what is not. The following items would not be candidates for copyright:

- Speeches, dances or improvisational material that is not written or recorded in fixed form,
- Slogans, titles or names that are ordinary or mere listings of ingredients or contents,
- Concepts or ideas that have no original authorship, such as calendars, unit-conversion charts or tape measures.

Securing and Declaring a Copyright

The way in which a copyright is secured is frequently misunderstood. A copyright is secured automatically at the moment a work is created in fixed form. For example, a song may be fixed by an audio recording or as sheet music. You must declare the copyright, though, to give notice to others of its existence.

Publication is no longer a requirement for copyright as it was under the Copyright Act of 1909. Before 1978, copyright was generally secured by the act of publication with notice of copyright, assuming compliance with other legal conditions. US works in the public domain as of January 1, 1978 remain in the public domain under the 1976 Copyright Act.

Under the present law in the US, works that are published in the US must be submitted to the Library of Congress. The law says you have to deposit your work with the government!

Notice of copyright is important because it identifies the copyright owner and date of first publication. Presentation of a confirmed copyright notice prevents parties in infringement lawsuits from giving any weight to defenses based on ignorance of the copyright, except as provided in Section 504(c)(2) of the copyright law.

Length of Copyright Protection

Works created (established in fixed form) on or after January 1, 1978 are automatically protected for the lifetime of the author and for 70 years after his or her death. In the case of a joint work created by co-authors, the protection lasts for 70 years after the last surviving author's death. For works made for hire, the protection extends to 95 years from publication or to 120 years from creation, whichever is shorter.

Works created before January 1, 1978 have been automatically brought under the statute and their terms are generally computed according to the formulas cited above. The law says that in no case shall the term of copyright expire before December 31, 2002 and for works published on or before December 31, 2002, the term of the copyright shall not expire before December 1, 2047.

Under the law in effect before January 1, 1978, copyright was secured either on the date the work was first published with copyright notice or on the date the work was registered if unpublished. In either case, the copyright endured for a first term of 28 years from the date it was secured. In the last year (the 28th) of the copyright, it was eligible for renewal. The Copyright Act of 1976 extended the renewal term from 28 to 47 years for copyrights that existed on January 1, 1978, or for pre-1978 copyrights under the Uruguay Round Agreements Act, making those works eligible for a total term of protection of 28 + 47 = 75 years. Public Law 105-298 further extended the renewal term of copyrights existing on the date the law was enacted by an additional 20 years, providing for a renewal term of 67 years and a total protection term of 28 + 67 = 95 years.

Public Law 102-307, enacted on June 26, 1992, amended the 1976 Copyright Act to provide for the automatic renewal of the term of copyrights secured between January 1, 1964 and December 31, 1977. That law makes registration renewal optional, so filing for renewal registration is no longer required to extend the original 28-year term to the full 95 years. However, filing a renewal registration during the 28th year of the original term achieves some benefits.

International Copyright Protection

Despite notices of "international copyright secured," there is no such thing as an international copyright that will protect an author's rights throughout the world. Protections in any particular country depend largely on the laws of that country. Many countries, though, provide protection to foreign works under the reciprocity conditions of a treaty.

Copyright Registration

Although it is not strictly necessary to obtain protection, copyright registration has its advantages. For example, lawsuits filed against infringement must show registration for works of US origin. If made within five years of pub-

The Digital Millennium Copyright Act of 1998 (DMCA)

The DMCA is Congress' attempt to bring copyright law in line with the information age. The main intent of the law is to prevent efforts to circumvent copy-protection schemes. Congress evidently did not quite fathom all the implications of what it wrote into the Act before they voted to approve it.

Among other things, the DMCA provides that:

1. Anyone accessing copyrighted material without the consent of the copyright owner is liable for criminal and civil penalties, thereby perhaps circumventing the fair-use doctrine.

2. It is illegal to import or sell technologies that exclusively seek to circumvent copy-protection schemes, with certain exceptions.

3. It is illegal to willfully remove or alter copyright management information, including the names of the author and copyright holder, terms of allowable use of the work and other conditions such as the Register of Copyrights may reasonably apply.

The restriction of clause (1) has certain citizens up in arms. They contend that the Constitution enabled Congress to allow the granting of limited monopolies on intellectual property in return for its free and public dissemination. Any law that prohibits one from actually learning the contents of copyrighted material without permission appears to go against that tenet. Some say the DMCA should be declared unconstitutional for that reason.

Libraries and writers are not happy with the DMCA because it appears to restrict their ability to provide information. A prior opinion from the Supreme Court seems to apply: "The author's consent to a reasonable use of his copyrighted works had always been implied by the courts as a necessary incident of the constitutional policy of promoting the progress of science and the useful arts, since a prohibition of such use would inhibit subsequent writers from attempting to improve upon prior works and thus ... frustrate the very ends sought to be attained." (Justice O'Connor writing on *Harper & Row v. Nation Enterprises*, 1985)

The US Supreme Court addressed the DCMA in a decision rendered in January 2003. Their holding was that, while the Act did run counter to the tenor of copyright law, it was a valid Act of Congress in its Legislative function and therefore not within the purview of the Court to reject it. That is, all amendments of the perceived inequities must be dealt with by Congress.

To some, the DMCA looks like it limits the rights of purchasers of tangible goods. For example, it may limit their right to privately display copyrighted works, to archive their software or make it compatible with their individual machines. Those folks say the DMCA goes exactly in the opposite direction of what was intended by our constitutional framers.

Whatever its implications, the DMCA is certainly controversial. It remains to be seen whether Congress will amend it or just ignore it. What do you think?
—Doug Smith, KF6DX

lication, registration will establish *prima facie* evidence of the copyright. If registration of the work were made within three months of publication of the work or prior to infringement of the work, statutory damages and attorney's fees would be available to the copyright owner in court actions. Otherwise, only actual damages would be available to the copyright owner. More information about copyrights can be found at the Library of Congress' Web site, **www.loc.gov**.

Trademarks and Service Marks

The original purpose of trademarks was to indicate the origin of goods. As early as the Middle Ages, statutes were enacted in England, the Middle and Far East supporting trademarks to protect the public against the sale of goods of questionable origin and quality. In the beginning, trademark law was intended only to prevent "palming off" the goods of one producer as those of another. Today, trademarks allow the buyer to assert his or her choice of brand and to distinguish products by name.

Congress enacted the first US trademark laws in 1870 and 1876. Those late dates indicate inattention by legislators to trademarks. They also must not have done a very good job, since the Supreme Court declared the laws unconstitutional in 1879. The court held that Congress had no power to regulate trademark matters in the way that they did, finding that those powers were reserved to the states. In response to that decision, Congress passed laws in 1881 and 1905 solely addressing interstate use of trademarks. In 1946, the Lanham Act was passed and signed into law. That act, primarily concerned with enforcement of trademark law at the federal level, constitutes the most recent major revision of trademark law.

A trademark is something that exists only with respect to the sale of goods commercially. Absent the sale of goods, a mark indicating origin is simply not a trademark. Rights to a trademark are thus acquired solely through priority of use. Somewhat like copyrights, trademarks may be declared without registration; but registration may be necessary in cases of interference and it is certainly valuable in lawsuits. A service mark is very much like a trademark except that it indicates the source of services and not goods.

What is Eligible For Trademark?

Words or signs that uniquely distinguish the goods of one producer from another are eligible for trademark as long as they are distinctive and meet requirements against prior use in similar markets. The requirements about similar markets include geographical and other limitations. For example, *QEX* could be the trademarked name of a magazine devoted entirely to Palm Pilots were its readership and market wholly different from a magazine devoted to communications experimentation.

Eligibility is determined primarily by usage. Courts find for the complainant who can show association between the buyer and the trademark in a particular market. Judgements involve decisions about whether buyers were deceived, confused or mistaken by the use of conflicting marks.

A descriptive term in a trademark is ineligible for protection unless it acquires secondary meaning. Secondary meaning means that a formerly descriptive mark has developed a quality that uniquely distinguishes the goods or services of the provider. For example, a trademark such as "bright finish" might apply to apples, automobiles and furniture alike. A trademark acquires secondary meaning through market association of the product with the term. If Buicks become known for their bright finishes, then trademark law applies. If Washington apples also acquire such an association, no interference exists because the markets are different.

The best association between a product and a trademarked name can be found with Aspirin. This was the brand name used by AG Bayer at the turn of the 20th century for acetylsalicylic acid. Over the years, the term became totally identified with the product. Clearly, although competitors could package and market the drug under various brand names, only the Bayer product could properly be marketed and sold as Aspirin.

The patents on forms of salicylic acid have long since expired and the term aspirin is now associated with the products of a great many manufacturers. It appears in virtually all dictionaries, and it has therefore lost its status as a trademark.

Conclusion

We have tried to show how US intellectual-property law applies to various cases. We welcome your feedback and look forward to your comments.

Dr Dan Handelsman, MD, JD, N2DT, has been an Amateur Radio operator since 1957 when he received his original call sign WA2BCG. He received his present call sign in 1977.

Dan is a physician with a specialty in Pediatric Endocrinology and is on the Clinical Faculty of the New York Medical College. In addition, he received his Law Degree from Pace University in 1988, and he is a member of the Bar in New York.

His major interest is in antennas and some of his prior designs have appeared in QEX. His first patent was received for the ultra-wide-bandwidth design called the Prismatic Polygon. At the present time, he is in the process of obtaining a patent for a compact antenna that considerably improves on the performance of the Compact or Magnetic Loop.

Notes

[1] *US Constitution*, Article 1, Section 8.

[2] A. Miller and M. Davis, *Intellectual Property: Patents, Trademarks and Copyright*, (St Paul, Minnesota: West Group, 1987).

[3] The antenna is now called the Prismatic Polygon and my company is called Prismatic Antennas, Inc. At the time of the patent application, I had named the antennas Three-Dimensional Polygons; but David Jefferies, PhD, G6GPR, came up with the better name and I adopted it.

[4] **www.antennex.com/archive5/Feb02/Feb402/patents.html**, a membership is required to read the article.

[5] Visit **www.uspto.gov**.

[6] **www.antennex.com**.

[7] *Op. cit.* 2, pp 378-380.

A Software Defined Radio for the Masses, Part 4

We conclude this series with a description of a dc-60 MHz transceiver that will allow open-software experimentation with software defined radios.

By Gerald Youngblood, AC5OG

It has been a pleasure to receive feedback from so many *QEX* readers that they have been inspired to experiment with software-defined radios (SDRs) through this article series. SDRs truly offer opportunities to reinvigorate experimentation in the service and attract new blood from the ranks of future generations of computer-literate young people.[1] It is encouraging to learn that many readers see the opportunity to return to a love of experimentation left behind because of the complexity of modern hardware. With SDRs, the opportunity again exists for the experimenter to achieve results that exceed the performance of existing commercial equipment.

Most respondents indicated an interest in gaining access to a complete SDR hardware solution on which they can experiment in software. Based on this feedback, I have decided to offer the SDR-1000 transceiver described in this article as a semi-assembled, three-board set. The SDR-1000 software will also be made available in open-source form along with support for the GNU Radio project on *Linux*.[2] Table 1 outlines preliminary specifications for the SDR-1000 transceiver. I expect to have the hardware available by the time this article is in print.

The ARRL SDR Working Group includes in its mission the encouragement of SDR experimentation through educational articles and the availability of SDR hardware on which to experiment. A significant advance toward this end has been seen in the pages of *QEX* over the last year, and it continues into 2003.

This series began in Part 1 with a general description of digital signal processing (DSP) in SDRs.[3] Part 2 described *Visual Basic* source code to implement a full-duplex, quadrature interface on a PC sound card.[4] Part 3 described the use of DSP to make the PC sound-card interface into a functional software-defined radio.[5] It also explored the filtering technique called FFT fast-convolution filtering. In this final article, I will describe the SDR-1000 transceiver hardware including an analysis of gain distribution, noise figure and dynamic range. There is also a discussion of frequency control using the AD9854 quadrature DDS.

[1]Notes appear on page 28.

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To further support the interest generated by this series, I have established a Web site at **home.earthlink.net/~g_youngblood**. As you experiment in this interesting technology, please e-mail suggested enhancements to the site.

Is the "Tayloe Detector" Really New?

In Part 1, I described what I knew at the time about a potentially new approach to detection that was dubbed the "Tayloe Detector." In the same issue, Rod Green described the use of the same circuit in a multiple conversion scheme he called the "Dirodyne".[6] The question has been raised: Is this new technology or rediscovery of prior art? After significant research, I have concluded that both the "Tayloe Detector" and the "Dirodyne" are simply rediscovery of prior art; albeit little known or understood. In the September 1990 issue of *QEX*, D. H. van Graas, PAØDEN, describes "The Fourth Method: Generating and Detecting SSB Signals."[7] The three previous methods are commonly called the *phasing method*, the *filter method* and the *Weaver method*. The "Tayloe Detector" uses *exactly* the same concept as that described by van Grass with the exception that van Grass uses a double-balanced version of the circuit that is actually superior to the singly-balanced detector described by Dan Tayloe[8] in 2001.

In his article, van Graas describes how he was inspired by old frequency-converter systems that used ac motor-generators called "selsyn" motors. The selsyn was one part of an electric axle formerly used in radar systems. His circuit used the CMOS 4052 dual 1-4 multiplexer (an early version of the more modern 3253 multiplexers referenced in Part 1 of this series) to provide the four-phase switching. The article describes circuits for both transmit and receive operation.

Phil Rice, VK3BKR, published a nearly identical version of the van Graas transmitter circuit in *Amateur Radio* (Australia) in February 1998, which may be found on the Web.[9] While he only describes the transmit circuitry, he also states, ". . . the switching modulator should be capable of acting as a demodulator."

It's the Capacitor, Stupid!

So why is all this so interesting? First, it appears that this truly is a "fourth method" that dates back to at least 1990. In the early 1990s, there was a saying in the political realm: "It's the economy, stupid!" Well, in this case, it's the capacitor, stupid! Traditional commutating mixers do not have capacitors (or integrators) on their output. The capacitor converts the commutating switch from a mixer into a sampling detector (more accurately a track-and-hold) as discussed on page 8 of Part 1 (see Note 3). Because the detector operates according to sampling theory, the mixing products sum aliases back to the same frequency as the difference product, thereby limiting conversion loss. In reality, a switching detector is simply a modified version of a digital commutating filter as described in previous *QEX* articles.[10, 11, 12]

Instead of summing the four or more phases of the commutating filter into a single output, the sampling detector sums the 0° and 180° phases into the in-phase (I) channel and the 90° and 270° phases into the quadrature (Q) channel. In fact, the mathematical analysis described in Mike Kossor's article (see Note 10) applies equally well to the sampling detector.

Is the "Dirodyne" Really New?

The Dirodyne is in reality the sampling detector driving the sampling generator as described by van Graas, forming the architecture first described by Weaver in 1956.[13] The Weaver method was covered in a series of *QEX* articles[14, 15, 16] that are worth reading. Other interesting reading on the subject may be found on the Web in a Phillips Semiconductors application note[17] and an article in *Microwaves & RF*.[18]

Peter Anderson in his Jul/Aug 1999 letter to the *QEX* editor specifically describes the use of back-to-back commutating filters to perform frequency shifting for SSB generation or reception.[19] He states that if, on the output of a commutating filter, we can, "…add a second commutator connected to the same set of capacitors, and take the output from the second commutator. Run the two commutators at different frequencies and find that the input passband is centered at a frequency set by the input commutator; the output passband is centered at a frequency set by the output commutator. Thus, we have a device that shifts the signal frequency, an SSB generator or receiver." This is exactly what the Dirodyne does. He goes on to state, "The frequency-shifting commutating filter is a generalization of the Weaver method of SSB generation."

So What Shall We Call It?

Although Dan Tayloe popularized the sampling detector, it is probably not appropriate to call it the Tayloe detector, since its origin was at least 10 years earlier, with van Graas. Should we call it the "van Graas Detector" or just the "Fourth Method?" Maybe we should, but since I don't know if van Graas originally invented it, I will simply call it the quadrature-sampling detector (QSD) or quadrature-sampling exciter (QSE).

Dynamic Range— How Much is Enough?

The QSD is capable of exceptional dynamic range. It is possible to design a QSD with virtually no loss and 1-dB compression of at least 18 dBm (5 $V_{P\text{-}P}$). I have seen postings on e-mail

Table 1—SDR-1000 Preliminary Hardware Specifications

Frequency Range	0-60 MHz
Minimum Tuning Step	1 μHz
DDS Clock	200 MHz, <1 ps RMS jitter
1dB Compression	+6 dBm
Max. Receive Bandwidth	44 kHz-192 kHz (depends on PC sound card)
Transmit Power	1 W PEP
PC Control Interface	PC parallel port (DB-25 connector)
Rear Panel Control Outputs	7 open-collector Darlington outputs
Input Controls	PTT, Code Key, 2 Spare TTL Inputs
Sound Card Interface	Line in, Line out, Microphone in
Power	13.8 V dc

Table 2—Acceptable Noise Figure for Terrestrial Communications

Frequency (MHz)	*Acceptable NF (dB)*
1.8	45
3.5	37
4.0	27
14.0	24
21.0	20
28.0	15
50.0	9
144.0	2

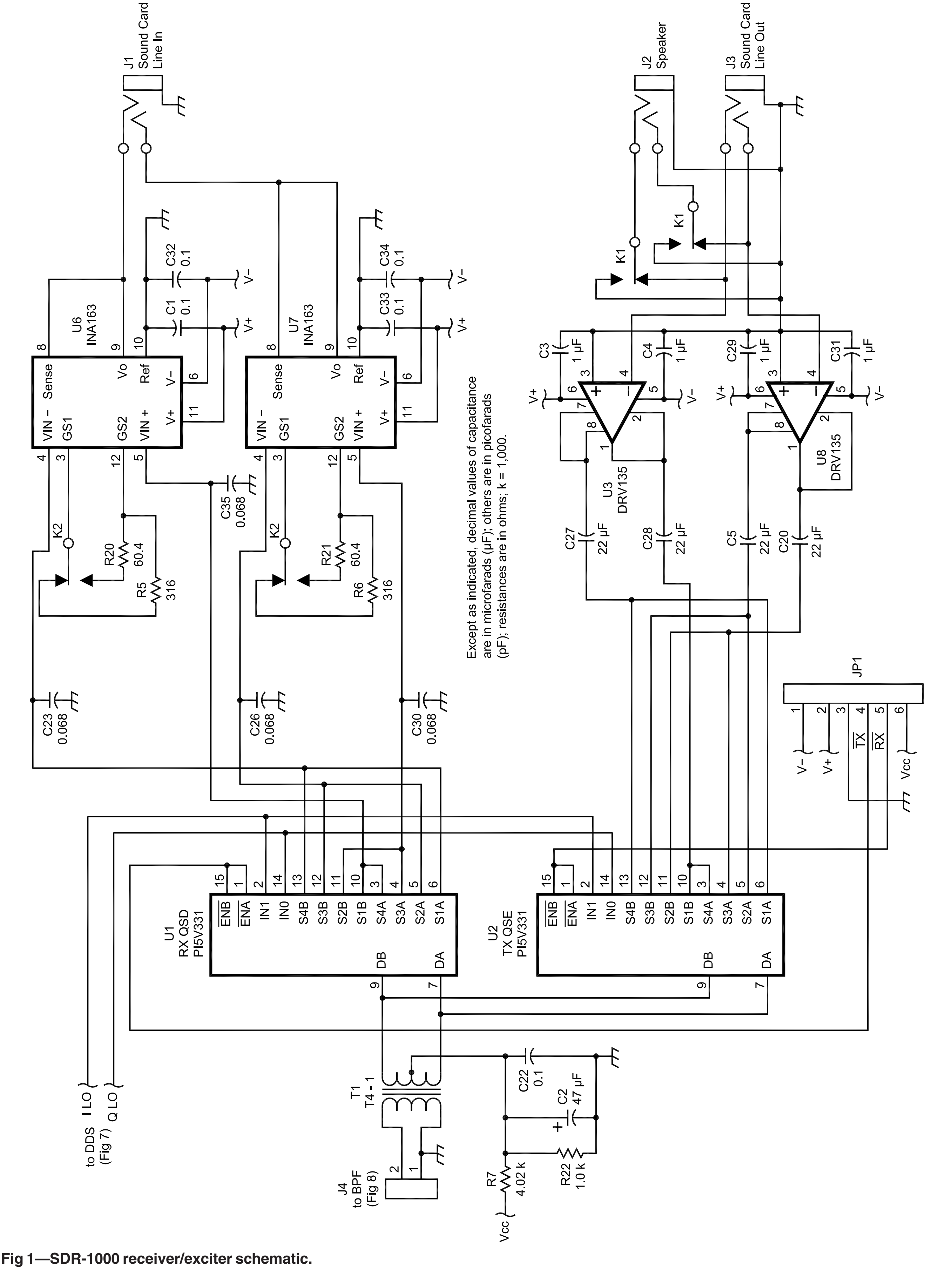

Fig 1—SDR-1000 receiver/exciter schematic.

reflectors claiming measured *IP3* in the +40 dBm range for QSD detectors using 5-V parts. With ultra-low-noise audio op amps, it is possible to achieve an analog noise figure on the order of 1 dB without an RF preamplifier. With appropriately designed analog AGC and careful gain distribution, it is theoretically possible to achieve over 150 dB of total dynamic range. The question is whether that much range is needed for typical HF applications. In reality, the answer is no. So how much is enough?

Several *QEX* writers have done an excellent job of addressing the subject.[20, 21, 22] Table 2 was originally published in an October 1975 *ham radio* article.[23] It provides a straightforward summary of the acceptable receiver noise figure for terrestrial communication for each band from 160 m to 2 m. Table 3 from the same article illustrates the acceptable noise figures for satellite communications on bands from 10 m to 70 cm.

For my objective of dc-60 MHz coverage in the SDR-1000, Table 2 indicates that the acceptable noise figure ranges from 45 dB on 160 m to 9 dB on 6 m. This means that a 1-dB noise figure is overkill until we operate near the 2-m band. Further, to utilize a 1-dB noise figure requires almost 70 dB of analog gain ahead of the sound card. This means that proper gain distribution and analog AGC design is critical to maximize IMD dynamic range.

After reading the referenced articles and performing measurements on the Turtle Beach Santa Cruz sound card, I determined that the complexity of an analog AGC circuit was unwarranted for my application. The Santa Cruz card has an input clipping level of 12 V (RMS, 34.6 dBm, normalized to 50 Ω) when set to a gain of –10 dB. The maximum output available from my audio signal generator is 12 V (RMS). The SDR software can easily monitor the peak signal input and set the corresponding sound card input gain to effectively create a digitally controlled analog AGC with no external hardware. I measured the sound card's 11-kHz SNR to be in the range of 96 dB to 103 dB, depending on the setting of the card's input gain control. The input control is capable of attenuating the gain by up to 60 dB from full scale. Given the large signal-handling capability of the QSD and sound card, the 1-dB compression point will be determined by the output saturation level of the instrumentation amplifier.

Of note is the fact that DVD sales are driving improvements in PC sound cards. The newest 24-bit sound cards sample at a rate of up to 192 kHz. The Waveterminal 192X from EGO SYS is one example.[24] The manufacturer boasts of a 123 dB dynamic range, but that number should be viewed with caution because of the technical difficulties of achieving that many bits of true resolution. With a 192-kHz sampling rate, it is possible to achieve real-time reception of 192 kHz of spectrum (assuming quadrature sampling).

Quadrature Sampling Detector/ Exciter Design

In Part 1 of this series (Note 3), I described the operation of a single-balanced version of the QSD. When the circuit is reversed so that a quadrature excitation signal drives the sampler, a SSB generator or exciter is created. It is a simple matter to reverse the SDR receiver software so that it transforms microphone input into filtered, quadrature output to the exciter.

While the singly-balanced circuit described in Part 1 is extremely simple, I have chosen to use the double-balanced QSD as shown in Fig 1 because of its superior common mode and even-harmonic rejection. U1, U6 and U7 form the receiver and U2, U3 and U8 form the exciter. In the receive mode, the QSD functions as a two-capacitor commutating filter, as described by Chen Ping in his article (Note 11). A commutating filter works like a comb filter, wherein the circuit responds to harmonics of the commutation frequency. As he notes, ". . . it can be shown that signals having harmonic numbers equal to any of the integer factors of the number of capacitors may pass." Since two capacitors are used in each of the I and Q channels, a two-capacitor commutating filter is formed. As Ping further states, this serves to suppress the even-order harmonic responses of the circuit. The output of a two-capacitor filter is extremely phase-sensitive, therefore allowing the circuit to perform signal detection just as a CW demodulator does. When a signal is near the filter's center frequency, the output amplitude would be modulated at the difference (beat) frequency. Unlike a typical filter, where phase sensitivity is undesirable, here we actually take advantage of that capability.

The commutator, as described in Part 1, revolves at the center frequency of the filter/detector. A signal tuned exactly to the commutating frequency will result in a zero beat. As the signal is tuned to either side of the commutation frequency, the beat note output will be proportional to the difference frequency. As the signal is tuned toward the second harmonic, the output will decrease until a null occurs at the harmonic frequency. As the signal is tuned

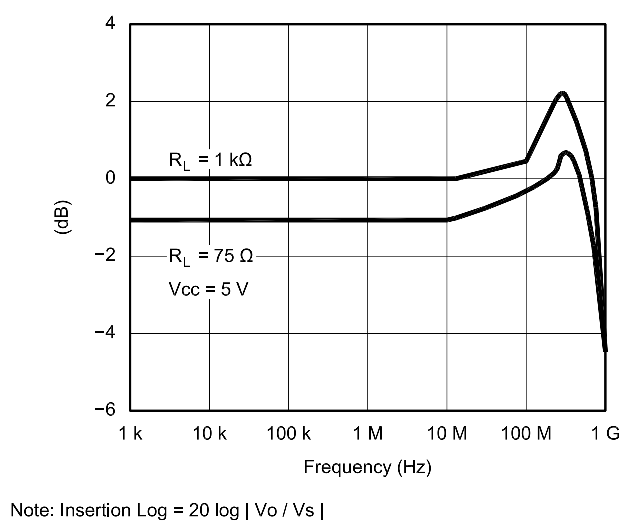

Fig 2—QS4A210 insertion loss versus frequency

Table 3—Acceptable Noise Figure for Satellite Communications

Frequency (MHz)	*Galactic Noise Floor (dBm/Hz)*	*Acceptable NF (dB)*
28	–125	8
50	–130	5
144	–139	1
220	–140	0.7
432	–141	0.2

further, it will rise to a peak at the third harmonic and then decrease to another null at the fourth harmonic. This cycle will repeat indefinitely with an amplitude output corresponding to the $\sin(x)/x$ curve that is characteristic of sampling systems as discussed in DSP texts. The output will be further attenuated by the frequency-response characteristics of the device used for the commutating switch. The PI5V331 multiplexer has a 3-dB bandwidth of 150 MHz. Other parts are available with 3-dB bandwidths of up to 1.4 GHz (from IDT Semiconductor).

Fig 2 shows the insertion loss versus frequency for the QS4A210. The upper frequency limitation is determined by the switching speed of the part (1 ns = T_{on} / T_{off}, best-case or 12.5 ns worst-case for the 1.4-GHz part) and the $\sin(x)/x$ curve for under-sampling applications.

The PI5V331 (functionally equivalent to the IDT QS4A210) is rated for analog operation from 0 to 2 V. The QS4A210 data sheet provides a drain-to-source on-resistance curve versus the input voltage as shown in Fig 3. From the curve, notice that the on resistance (R_{on}) is linear from 0 to 1 V and increases by less than 2 Ω at 2 V. No curve is provided in the PI5V331 data sheet, but we should be able to assume the two are comparable. In fact, the PI5V331 has a R_{on} specification of 3 Ω (typical) versus the 5 Ω (typical) for the QS41210. In the receive application of the QSD, the R_{on} is looking into the 60-MΩ input of the instrumentation amplifier. This means that ΔR_{on} modulation is virtually nonexistent and will have no material effect on circuit linearity.[25] Unlike typical mixers, which are nonlinear, the QSD is a linear detector!

Eq 1 determines the bandwidth of the QSD, where R_{ant} is the antenna impedance, C_S is the sampling capacitor value and n is the total number of sampling capacitors ($1/n$ is effectively the switch duty cycle on each capacitor). In the doubly balanced QSD, n is equal to 2 instead of 4 as in the singly balanced circuit. This is because the capacitor is selected twice during each commutation cycle in the doubly balanced version.

$$BW_{det} = \frac{1}{\pi \, n \, R_{ant} C_S} \qquad \text{(Eq 1)}$$

A tradeoff exists in the choice of QSD bandwidth. A narrow bandwidth such as 6 kHz provides increased blocking and IMD dynamic range because of the very high Q of the circuit. When designed for a 6-kHz bandwidth, the response at 30 kHz—one decade from the 3-kHz 3-dB point—either side of the center frequency will be attenuated by 20 dB. In this case, the QSD forms a 6-kHz-wide tracking filter centered at the commutating frequency. This means that strong signals outside the passband of the QSD will be attenuated, thereby dramatically increasing *IP3* and blocking dynamic range.

I am interested in wider bandwidth for several reasons and therefore willing to trade off some of the IMD-reduction potential of the QSD filter. In SDR applications, it is desirable in many cases to receive the widest bandwidth of which the sound card is capable. In my original design, that is 44 kHz with quadrature sampling. This capability increases to 192 kHz with the newest sound cards. Not only does this allow the capability of observing the real-time spectrum of up to 192 kHz, but it also brings the potential for sophisticated noise and interference reduction.[26]

Further, as we will see in a moment, the wider bandwidth allows us to reduce the analog gain for a given sensitivity level. The 0.068-µF sampling capacitors are selected to provide a QSD bandwidth of 22 kHz with a 50-Ω antenna. Notice that any variance in the antenna impedance will result in a corresponding change in the bandwidth of the detector. The only way to avoid this is to put a buffer in front of the detector.

The receiver circuit shown in Part 1 used a differential summing op amp after the detector. The primary advantage of a low-noise op amp is that it can provide a lower noise figure at low gain settings. Its disadvantage is that the inverting input of the op amp will be at virtual ground and the non-inverting input will be high impedance. This means that the sampling capacitor on the inverting input will be loaded differently from the non-inverting input. Thus, the respective passbands of the two inputs will not track one another. This problem is eliminated if an instrumentation amplifier is used. Another advantage of using an instrumentation amplifier as opposed to an op amp is that the antenna impedance is removed from the amplifier gain equation. The single disadvantage of the instrumentation amplifier is that the voltage noise

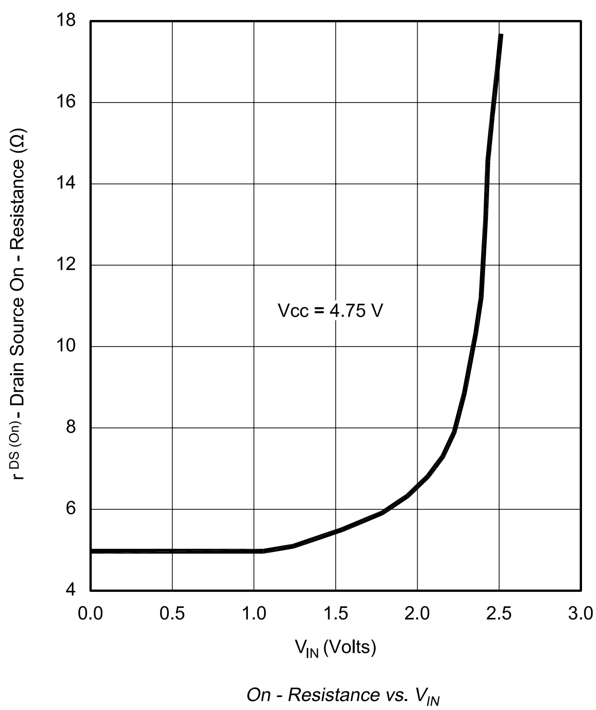

Fig 3—QS4A210 R_{on} versus V_{IN}.

Table 4—INA 163 Noise Data at 10 kHz

Gain (dB)	e_n	i_n	*NF (dB)*
20	7.5 nV/√ Hz	0.8 pA/√ Hz	12.4
40	1.8 nV/√ Hz	0.8 pA/√ Hz	3.0
60	1.0 nV/√ Hz	0.8 pA/√ Hz	1.3

and thus the noise figure increases with decreasing gain.

Table 4 shows the voltage noise, current noise and noise figure for a 200-Ω source impedance for the TI INA163 instrumentation amplifier. Since a single resistor sets the gain of each amplifier, it is a simple matter to provide two or more gain settings with relay or solid-state switching.

Unlike typical mixers, which are normally terminated in their characteristic impedances, the QSD is a high-impedance, sampling device. Within the passband, the QSD outputs are terminated in the 60-MΩ inputs of the instrumentation amplifiers. The IDT data sheet for the QS4A210 indicates that the switch has *no insertion loss* with loads of 1 kΩ or more! This coincides with my measurements on the circuit. If you apply 1 V of RF into the detector, you get 1 V of audio out on each of the four capacitors—a no-loss detector. Outside the passband, the decreasing reactance of the sampling capacitors will reduce the signal level on the amplifier inputs. While it is possible to insert series resistors on the output of the QSD, so that it is terminated outside the passband, I believe this is unnecessary. For receive operation, filter reflections outside the passband are not very important. Further, the termination resistors would create an additional source of thermal noise.

As stated earlier, the circuitry of the QSD may be reversed to form a quadrature sampling exciter (QSE). To do so, we must differentially drive the I and Q inputs of the QSE. The Texas Instruments DRV135 50-Ω differential audio line driver is ideally suited for the task. Blocking capacitors on the driver outputs prevent dc-offset variation between the phases from creating a carrier on the QSE output. Carrier suppression has been measured to be on the order of –48 dBc relative to the exciter's maximum output of +10 dBm. In transmit mode, the output impedance of the exciter is 50 Ω so that the band-pass filters are properly terminated.

Conveniently, T/R switching is a simple matter since the QSD and QSE can have their inputs connected in parallel to share the same transformer. Logic control of the respective multiplexer-enable lines allows switching between transmit and receive mode.

Level Analysis

The next step in the design process is to perform a system-level analysis of the gain required to drive the sound card A/D converter. One of the better references I have found on the subject is the book by W. Sabin and E. Schoenike, *HF Radio Systems and Circuits.*[27] The book includes an *Excel* spreadsheet that allows interactive examination of receiver performance using various A/D converters, sample rates, bandwidths and gain distributions. I have placed a copy of the SDR-1000 Level Analysis spreadsheet (by permission, a highly modified version of the one provided in the book) for download from ARRLWeb.[28] Another excellent resource on the subject is the Digital Receiver/Exciter Design chapter from the book *Digital Signal Processing in Communication Systems.*[29]

Notice that the former reference has a better discussion of the minimum gain required for thermal noise to transition the quantizing level as discussed here. Neither text deals with the effects of atmospheric noise on the noise floor and hence on dynamic range. This is—in my opinion—a major oversight for HF communications since atmospheric noise will most likely limit the minimum discernable signal, not thermal noise.

For a weak signal to be recovered, the minimum analog gain must be great enough so that the weakest signal to be received, plus thermal and atmospheric noise, is greater than at least one A/D converter quantizing level (the least-significant usable bit). For the A/D converter quantizing noise to be evenly distributed, several quantizing levels must be traversed. There are two primary ways to achieve this: Out-of-band dither noise may be added and then filtered out in the DSP routines, or in-band thermal and atmospheric noise may be amplified to a level that accomplishes the same. While the first approach offers the best sensitivity at the lowest gain, the second approach is simpler and was chosen for my application. *HF Radio Systems and Circuits* states, "Normally, if the noise is Gaussian distributed, and the RMS level of the noise at the A/D converter is greater than or equal to the level of a sine wave which just bridges a single quantizing level, an adequate number of quantizing levels will be bridged to guarantee uniformly distributed quantizing noise." Assuming uniform noise distribution, Eq 2 is used to determine the quantizing noise density, N_{0q}:

$$N_{0q} = \frac{\left(\frac{V_{pp}}{2^b}\right)^2}{6 f_s R}\ \text{W/Hz} \qquad \text{(Eq 2)}$$

where
$V_{P\text{-}P}$= peak-to-peak voltage range
b = number of valid bits of resolution
f_s = A/D converter sampling rate
R = input resistance
N_{0q} = quantizing noise density

The quantizing noise decreases by 3 dB when doubling the sampling rate and by 6 dB for every additional bit of resolution added to the A/D converter. Notice that just because a converter is specified to have a certain number of bits does not mean that they are all usable bits. For example, a converter may be specified to have 16 bits; but in reality, only be usable to 14-bits. The Santa Cruz card utilizes an 18-bit A/D converter to deliver 16 usable bits of resolution. The maximum signal-to-noise ratio may be determined from Eq 3:

$$SNR = 6.02b + 1.75\ \text{dB} \qquad \text{(Eq 3)}$$

For a 16-bit A/D converter having a maximum signal level (without input attenuation) of 12.8 $V_{P\text{-}P}$, the minimum quantum level is –70.2 dBm. Once the quantizing level is known, we can compute the minimum gain required from Eq 4:

$$Gain = quantizing\ level - kTB + analog\ NF + atmospheric\ NF - 10\log_{10} BW \qquad \text{(Eq 4)}$$

where:

$$quantizing\ level = 10\log_{10}\left(\frac{\left(\frac{V_{pp}\times 0.707}{2^b \times 2}\right)^2}{50\times 0.001}\right)\text{dBm}$$

kTB / Hz = –174 dBm/Hz
analog NF = analog receiver noise figure, in decibels
atmospheric NF = atmospheric noise figure for a given frequency
BW = the final receive filter bandwidth in hertz

Table 5, from the SDR-1000 Level Analysis spreadsheet, provides the cascaded noise figure and gain for the circuit shown if Fig 1. This is where things get interesting.

Fig 4 shows an equivalent circuit for the QSD and instrumentation amplifier during a respective switch period. The transformer was selected to have a 1:4 impedance ratio. This means that the turns ratio from the primary to the secondary for each switch to ground is 1:1, and therefore the voltage on each switch is equal to the input signal voltage. The differential impedance across the transformer secondary will be 200 Ω, providing a good noise match to the INA163 amplifier. Since the input impedance of the INA163 is 60 MΩ, power loss through the circuit is virtually nonexistent. We must therefore analyze the circuit based on voltage gain, not power gain.

Table 5 –Cascaded Noise Figure and Gain Analysis from the SDR-1000 Level Analysis Spreadsheet

		BPF	*T1-4*	*PI5V331*	*INA163*	*ADC*
dB	Noise Figure	0.0	0.0	0.0	3.0	58.6
dB	Gain	0.0	6.0	0.0	40.0	0.0
Equivalent Power Factor	Noise Factor	1.00	1.00	1.00	1.99	720,482
Equivalent Power Factor	Gain	1	4	1	10,000	1
Clipping Level	Vpk			1.0	13.0	6.4
Clipping Level	dBm			10.0	32.3	26.1
Cascaded Gain	dB	0.0	6.0	6.0	46.0	46.0
Cascaded Noise Factor		1.00	1.00	1.00	1.25	19.06
Cascaded Noise Figure	dB	0.0	0.0	0.0	1.0	12.8
Output Noise	dBm/Hz	–174.0	–174.0	–174.0	–173.0	–161.2

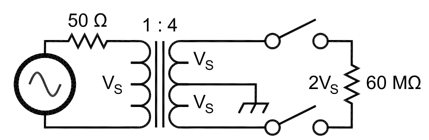

Fig 4—Doubly balanced QSD equivalent circuit.

Table 6—Atmospheric Equivalent Noise Figure By Band

Band (Meters)	*Ext Noise (dBm/Hz)*	*Ext NF (dB)*
160	–128	46
80	–136	38
40	–144	30
30	–146	28
20	–146	28
17	–152	22
15	–152	22
12	–154	20
10	–156	18
6	–162	12

Table 7—SDR-1000 Level Analysis Assumptions for the 10-Meter Band with 40 dB of INA Gain

Receiver Gain Distribution and Noise Performance	
Turtle Beach Santa Cruz Audio Card	
Band Number	9
Band	10 Meters
Include External NF? (True=1, False=0)	1
External (Atmospheric) Noise Figure	18 dB
A/D Converter Resolution (bits)	16 bits (98.1 dB)
A/D Converter Full–Scale Voltage	6.4 V-peak (26.1 dBm)
A/D Converter Quantizing Signal Level	–70.2 dBm
Quantizing Gain Over/(Under)	7.2 dB
A/D Converter Sample Frequency	44.1 kHz
A/D Converter Input Bandwidth (BW1)	40.0 kHz
Information Bandwidth (BW2)	0.5 kHz
Signal at Antenna for INA Saturation	–13.7 dBm
Nominal DAC Output Level	0.5 V peak (4.0 dBm)
AGC Threshold at Ant (40 dB Headroom)	–51.4 dBm
Sound Card AGC Range	60.0 dB

That means that we get a 6-dB differential voltage gain from the input transformer—the equivalent of a 0-dB noise figure amplifier! Further, there is no loss through the QSD switches due to the high-impedance load of the INA. With a source impedance of 200 Ω, the INA163 has a noise figure of approximately 12.4 dB at 20 dB of gain, 3 dB at 40 dB of gain and 1.3 dB at 60 dB of gain.

In fact, the noise figure of the analog front end is so low that if it were not for the atmospheric noise on the HF bands, we would need to add a lot of gain to amplify the thermal noise to the quantizing level. The textbook references ignore this fact. In addition to the *ham radio* article (Note 23) and Peter Chadwick's *QEX* article (Note 20), John Stephenson in his *QEX* article[30] about the ATR-2000 HF transceiver provides further insight into the subject. Table 6 provides a summary of the external noise figure for a by-band quiet location as determined from Fig 1 in Stephenson's article. As can be seen from the table, it is counterproductive to have high gain and low receiver noise figure on most of the HF bands.

Tables 7 and 8 are derived from the *SDR-1000 Level Analysis* spreadsheet (Note 28) for the 10-m band. The spreadsheet tables interact with one another so that a change in an assumption will flow through all the other tables. A detailed discussion of the spreadsheet is beyond the scope of this text. The best way to learn how to use the spreadsheet is to plug in values of your own. It is also instructive to highlight cells of interest to see how the formulas are derived. Based on analysis using the spreadsheet, I have chosen to make the gain setting relay-selectable between INA gain settings of 20 dB for the lower bands and 40 dB for the higher bands.

It is important to remember that my noise and dynamic-range calculations include external noise figure in addition to the thermal noise figure. This is much more realistic for HF applications than the typical lab testing and calculations you see in most references. With the INA163 gain set to 40 dB, the cascaded analog thermal NF is calculated to be just 1 dB at the input to the sound card. If it were not for the external noise, nearly 70 dB of analog gain would be required to amplify the thermal noise

Table 8—SDR-1000 Level Analysis Detail for the 10-Meter Band with 40dB of INA Gain

Antenna Signal Level (dBm)	*INA Output Level (dBm)*	*Antenna Overload Level (dBm)*	*Total Analog Gain (dB)*	*Sound Card AGC Reduction (dB)*	*Noise at A/D Input in BW1 (dBm)*	*A/D Signal Level (dBm)*	*Noise in A/D Input in BW2 (dBm)*	*Quantizing Noise of A/D in BW2 (dBm)*	*Total Noise in BW2 (dBm)*	*Output S/N Ratio in BW2 (dB)*	*Digital Gain Required (dB)*
–128	–82		46.0	0.0	–63.0	–82.0	–82.0	–88.4	–81.1	–0.9	86.0
–118	–72		46.0	0.0	–63.0	–72.0	–82.0	–88.4	–81.1	9.1	76.0
–108	–62		46.0	0.0	–63.0	–62.0	–82.0	–88.4	–81.1	19.1	66.0
–98	–52		46.0	0.0	–63.0	–52.0	–82.0	–88.4	–81.1	29.1	56.0
–88	–42		46.0	0.0	–63.0	–42.0	–82.0	–88.4	–81.1	39.1	46.0
–78	–32		46.0	0.0	–63.0	–32.0	–82.0	–88.4	–81.1	49.1	36.0
–68	–22		46.0	0.0	–63.0	–22.0	–82.0	–88.4	–81.1	59.1	26.0
–58	–12		46.0	0.0	–63.0	–12.0	–82.0	–88.4	–81.1	69.1	16.0
–48	–2		42.7	–3.3	–66.4	–5.4	–85.4	–88.4	–83.6	78.2	9.4
–38	8		32.7	–13.3	–76.4	–5.4	–95.4	–88.4	–87.6	82.2	9.4
–28	18		22.7	–23.3	–86.4	–5.4	–105.4	–88.4	–88.3	82.9	9.4
–18	28		12.7	–33.3	–96.4	–5.4	–115.4	–88.4	–88.4	83.0	9.4
–8	38	6	2.7	–43.3	–106.4	–5.4	–125.4	–88.4	–88.4	83.0	9.4
2	48	16	–7.3	–53.3	–116.4	–5.4	–135.4	–88.4	–88.4	83.0	9.4
12	58	26	–14.0	–60.0	–123.0	–2.0	–142.0	–88.4	–88.4	86.4	6.0
22	68	36	–14.0	–60.0	–123.0	8.0	–142.0	–88.4	–88.4	96.4	–4.0
32	78	46	–14.0	–60.0	–123.0	18.0	–142.0	–88.4	–88.4	106.4	–14.0

to the quantizing level or dither noise would have to be added outside the passband. Fig 6 illustrates the signal-to-noise ratio curve with external noise for the 10-m band and 40 dB of INA gain. Fig 5 shows the same curve without external noise and with INA gain of 60 dB. This much gain would not improve the sensitivity in the presence of external noise but would reduce blocking and IMD dynamic range by 20 dB. On the lower bands, 20 dB or lower INA gain is perfectly acceptable given the higher external noise.

Frequency Control

Fig 7 illustrates the Analog Devices AD9854 quadrature DDS circuitry for driving the QSD/QSE. Quadrature local-oscillator signals allow the elimination of the divide-by-four Johnson counter, described in Part 1, so that the DDS runs at the carrier frequency instead of its fourth harmonic. I have chosen to use the 200-MHz version of the part to minimize heat dissipation, and because it easily meets my frequency coverage requirements of dc-60 MHz. The DDS outputs are connected to seventh-order elliptical low-pass filters that also provide a dc reference for the high-speed comparators. The AD9854 may be controlled either through a SPI port or a parallel interface. There are timing issues in SPI mode that require special care in programming. Analog Devices have developed a protocol that allows the chip to be put into external I/O update mode to work around the serial

About Intel Performance Primitives

Many readers have inquired about Intel's replacement of its Signal Processing Library (SPL) with the Intel Performance Primatives (IPP). The SPL was a free distribution, but the Intel Web site states that IPP requires payment of a $199 fee after a 30 day evaluation period. A fully functional trial version of IPP may be downloaded from the Intel site at **www.intel.com/software/products/global/eval.htm**. The author has confirmed with Intel Product Management that no license fee is required for amateur experimentation using IPP, and there is no limit on the evaluation period for such use. Intel actually encourages this type of experimental use. Payment of the license fee is required if and only if there is a commercial distribution of the DLL code.—*Gerald Youngblood* timing problem. In the final circuit, I chose to use the parallel mode.

According to Peter Chadwick's article (Note 20), phase-noise dynamic range is often the limiting factor in receivers instead of IMD dynamic range. The AD9854 has a residual phase noise of better than –140 dBc/Hz at a 10-kHz offset when directly clocked at 300 MHz and programmed for an 80-MHz output. A very low-jitter clock oscillator is required so that the residual phase noise is not degraded significantly.

High-speed data communications technology is fortunately driving the introduction of high-frequency crystal oscillators with very low jitter specifications. For example, Valpey Fisher makes oscillators specified at less than 1 ps RMS jitter that operate in the desired 200-300 MHz range. According to Analog Devices, 1 ps is on the order of the residual jitter of the AD9854.

Band-Pass Filters

Theoretically, the QSD will work just fine with low-pass rather than band-pass filters. It responds to the carrier frequency and odd harmonics of the carrier; however, very large signals at half the carrier frequency can be heard in the output. For example, my measurements show that when the receiver is tuned to 7.0 MHz, a signal at 3.5 MHz is attenuated by 49 dB. The measurements show that the attenuation of the second harmonic is 37 dB and the third harmonic is down 9 dB from the 7-MHz reference. While a simple low-pass filter will suffice in some applications, I chose to use band-pass filters.

Fig 8 shows the six-band filter design for the SDR-1000. Notice that only the 2.5-MHz filter has a low-pass characteristic; the rest are band-pass filters.

SDR-1000 Board Layout

For the final PC-board layout, I decided on a 3×4-inch form factor. The receiver, exciter and DDS are located on one board. The band-pass filter and a 1-W driver amplifier are located on a second board. The third board has a PC parallel-port interface for control, and power regulators for operation from a 13.8-V dc power source. The three boards sandwich together into a small 3×4×2-inch module with rear-mount connectors and no interconnection wiring required. The boards use primarily surface-mount components, except for the band-pass filter, which uses mostly through-hole components.

Acknowledgments

I would like to thank David Brandon and Pascal Nelson of Analog Devices for their answering my questions about the AD9854 DDS. My appreciation also goes to Mike Pendley, WA5VTV, for his assistance in design of the band-pass filters as well as his ongoing advice.

Conclusion

This series has presented a practical approach to high-performance SDR development that is intended to spur broad-scale amateur experimentation. It is my hope—and that of the ARRL SDR Working Group—that many will be encouraged to contribute to the technical art in this fascinating area. By making the SDR-1000 hardware and software available to the amateur community, software extensions may be easily and quickly added. Thanks for reading.

Notes

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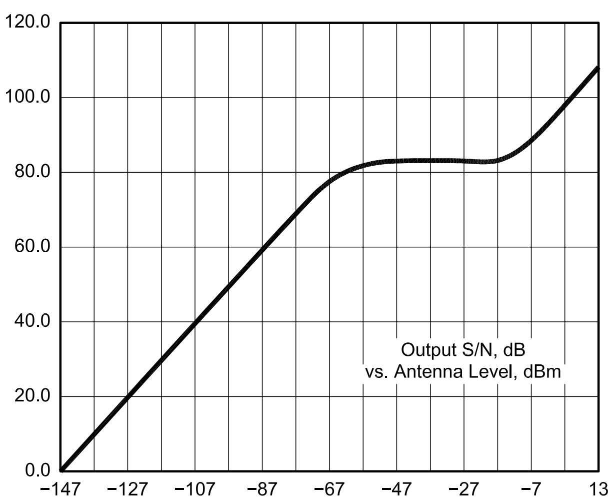

Fig 5—Output signal-to-noise ratio excluding external (atmospheric) noise. INA gain is set to 60 dB. Antenna signal level for saturation is –33.7 dBm.

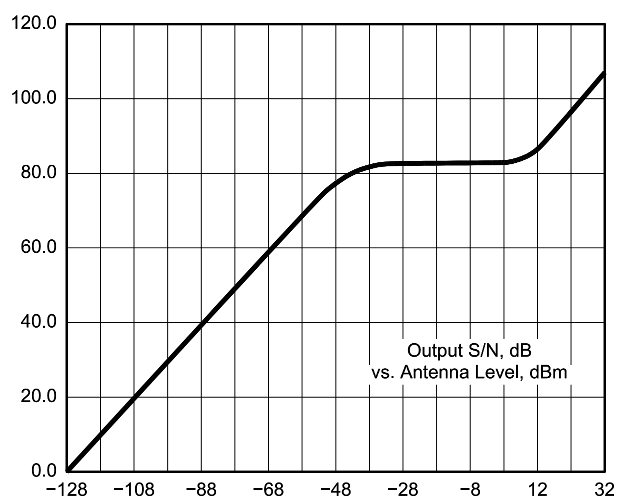

Fig 6—Output signal-to-noise ratio for the 10-m band including external (atmospheric) noise. INA gain is set to 40 dB. Antenna signal level for INA saturation is –13.7dBm.

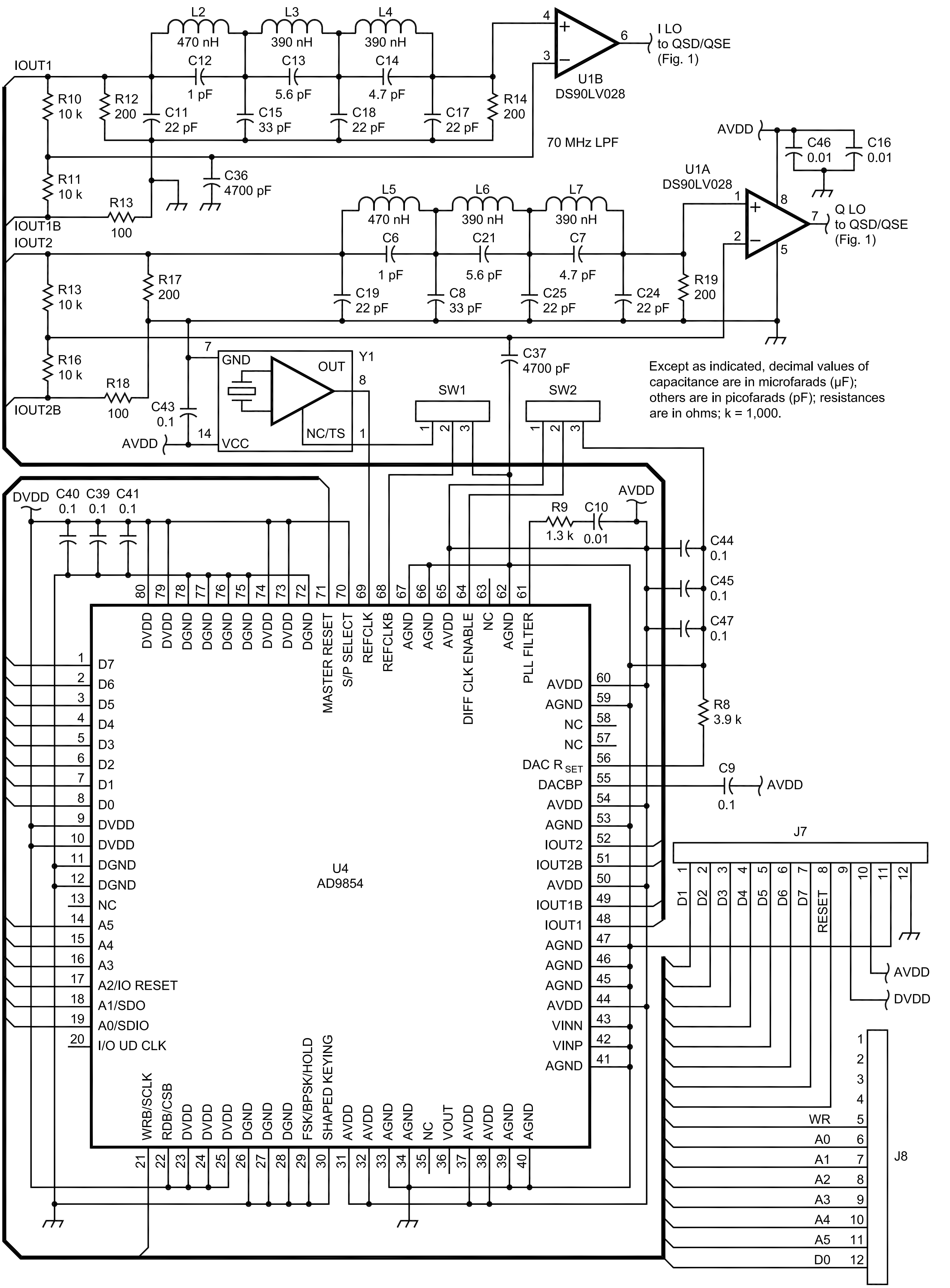

Fig 7—SDR-1000 quadrature DDS schematic.

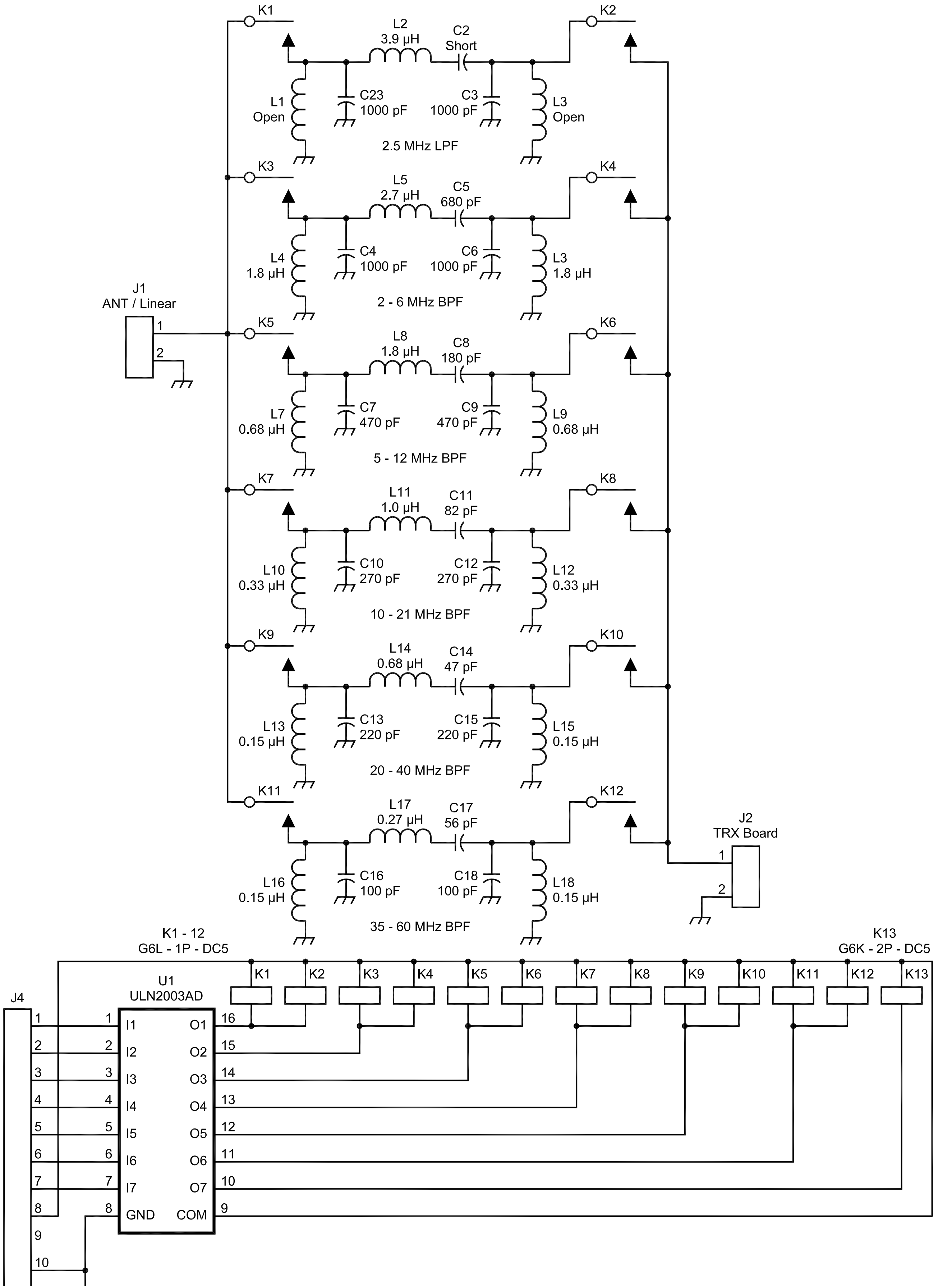

Fig 8—SDR-1000 six-band filter schematic.

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Put Your Antenna Modeling Programs on Autopilot

Run a simulated sweep on your antenna model, but automatically change something other than just frequency.

By Dan Maguire, AC6LA

In October of 2001, I was asked to be a part of the team doing the beta test on the ARRL Antenna Modeling course. That course discusses a number of interesting concepts and techniques for building antenna models. Among these is a method known as modeling by equation, wherein the wire geometry of a model may be modified merely by changing the values for one or more variables as opposed to changing all the wire XYZ coordinates by hand.

Of the two popular commercial implementations of *NEC-2* that are described in the course, *NEC-Win Plus+*[1] by Nittany Scientific and *EZNEC*[2] by Roy Lewallen, W7EL, only the former offers built-in support for the modeling-by-equation technique. I wanted to use *EZNEC* in a similar manner, so I created a pair of Excel spreadsheets to let me do just that. The XYZ coordinates were entered as *Excel* formulas on one sheet. The other sheet controlled the values for the variables that were used in the formulas on the first sheet. Then an ASCII file was built containing the set of wire coordinates as generated per the *Excel* formulas and variable values. The ASCII file had to be imported manually into *EZNEC*, replacing the wire definitions of an existing *EZNEC* model.

Other aspects of the antenna, such as the source type and placement, the ground type and characteristics, and the desired plot type and angle were all defined as part of the existing model and were controlled using normal *EZNEC* methods. The model was run in *EZNEC*, the results were inspected, and then it was back to the *Excel* spreadsheets to perhaps create a new version of the wires. A lot of manual intervention was required.

I decided to automate the process steps of changing variable values, sending the model to *EZNEC*, and extracting information from the calculated output. But I didn't stop there. In the course of program development, several additional capabilities were added, including functions to:

• Use variables to control features of the model besides wire geometry, such as source and load parameters;

• Build complete model files, not just wire sets, in *EZNEC* binary format, *NEC* "deck" format, and in *Antenna Model*[3] (Teri Software) format;

• Import existing *EZNEC*, *NEC* and Antenna-Model-format files, as well as several other formats, which can then be modified if desired to have certain aspects controlled via variables;

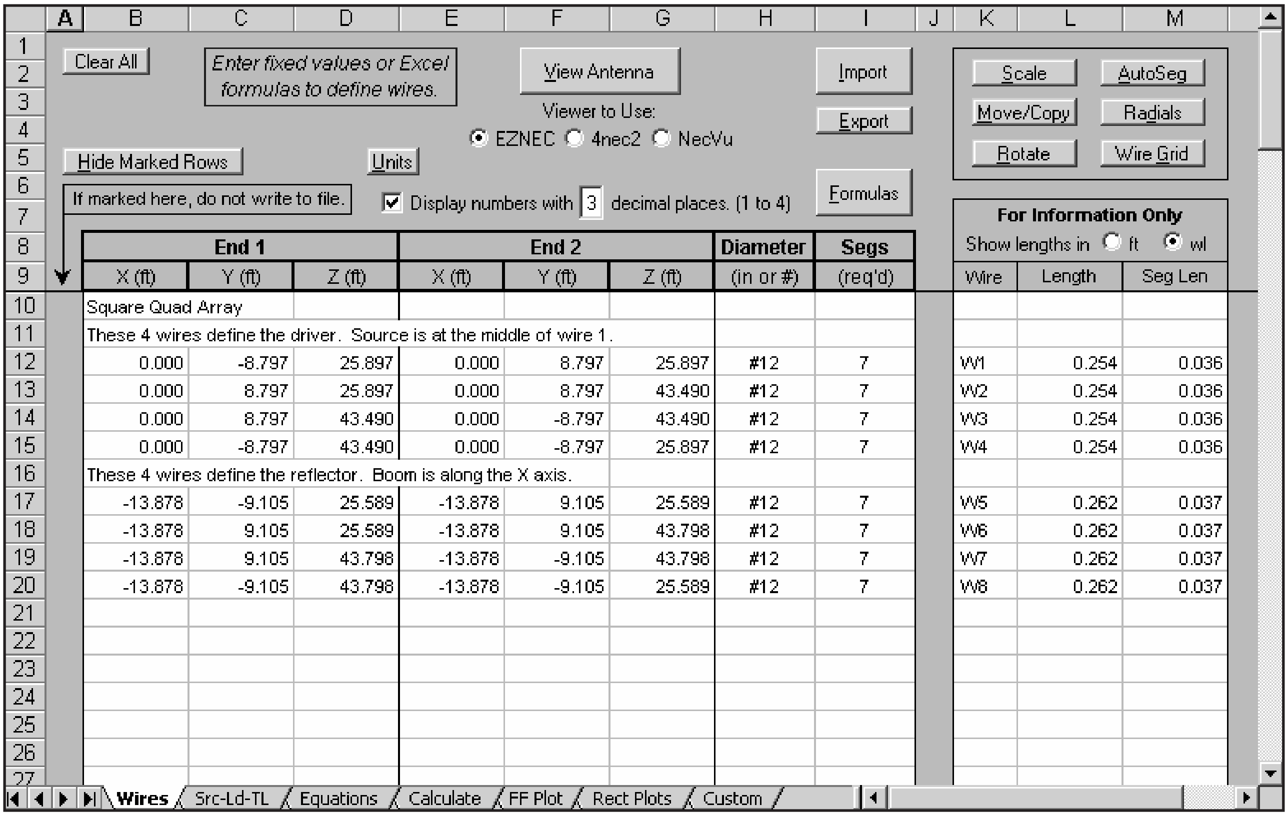
• Use calculating engines from several commercial vendors and a *NEC-2* public domain engine;

• Support user-friendly *EZNEC* conventions such as split sources and "trap"-type loads no matter what calculating engine is used;

• Allow easy comparison of

[1]Notes appear on page 39.

2150 Louis Holstrom Dr
Morgan Hill, CA 95037
ac6la@arrl.net



Clear All

Enter fixed values or Excel formulas to define wires.

View Antenna

Viewer to Use: EZNEC 4nec2 NecVu

Import

Export

Scale AutoSeg Move/Copy Radials Rotate Wire Grid

Hide Marked Rows

Units

If marked here, do not write to file.

Display numbers with 3 decimal places. (1 to 4)

Formulas

End 1			End 2			Diameter	Segs
X (ft)	Y (ft)	Z (ft)	X (ft)	Y (ft)	Z (ft)	(in or #)	(req'd)
Square Quad Array							
These 4 wires define the driver. Source is at the middle of wire 1.							
0.000	-8.797	25.897	0.000	8.797	25.897	#12	7
0.000	8.797	25.897	0.000	8.797	43.490	#12	7
0.000	8.797	43.490	0.000	-8.797	43.490	#12	7
0.000	-8.797	43.490	0.000	-8.797	25.897	#12	7
These 4 wires define the reflector. Boom is along the X axis.							
-13.878	-9.105	25.589	-13.878	9.105	25.589	#12	7
-13.878	9.105	25.589	-13.878	9.105	43.798	#12	7
-13.878	9.105	43.798	-13.878	-9.105	43.798	#12	7
-13.878	-9.105	43.798	-13.878	-9.105	25.589	#12	7

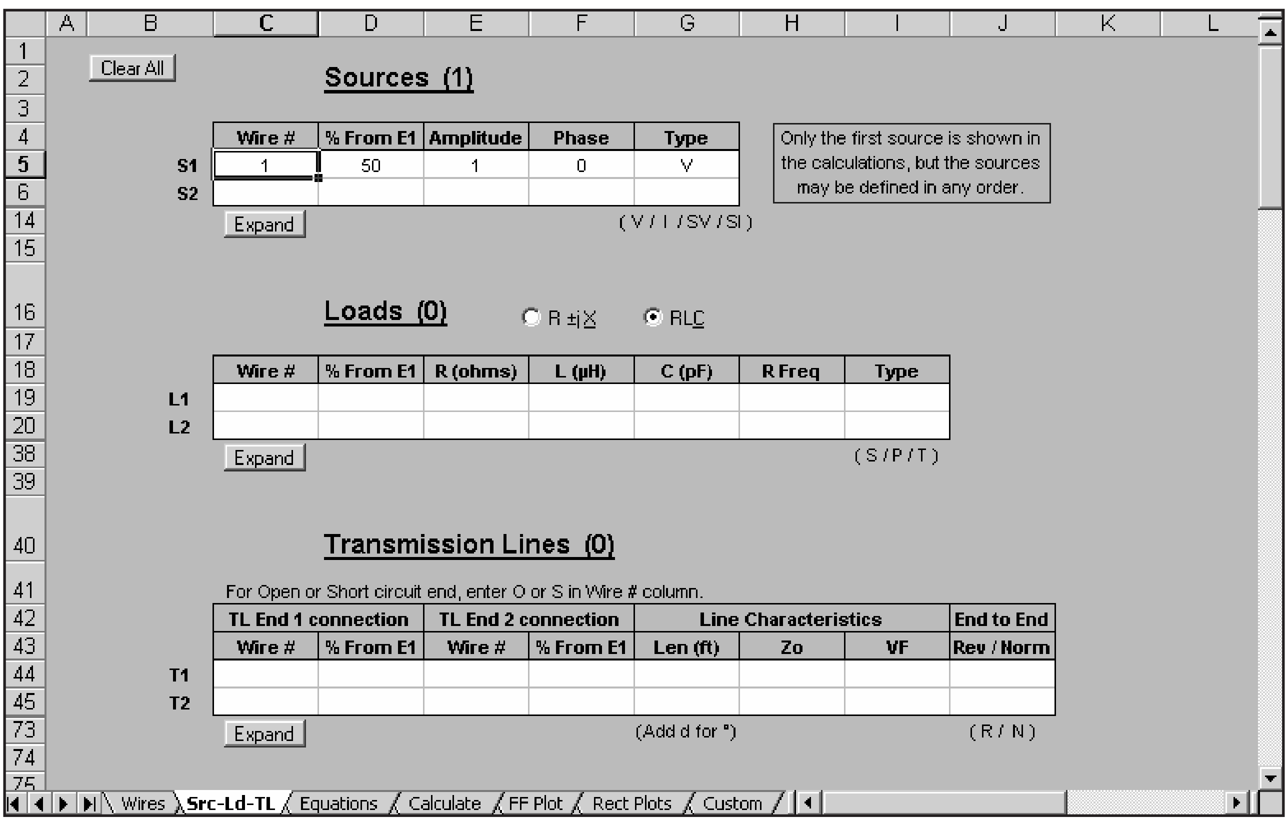
For Information Only

Show lengths in ft wl

Wire	Length	Seg Len
W1	0.254	0.036
W2	0.254	0.036
W3	0.254	0.036
W4	0.254	0.036
W5	0.262	0.037
W6	0.262	0.037
W7	0.262	0.037
W8	0.262	0.037

Wires / Src-Ld-TL / Equations / Calculate / FF Plot / Rect Plots / Custom

Fig 1—The *Wires* sheet. This view shows the numeric values for the End 1 and End 2 *XYZ* coordinates. It is also possible to show the *Excel* formulas for the coordinates, applicable for models built using the "Modeling by Equation" technique.



Clear All

Sources (1)

	Wire #	% From E1	Amplitude	Phase	Type
S1	1	50	1	0	V
S2					

Only the first source is shown in the calculations, but the sources may be defined in any order.

Expand (V / I / SV / SI)

Loads (0) R ±jX RLC

	Wire #	% From E1	R (ohms)	L (µH)	C (pF)	R Freq	Type
L1							
L2							

Expand (S / P / T)

Transmission Lines (0)

For Open or Short circuit end, enter O or S in Wire # column.

	TL End 1 connection		TL End 2 connection		Line Characteristics			End to End
	Wire #	% From E1	Wire #	% From E1	Len (ft)	Zo	VF	Rev / Norm
T1								
T2								

Expand (Add d for °) (R / N)

Wires / Src-Ld-TL / Equations / Calculate / FF Plot / Rect Plots / Custom

Fig 2—The *Src-Ld-TL* sheet. All cell values on this sheet may be set using either constant values or via variables and formulas.

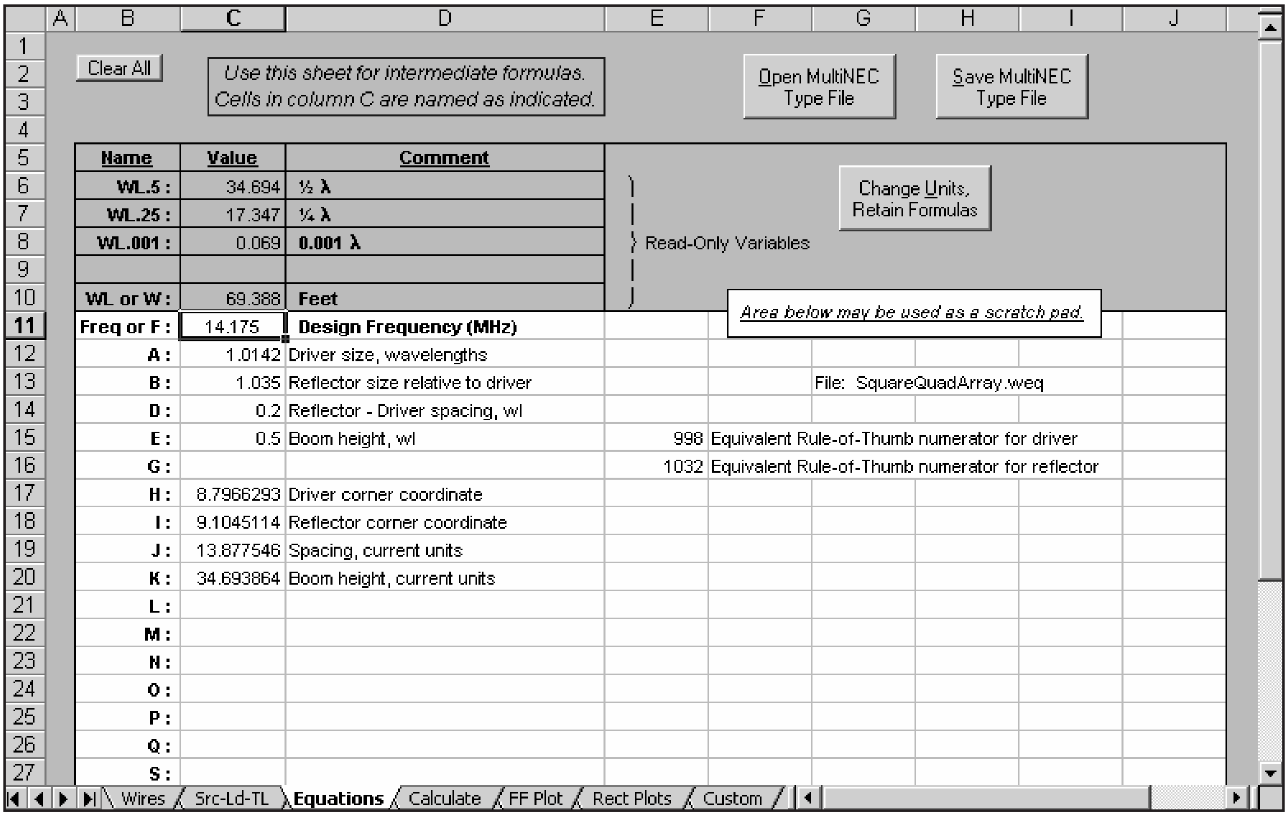


Fig 3—The *Equations* sheet, used to define the variables that may be used on the *Wires* and *Src-Ld-TL* sheets.

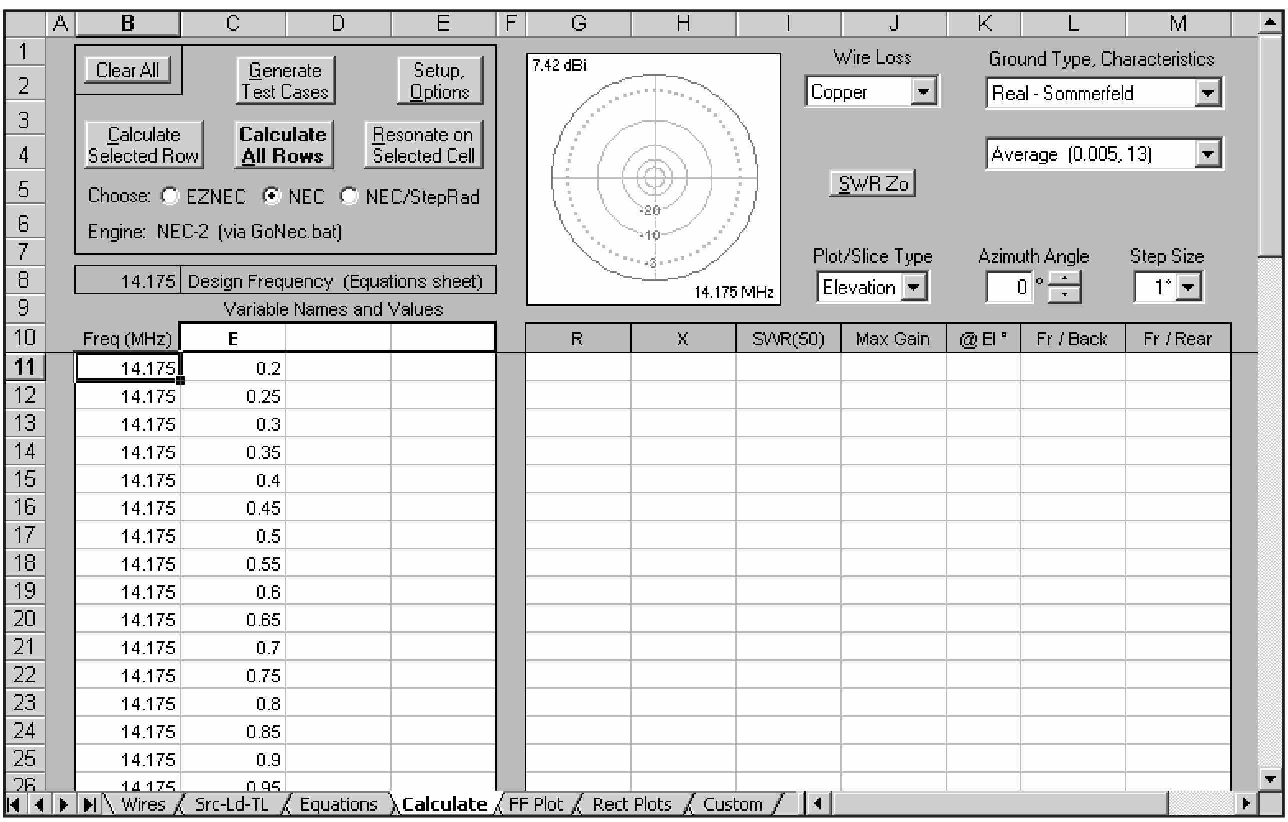


Fig 4—The left side of the *Calculate* sheet is used to define the test cases, including the specification of values for any variables that are to be changed between runs.

calculated results for the same model across multiple engines;

• Show multiple radiation patterns as an animation instead of overlaying multicolored patterns on the same plot;

• Use *Excel* built-in functions to do regression analysis on calculated data;

• Automatically find the resonant frequency of a model, or automatically change the geometry or other characteristics of a model, so that it is resonant at a desired frequency.

The result was a program called *MultiNEC*, described in this article.

MultiNEC

MultiNEC is an *Excel* application that can make multiple simulation runs of an antenna model while automatically changing one or more aspects of the model between runs. You can change the model in a variety of ways, choose from a variety of calculating engines and view the results in a variety of formats.

The model can be automatically changed between runs in several ways. The simplest is to just use a different frequency for each test case, equivalent to a normal frequency sweep. You can also specify that one or more wires in the model are to be rotated, moved, scaled or have a different segmentation level between runs. You can change the parameters of sources, loads and transmission lines, such as the position of a source along a wire or the resistance value for a load. Finally, *MultiNEC* contains a complete modeling-by-equation facility. You can create the model using *Excel* spreadsheets and then instruct the program to change the geometry or other aspects of the model in any way that can be specified using *Excel* formulas. All these changes may be made automatically without the need for any manual intervention between simulation runs.

You have several choices of the calculating engine that will be used for the simulations. If you have either standard *EZNEC* or *EZNEC Pro* you may use the *NEC-2* and *NEC-4* engines that are included in those packages. If you have *NEC-Win Plus+*, *NEC-Win Pro* or *GNEC,* you may use the engines from those as well. If you have *Antenna Model* you may use that. You may also use the public domain *NEC-2* engine that is bundled with *MultiNEC*.

The simulation results are available in tabular format or in several different plot formats, including a far-field polar plot, three standard rectangular plots and a custom rectangular plot on which you may choose any desired X and Y axes. The polar plot can be used to play back the radiation patterns for the test cases as animation. The custom plot can show a regression analysis curve and matching equation for the plotted points.

In addition to the "multi-run" capability, *MultiNEC* is also "multilingual." You can import existing antenna models that are in *EZNEC* binary format (.ez), *NEC* "deck" format (typically .nec or .inp), *Antenna Model* format (.def) or several other formats. Once the model has been imported, you may run it on any of the available engines. For example, you can import an *EZNEC* binary format model and run it using public domain *NEC-2*. Or you could import a *NEC* file and then use *EZNEC* as the calculating engine. Since *MultiNEC* can produce output files in .ez, .nec, or .def format, you can use the program as a translator between file types.

Program Description

The use of *MultiNEC* may best be described as a series of steps:

1. Define an antenna model, either by building one from scratch or importing an existing model.

2. Specify how the model is to be changed from one simulation run to the next. This might be as simple as merely changing the test case frequency between runs, or it might involve changing the geometry or other aspects of the model with the use of *Excel* formulas and variables.

3. Calculate the results for an entire set of test cases, choosing any available calculating engine and letting *MultiNEC* automatically change the model between runs in the chosen manner.

4. Analyze the results, again for an entire set of test cases as opposed to just a single run.

These steps are explained further below.

Step 1, define the antenna to be modeled. This is done using three related *Excel* spreadsheets. The sheets are:

Wires (Fig 1): This sheet is very similar to the Wires window of *EZNEC*, except that you may use *Excel* formulas as well as numeric constants for the *XYZ* coordinates, wire diameters, and number of segments. If you prefer you may Import an existing antenna model in any of several supported formats.

Although not obvious in the figure, the *XYZ* coordinates of the model used for illustration purposes were not entered as numeric constants. Instead, they were defined in terms of *Excel* formulas, using variables with values set on another workbook sheet. For example, the *Excel* formula in cell **D12** is "= –H + K." Given the current values for variables *H* and *K* the numeric result of this formula is 25.897 when rounded to three decimal places.

Src-Ld-TL (Fig 2): This sheet is used to define the Sources, Loads and Transmission Lines used in the model. The three sections of this sheet are again very similar to the corresponding windows in *EZNEC*, and again you may use *Excel* formulas as well as numeric constants.

Equations (Fig 3): This sheet is used to define the variables that may be used in *Excel* formulas on the previous two sheets. The variables may be set equal to numeric constants (as with *F* through *E* in Fig 3) or may be defined in terms of other variables and intermediate *Excel* formulas (as with *H* through *K*). For example, the formula in cell C17 of Fig 3 is "= WL * A / 8" and the formula in cell C20 is "= WL * E." By changing the values for variables on the Equations sheet you (or *MultiNEC*) can change the geometry of the wires in your model and/or change the parameters that control the sources, loads and transmission lines.

You can build your antenna model from scratch by entering data into the cells on these three sheets, or you can use the **Import** function mentioned earlier. You can also **Open** and **Save** models in *MultiNEC* format, which preserves any *Excel* formulas you might have used on the various sheets. Several sample models in this format are included in the package.

Step 2, specify the way in which the model is to be changed for each simulation run. This is done using the columns on the *left side* of the Calculate sheet (Fig 4).

You may specify a frequency and from zero to three variables (from the Equations sheet) that are to be changed for each run. For each test case, corresponding to a row on the Calculate sheet, you enter a specific frequency value and a specific value for the variable(s) to be modified. For example, you may have defined variable *E* on the Equations sheet to control the height of the antenna above ground. You could enter different values for *E* in successive rows on the Calculate sheet, with the same frequency value for each.

Another example (not shown) would be to define a variable on the Equations sheet, say *P*, which controls the placement of a source or load as a percentage from end 1 of a given wire. Or you could define variable *L* to control the length of a transmission line to be used as a stub in a hairpin match. Typically, you will change only one item between test cases, either the test case frequency or a variable value, but

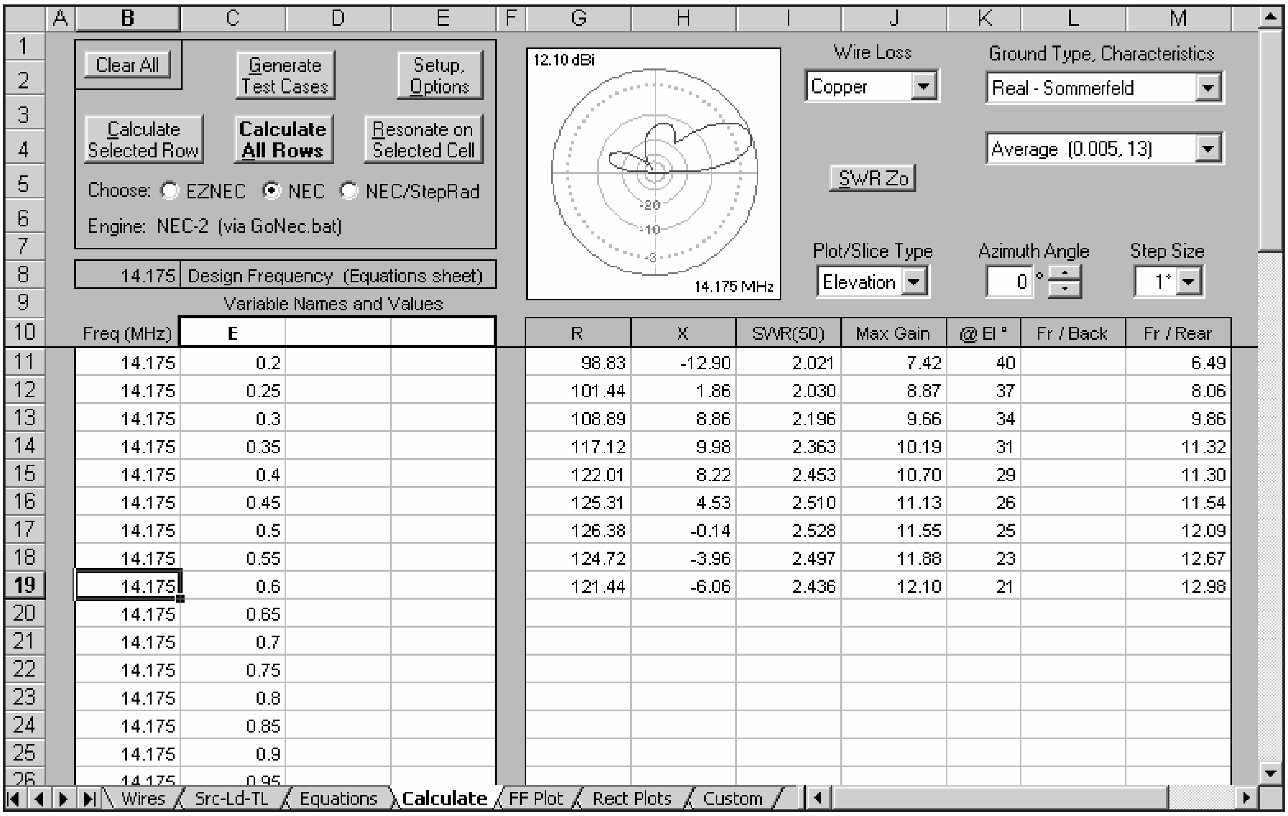


Fig 5—The right side of the *Calculate* sheet shows the calculated results for each test case, at the specified frequency and with the specified variables (if any) having the particular value(s) shown.

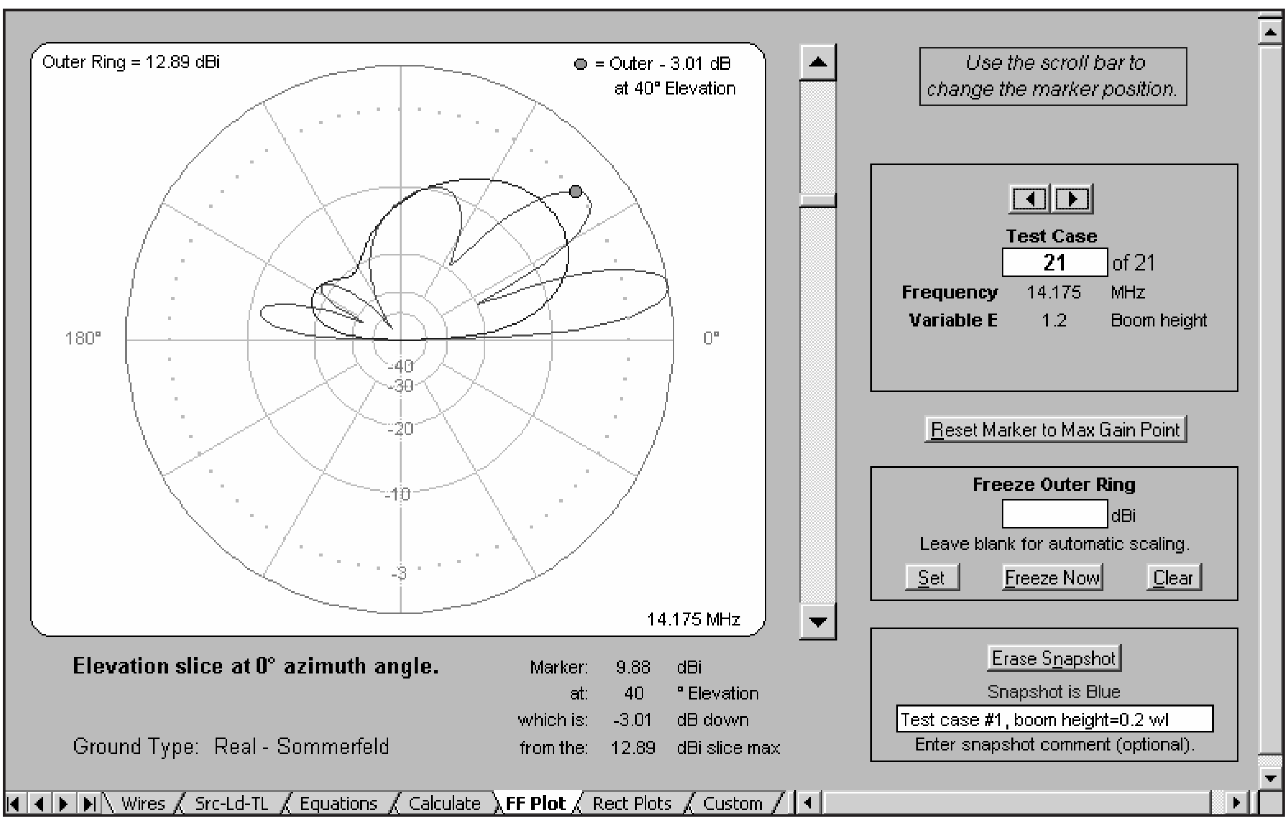


Fig 6—The *FF Plot* sheet shows azimuth or elevation radiation patterns. By holding down the spin button it is possible to see an "animation" of the complete set of patterns for a range of test cases.

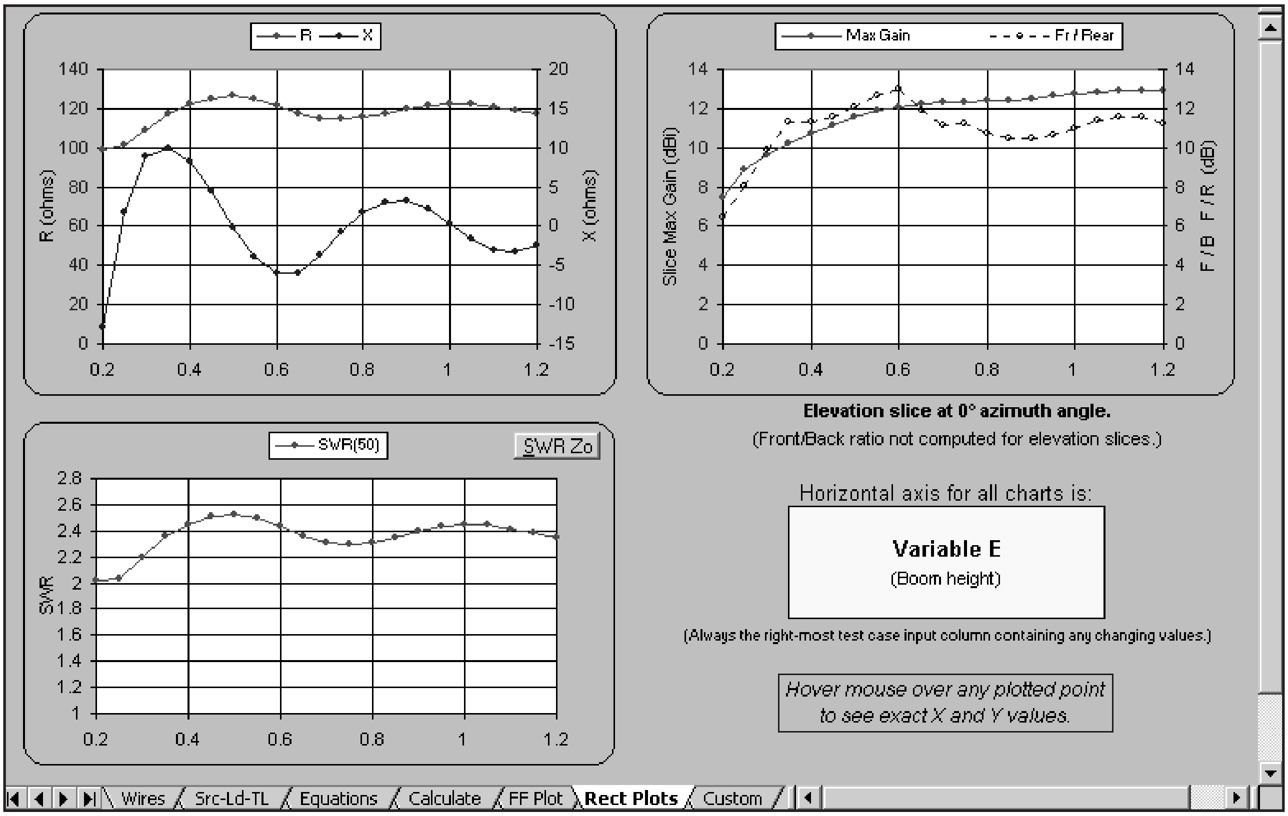


Fig 7—The *Rect Plots* sheet shows three standard rectangular plots. The horizontal scale for all plots is the same and will be either frequency or a changing variable value.

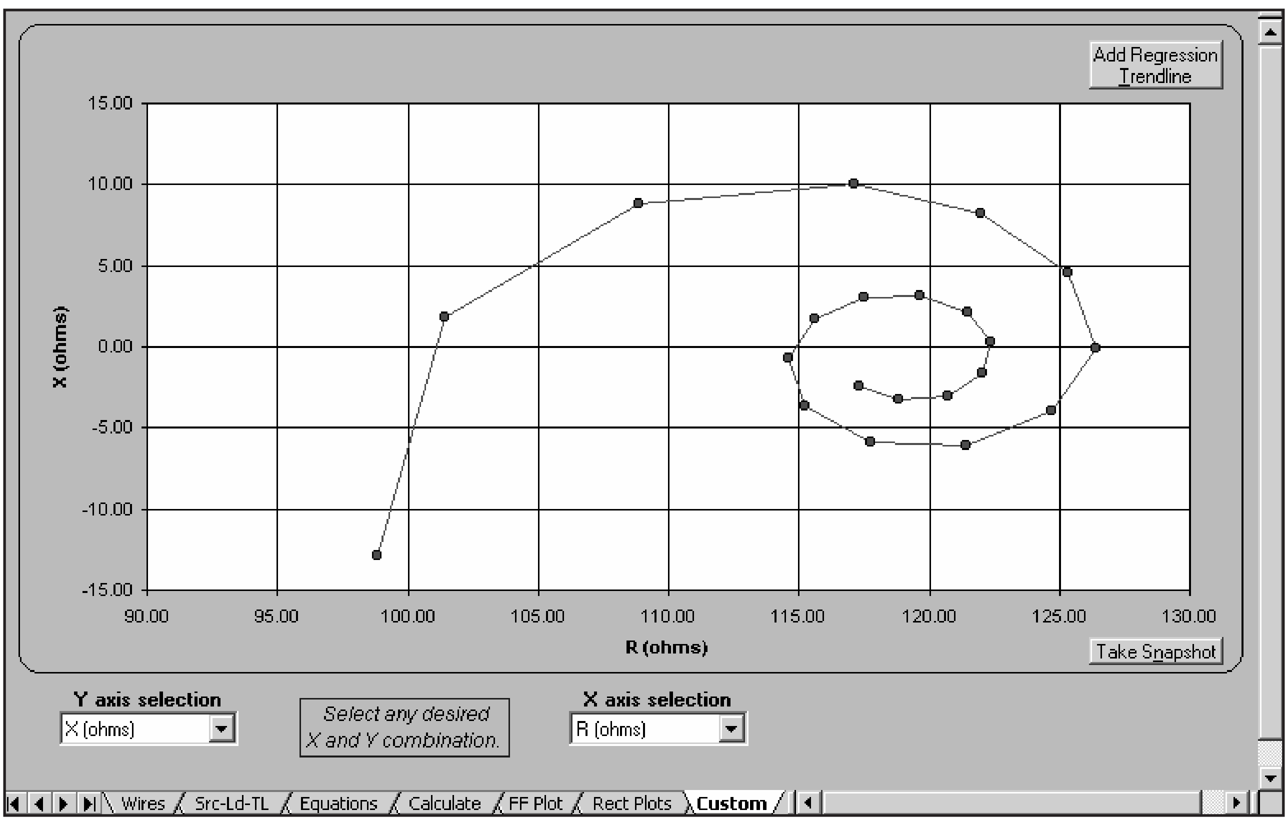


Fig 8—The *Custom* sheet can show any combination of the columns of data on the Calculate sheet. It can also be used to show a regression analysis curve and equation for the plotted points.

this is not an absolute requirement.

Step 3, run the simulations. This is usually done by clicking the **Calculate All Rows** button. For each row (each test case), *MultiNEC* will first change the value for the corresponding variable(s) on the Equations sheet to match the value specified for that row on the Calculate sheet. Then an antenna model will be constructed using the current values for all variables and at the specified test-case frequency. That model will be sent to the designated engine for processing. The resulting output data file will be read by *MultiNEC*, several items will be extracted from the file, and those items will be written to the right side of the Calculate sheet in the same row as the input data. The radiation-pattern data points will be collected, the thumbnail polar plot will be updated as each test case is completed, and the data points will be saved for later playback on a larger polar plot. The whole process will be repeated for each defined test case.

The various drop-down boxes and buttons at the top-right of the Calculate sheet define those modeling parameters that apply to all test cases. This includes the ground type and ground characteristics, the wire loss (if any) assigned to all wires in the model, the type of plot pattern to be generated and the reference base to be used in calculating the SWR at the source.

Step 4, interpret the results. This is done using four different sheets:

Calculate (Fig 5): As each test case is run, the numeric results are shown on the *right side* of the Calculate sheet and the thumbnail far-field plot is updated. The Fig 5 screen shot was taken just after the test case on row 19 was completed. The thumbnail plot shows the elevation slice radiation pattern when variable *E* (boom height, in wavelengths) has a value of 0.6.

FF Plot (Fig 6): After all test cases have been run, you may switch to the FF Plot sheet. Initially this sheet will show the azimuth or elevation plot for the first test case, corresponding to row 11 on the Calculate sheet. You may use the spinner to cycle through the other test cases or you may enter a desired test case number directly. By holding down the spinner, you may "animate" the playback of the patterns. The screen capture in Fig 6 shows the last test case in the set. Variable *E*, the boom height in wavelengths, has a value of 1.2. The red trace shows the elevation pattern for that test case.

The scroll bar is used to position the marker to a particular azimuth or elevation angle and see detailed information concerning that point on the plot. The plot is scaled with the ARRL modified log scale, the same as the default for *EZNEC* polar plots.

The **Take Snapshot/Erase Snapshot** button makes it easy to capture and freeze a plot trace (shown in blue) so that you may compare that trace to others from the same set of test cases (as in Fig 6) or to plots generated with different parameters. For example, you could request an elevation plot, take a snapshot, then request an azimuth plot and show the results superimposed on the same chart. The snapshot trace will remain in place until it is manually erased, and it will auto-scale along with the primary trace.

Rect Plots (Fig 7): This sheet contains three standard rectangular plots: (1) R and X, (2) SWR and (3) Max Gain, Front/Back Ratio, Front/Rear Ratio. (Front/Back is calculated for azimuth patterns only.) The horizontal scale for all three plots is the same and is automatically set by *MultiNEC*. It will typically be whatever the changing value was between the test cases, such as antenna height in the Fig 7 screen capture. If no variables were changed between test cases on the Calculate sheet and instead the frequency was changed, the horizontal scale would be frequency, and the plots would be equivalent to those you may have seen in other antenna modeling programs.

Custom (Fig 8): This sheet can be used to build a completely free-form rectangular plot. Both the horizontal (*X*) and vertical (*Y*) scales may be set to your choosing, picking from frequency, any variables that were set on the Calculate sheet and any data items extracted from the simulation output data files.

You can invoke a built-in *Excel* function to do regression analysis on the plotted points and show both the resulting curve (called a **Trendline** in *Excel* terms) and the corresponding equation. You may choose from several different regression types including Linear, Logarithmic, Power, Exponential and Polynomial from second to sixth order.

The **Take Snapshot** button works exactly like the one on the FF Plot sheet. You can capture any trace and compare it against any other trace. For example, you could plot SWR(50) versus frequency, take a snapshot, then change the SWR reference to 75 Ω and compare the traces. (The SWR reference base may be changed at any time without the need to recalculate all the test case results.) Or you could plot gain versus some other parameter, take a snapshot, change the geometry of the model, recalculate the test cases, then compare the old gain curve to the new one.

Finding Antenna Resonance

There is an additional way to run multiple calculations for an antenna model besides defining multiple test cases on the Calculate sheet. Instead, you may use the **Resonate** button. That button will cause automatic changes to either the frequency for a single test-case row or to a variable value for a single row such that the resulting source reactance value is within a range of ±0.1 Ω.

If frequency is changed, the other aspects of the model such as the geometry coordinates will be held constant at the current values. On the other hand, if a variable value is changed while frequency is held constant, that variable can control anything that varies reactance. This could include the length of a dipole, the circumference of a loop, the inductance of a loading coil or coils, or any other aspect of the model that produces a change in the source reactance. Hence, the **Resonate** button may be used to automatically answer not only "What's the frequency at which my 34-foot dipole is resonant?" but also "What's the exact length I need for a dipole if it is to be resonant at 14 MHz?"

MultiNEC does not just try random values to answer such questions. Three test cases are calculated (under the covers) and the three sets of results are used to perform a second-order-polynomial regression analysis. The Y intercept of this analysis is used as the "best guess" for a second set of three test cases. This process is repeated until the antenna resonance point is found, with automatic checkpoints to allow for manual intervention in case the process is not converging.

What *MultiNEC* is *Not*

MultiNEC may be used by itself. However, you may find it more convenient to use in conjunction with other antenna modeling programs because *MultiNEC* does not include *all* the functions you may want. Specifically, *MultiNEC* does *not* have:

An antenna model viewer: *MultiNEC* has built-in interfaces to the *EZNEC* view panel, the Nittany *NecVu* module, the *Antenna Model* viewer, and to the viewer that is included in the very nice freeware program *4nec2*.[4]

A wire geometry guideline checker: Guideline checks are done by all of the above viewers. *MultiNEC* itself does only the most basic checks such as making sure that the model contains at least one source.

A built-in stepped diameter correction module: NEC-2 produces inaccurate results when modeling connected wires of different diameters. A correction (the "Leeson" correction) may be applied in certain cases, most notably when modeling Yagi antennas constructed with telescoping sections of aluminum tubing. If you run your test cases using *EZNEC* as the calculating engine this correction will be applied automatically, in situations where it is possible to do so. If you run using any other *NEC-2* engine you must make available a separate correction module, such as the one included with *NEC-Win Plus+*, and explicitly request that it be used.

A viewer for currents on wires: *MultiNEC* contains no facilities for examining the currents on the wires of your model. You may use the *EZNEC* view window for this purpose. You may also use the *NecVu* or *4nec2* facilities for viewing currents on wires, although in that case you must first export the model and then use the standard user interface to open and process it with the other program.

A detailed reference guide for antenna model construction: The *EZNEC* user manual is a treasure chest of information on antenna modeling and serves as a very good introduction to the subject. Since the user-interface portion of *MultiNEC* is, by design, very similar to *EZNEC* (at least on the "input side" of the process), spending a few hours reading the *EZNEC* manual will be time well spent. The documentation provided by Nittany Scientific and Teri Software for use with their products serves the same purpose. Users of any of these commercial programs will have no trouble adapting to the *MultiNEC* interface.

Download

The *MultiNEC* package is shareware. You can find additional details and a downloadable zip file at **www.qsl.net/ac6la**. Since a public-domain version of the *NEC-2* calculating engine is bundled with the package, you need not have any of the commercial modeling programs installed in order to use *MultiNEC*. However, you must have *Excel 97* or later (*Excel 2000* or *Excel 2002/XP*) available on your PC.

Notes

[1] *NEC-Win Plus+*, *NEC-Win Pro*, and *GNEC* are products of Nittany Scientific Inc; **www.nittany-scientific.com**.

[2] *EZNEC* is a registered trademark of Roy W. Lewallen, W7EL; **www.eznec.com**.

[3] *Antenna Model* is a trademark of Teri Software; **www.antennamodel.com**.

[4] *4nec2* is a free *NEC-2* antenna modeling program. For additional details see **www.qsl.net/wb6tpu/swindex.html**.

Dan was first licensed in 1969 as WN3LVE. His Elmer, W3ZQU, also happened to be the father of his high school girlfriend; this was cause for consider! It was even suggested that Dan actually wanted the FCC to issue a special 1×4 call (W3LOVE) when he upgraded.

Dan holds a BSEE from the University of Pennsylvania. His radio-related spare time is devoted to QRP operating and writing antenna modeling and related software just for the fun of it. In the summer months, he spends a considerable amount of time on 2 meters operating from a somewhat unusual mobile location, his hang glider.

Professional Path Analysis Using a Spreadsheet

Use any common spreadsheet application to do your path-analysis number crunching.

By James Lawrence Sr, WB5HVH

Modern spreadsheet programs have the power to compute and to present line-of-sight radio-path analyses. The accuracy of the predictions and the quality of the presentations can exceed that obtained with commercial software. By developing your own spreadsheet application, you can supplement or replace your current path-analysis application. Capable spreadsheet programs can be obtained for a modest price or even free of charge.[1]

The formulas and methods required to build a path-analysis spreadsheet application are in the public domain and are readily discovered using the Internet or your public library. Virtually any modern spreadsheet program has the horsepower to handle the straightforward math. The only thing remaining is "elbow grease" and ingenuity. This paper will explain the math and the graphics required to build your own application for frequencies ranging from VHF to G-band microwave. It is based on a working application with field-proven results. The resulting application is accurate, inexpensive, simple to use and easily transportable.

[1]Notes appear on page 44.

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Cover Sheet

Having come from a telecom and control-systems engineering background, I prefer the "data sheet" format for presentation. This format lends itself nicely to the path analysis application. Also, I prefer a "3D" spreadsheet approach. That is, multiple "sheets" within a "book," with each sheet having a specific purpose. Fig 1 shows the cover sheet of my multi-sheet example application. Notice the datasheet format.

The "cover sheet" provides for user input, and it presents the numeric and graphic results. At the top of the sheet, there is a typical data-sheet area for title, revision history and other documentation. Both ends of the path (the sites) are documented in parallel columns. Just below the user input, there is a section for the numeric results. At the bottom, there is the obligatory visual representation of the path complete with ground elevation, foliage and Fresnel clearances.

Only the cells for user input are left unprotected. All other cells are protected for the purpose of document integrity. Calculation formulas may be hidden if desired. This portion of the document can be left austere or embellished with "pull-down" choices for user input if desired.

The example presented in this paper calculates path loss from site "A" to site "B" only. If there are separate transmit and receive antennas at the sites,

WIDGET LOGO HERE **WiDGi-CORP** **COMUNICATIONS DEPARTMENT** **RADIO HOP DATA SHEET**	**REV**	**BY** **DATE**	**REVISION DESCRIPTION**	**FILE NAME** S-Path.xls — **REV** **A**
	A	J.R. Lawrence 29-May-02	EXAMPLE	**HOP NAME** SITE A to SITE B
				HOP TYPE 2.4 GHz Spread Spectrum
				BY jrl — **CHKD** — **APRVD**

			SITE A	SITE B
GEN	01	SITE NAME	SITE ALPHA — **SITE A**	SITE BETA — **SITE B**
	02	ADDRESS	#1 ALPHA DRIVE	#2 BETA COURT
	03	CITY, STATE, ZIP	ALPHA HILLS, TEXAS 78957	BETA VALLEY, TEXAS 77478
	04	NORTH LATITUDE — NAD 27	32 DEG 26 MIN 52.0 SEC *32.448* DEG	32 DEG 20 MIN 23.0 SEC *32.340* DEG
	05	WEST LONGITUDE — NAD 27	87 DEG 46 MIN 5.0 SEC *87.768* DEG	87 DEG 55 MIN 1.0 SEC *87.917* DEG
TOWER	06	ELEVATION / TOWER HEIGHT / HEIGHT AGL	195 FT / 360 FT / *555* FT	310 FT / 270 FT / *580* FT
	07	CALCULATED IN METERS	*59.4* M / *109.7* M / *169.2* M	*94.5* M / *82.3* M / *176.8* M
	08	MANUFACTURER	ROHN	ALLIED
	09	MODEL	TYPE 45	BETA
	10	FCC/FAA TOWER NUMBER	87654321	12345678
	11	OTHER INFO	GUYED TOWER W/STROBE AND DIAL-UP ALARM	SELF SUPPORTING W/STROBE AND TOWER WATCHER
ANTENNA	12	TYPE	SOLID DISH W/RADOME	GRID DISH
	13	MANUFACTURER / MODEL		
	14	SIZE / BEAMWIDTH / GAIN	6 FT *1.8* M / 4.8 DEG / 30.8 dBd	8 FT *2.4* M / 3.6 DEG / 33.3 dBd
	15	HEIGHT AGL / BEARING	350 FT *106.7* M / ***229.4*** DEG A to B	260 FT *79.2* M / ***49.4*** DEG B to A
	16	OTHER INFO / POLARIZATION	HORIZ	HORIZ
FEED LINE	17	TYPE		
	18	MANUFACTURER		
	19	MODEL / SIZE		
	20	LENGTH 1 / FREQ / ATTEN. / 100 FT	360 FT / 2455.0 MHz / 2.22 dB / 100 FT	270 FT / 2455.0 MHz / 2.22 dB / 100 FT
	21	LENGTH 2 / FREQ / ATTEN. / 100 FT	30 FT / 2455.0 MHz / 5.93 dB / 100 FT	30 FT / 2455.0 MHz / 5.93 dB / 100 FT
	22	OTHER INFO		
RADIO	23	MANUFACTURER		
	24	MODEL / SERIAL NO.		
	25	TRANSMITTER / FREQ / GAIN	2455.0 MHz / 27.0 dBm	2455.0 MHz / 27.0 dBm
	26	RECEIVER / SENS	-94.0 dBm	-94.0 dBm
	27	OTHER INFO		

RESULTS	28	HOP DISTANCE / HOP FACTORS	***11.46*** MILES	*18.44* KM	0.40 (0.125 DRY to 0.500 HUMID)	1.00 (4.0 SMOOTH to 0.25 MOUNTAINS)	
	29	SYSTEM GAINS: SITE A TO SITE B	*27.0* dB XMIT	*94.0* dB XCEIVE	*30.8* dB "A" ANT	*33.3* dB "B" ANT	*185.2* TOTAL
	30	SYSTEM LOSSES	*131.1* dB PATH	*9.8* dB COAX "A"	*7.8* dB COAX B	*2.8* dB OTHER	*151.4* TOTAL
	31	FADE MARGIN / AVAILABILITY / DOWNTIME	***33.8*** dB	***99.99985%***	***49*** SEC / YR or	0.8 MIN / YR or	0.014 HR / YR

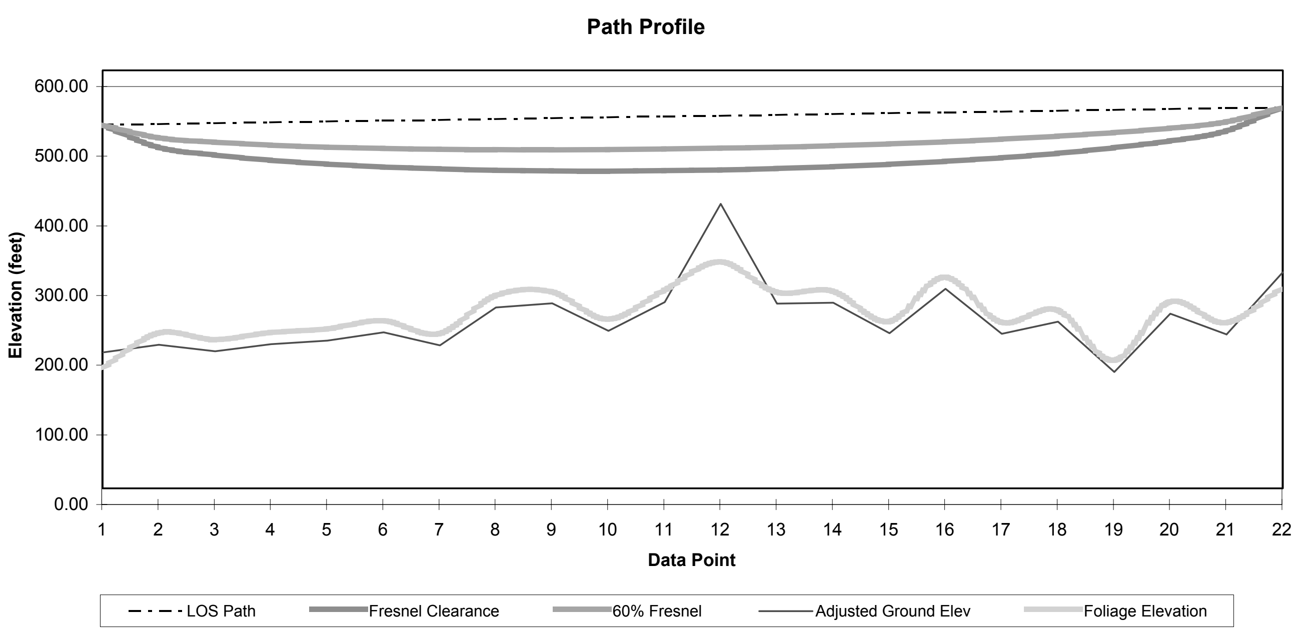

Fig 1—The author's cover sheet for his path-analysis spreadsheet.

it may be desirable to calculate from site B to site A also. Creating another instance of the file and swapping the site A and B parameters can easily do this. One could also get crafty and enhance the application to calculate both cases. Also, path diversity is not covered but can be readily handled if required.

As previously stated, my approach is to place the logical calculation groupings on separate sheets. This breaks the problem down into easy-to-manage sections. Each separate calculation sheet pulls data from the cover sheet, completes the calculations as required and pushes the result back up to the cover sheet. Examples of discrete calculation sheets are "Bearing and Distance," "Path Loss" and "Availability."

Bearing and Distance

For the bearing-and-distance calculations, you will need the latitude and the longitude of each site as input. The required output is the bearing from site A to site B, the bearing from site B to site A, and the distance between the sites in various units including Earth surface arc.

The latitude and longitude are available from user input on the cover sheet. The calculation will require these values in decimal format. Here is a method for making the latitude/longitude conversion to decimal.

=Cover!R24+((Cover!V24)+(Cover!Z24/60))/60

In this example, the values for "Degrees," "Minutes" and "Seconds" reside on the cover sheet in cells R24, V24 and Z24 respectively. Error checking can be implemented to ensure that the user does not enter values outside the valid range, that is "61 seconds."

Now that we have the latitudes and longitudes from user input and we have made the conversion to decimal, let us look at the distance calculation. The distance between any two points on the surface of a sphere can be determined with the formula:

$$\cos D = \sin A \sin B + \cos A \cos B \cos L \qquad \text{(Eq 1)}$$

where

D = distance in degrees of arc
A = site A latitude in decimal degrees
B = site B latitude in decimal degrees
L = site A longitude minus site B longitude

Since A, B and L are known, the value for distance is easy to calculate. Although "degrees of arc" will be required to find the bearing, it is not very useful from the human perspective. To resolve this, here are some useful conversions:

111.23 km = degree of arc (for this planet only)
0.621371 mi = 1 km

Now, let's consider the bearing calculation. The bearing from site A to site B can be obtained with the following formula:

$$\cos C = (\sin B - \sin A \cos D) / (\cos A \sin D) \qquad \text{(Eq 2)}$$

where

C = bearing from North
D = distance of arc between the sites, in degrees of arc (from the distance calculation above)

Again, this is a simple calculation, but there are caveats. First, the raw result only works half of the time. Second, know your spreadsheet's result format (degrees or radians) and make the necessary conversion when required. Knowing the sine of L allows you to resolve the "half the time" problem. Here is a way to work it all out:

=IF(E40 < 0, 360 – E46, E46)

confirms the raw bearing calculation from site A to site B. This means, "If the sine of L is less than zero, subtract the raw bearing from 360°, otherwise the answer is good.

In this calculation, cell E40 contains sin L and cell E46 contains the raw bearing in degrees,

=IF(E47 > 180, E47 – 180, E47 + 180)

derives the bearing from site B to site A. If the confirmed bearing is greater than 180°, subtract 180°, otherwise add 180°. Where cell E47 contains the confirmed bearing from site A to site B.

Please note that this calculation example works in north latitude and west longitude. It may require tweaking for the rest of the planet. This completes the bearing and distance calculations.

Path Loss

Next, we will tackle path loss. Path loss is loosely defined as the loss in signal associated with the line-of-sight or free-space distance between sites A and B. The formula for free-space path loss is readily available in textbooks and vendor catalogs.

$$L_{fs} = K + 20 \log f + 20 \log D \qquad \text{(Eq 3)}$$

where

L_{fs} = loss in decibels
K = pseudo constant (based on frequency)
f = frequency in gigahertz
D = distance in miles

We have already calculated the distance between the sites. The frequency of the hop is known from user input. The only thing remaining is the constant, K. In general, K is equal to 96.6 dB plus some small frequency-dependent amount or addend. This would be a good application for a "pull-down" pick-list. My path-analysis application was only used within a known range of frequencies, so I simply interpolated between the lowest and the highest "frequency addend." The results came out very close to those obtained with a commercial application.

Availability

Another required calculation sheet is that of availability. There are several mathematical models for availability. I chose Barnett's:[2]

$$Av = (1 - a \times b \times 10^{-5} \times f/4 \times d^3 \times 10^{-FM/10}) \times 100\% \qquad \text{(Eq 4)}$$

Where

Av = availability (uptime) expressed in percent
a = terrain roughness factor
b = climate factor
f = frequency in gigahertz
d = distance between sites in miles
FM = fade margin

This one is not tough, but it requires a bit more explanation. Terrain roughness and climate factors are subjective input from the user on the cover sheet. The terrain-roughness factor varies between 0.25 for mountains (rough) and 4.0 for Groom Lake (smooth). The climate factor varies between 0.125 for Groom Lake (dry) and 0.50 for Houston (humid). Again, frequency and distance are known values at this point. The last variable is fade margin. Fade margin is simply the sum of all gains minus the sum of all losses. This reminds me of one of the two "rules of engineering": "Some of it plus the rest of it equals all of it."[3] Fade margin will be discussed in more detail later; but for the sake of this calculation, its value resides in a cell on the cover sheet.

Miscellaneous Gains and Losses

There needs to be a place for the user to apply miscellaneous gains, losses and "guesstimates." This will be the last calculation sheet. It is an exception to the "user input on cover sheet" rule. Examples of losses would be connectors, lighting arrestors and old coaxial cable. Gains could include power amplifiers or receive preamplifiers. Just add it all up and stick the total on this sheet.

Path Profile

The most complex portion of creating this path-analysis application is the graphic presentation of the path profile. As a minimum, the path profile should show the straight line of sight between the two antenna elevations, the 60% Fresnel zone and the ground elevations corrected for earth curvature. I choose to add the full Fresnel zone and average foliage height plus an option for "point addends" to represent buildings or towers that might protrude through foliage and possibly into the Fresnel zone.

Like most graphs, this profile is created from a table of values. Some of the values are borrowed from user input on the cover sheet, some are input by the user into the table and most are calculated. The last exception to "user input on the cover sheet" is elevation data taken from a topographic map. Yes, you need a "topo" map to do a proper path analysis.[4] Even when you use the electronic topographic information from commercial programs, you should check it against actual maps. I have found that electronic three-second data are not suitable for a reliable line-of-sight analysis. Peaks in or near the path are often missed and I have found endpoint elevations to be off by over 100 feet. Fig 2 is an example of a table of values.

In Fig 2, the shaded areas are protected; user input is allowed in the "Ground Elevation" and the "Point Addend" columns. Here is an explanation of the table column by column:

Data Point: Data points A and B represent the elevations at the two sites. I chose to use a total of 20 data points between sites because that provides adequate resolution for most paths of 900 MHz and above.

Distance d1 (miles): This is simply the distance from each data point to site "A." These are equidistant segments of the entire path length, (in this case, 11.46 miles divided by 21 or 0.5457 miles each). Here is the easy way. Data Point 1: "= $^1/_{21}$ × D36" (where cell D36 contains the path length in miles). Data Point 2: "= $^2/_{21}$ × D36," and so on. This data column will be used in the earth-curvature calculation.

Ground Elevation: The ground elevations for points A and B come from user input on the cover sheet. Those for the other 20 points are read from a topo map. The user simply draws a line between the two Sites on the topo map (or maps), divides the line into 21 equal parts, and reads the highest elevation at, *or near,* each of the 20 resulting points.

Distance d2 (miles): This is the same data as distance d1 but from site B. This data column will also be used in the earth-curvature calculation.

LOS Path: The line-of-sight path column will draw a straight line between the antenna AGL (above ground level) elevations on the graph. The data for sites A and B are simply the corresponding ground elevation from that data column plus the antenna height (AGL) from the cover sheet. Each of the 20 data points builds on the previous point. For example, the Data Point 1 calculation would be "=F12 + ((F33 – F12) / 21)" where cell F12 represents Data Point A, and cell F33 represents Data Point B. The next calculation for Data Point 2 would be "=F13+((F33-F13)/21)" and so on.

Fresnel Clearance: Fresnel clearance elevation for each data point can be calculated with this series of calculations:

Data Point A: =F12 – 72.1 × SQRT((C12 × (D36 – C12)) / (D35 × D36))

Data Point 1: =F13 – 72.1 × SQRT((C13 × (D36 – C13)) / (D35 × D36))

Where cell D35 contains the frequency in gigahertz, and cell D36 contains the path length in miles.

I leave it to you to research and confirm this calculation

A	B	C	D	E	F	G	H	I	J	K	L
9	Path Information										
10	Data	Distance	Ground	Distance	LOS	Fresnel	60%	Adjusted	Foliage	Earth	Point
11	Point	d1 (miles)	Elevation	d2 (miles)	Path	Clearance	Fresnel	Ground Elev	Elevation	Curvature	Addend
12	A (1)	0.00	195	11.46	**545.00**	**545**	**545**	**195.00**	**195.00**	0	
13	2	0.55	203	10.91	**546.19**	**513**	**526**	**205.98**	**245.98**	2.98	
14	3	1.09	191	10.37	**547.38**	**502**	**520**	**196.66**	**236.66**	5.66	
15	4	1.64	199	9.82	**548.57**	**494**	**516**	**207.04**	**247.04**	8.04	
16	5	2.18	202	9.28	**549.76**	**489**	**513**	**212.12**	**252.12**	10.12	
17	6	2.73	212	8.73	**550.95**	**485**	**511**	**223.91**	**263.91**	11.91	
18	7	3.27	192	8.18	**552.14**	**482**	**510**	**205.40**	**245.40**	13.40	
19	8	3.82	245	7.64	**553.33**	**480**	**509**	**259.59**	**299.59**	14.59	
20	9	4.36	250	7.09	**554.52**	**479**	**509**	**265.48**	**305.48**	15.48	
21	10	4.91	210	6.55	**555.71**	**479**	**509**	**226.07**	**266.07**	16.07	
22	11	5.46	251	6.00	**556.90**	**479**	**510**	**267.37**	**307.37**	16.37	
23	12	6.00	292	5.46	**558.10**	**480**	**511**	**408.37**	**348.37**	16.37	100
24	13	6.55	249	4.91	**559.29**	**482**	**513**	**265.07**	**305.07**	16.07	
25	14	7.09	251	4.36	**560.48**	**485**	**515**	**266.48**	**306.48**	15.48	
26	15	7.64	208	3.82	**561.67**	**488**	**518**	**222.59**	**262.59**	14.59	
27	16	8.18	273	3.27	**562.86**	**492**	**521**	**286.40**	**326.40**	13.40	
28	17	8.73	210	2.73	**564.05**	**498**	**524**	**221.91**	**261.91**	11.91	
29	18	9.28	229	2.18	**565.24**	**504**	**529**	**239.12**	**279.12**	10.12	
30	19	9.82	159	1.64	**566.43**	**512**	**534**	**167.04**	**207.04**	8.04	
31	20	10.37	245	1.09	**567.62**	**522**	**540**	**250.66**	**290.66**	5.66	
32	21	10.91	218	0.55	**568.81**	**536**	**549**	**220.98**	**260.98**	2.98	
33	B(22)	11.46	310	0.00	**570.00**	**570**	**570**	**310.00**	**310.00**	0	

Frequency	2.455	GHz
Path Length	11.46	miles
Curvature Height	h = d1*d2 / 1.5k	feet
k	1.33	4/3 (default)
Foliage Height	40	feet
Beam elevation	0.001900949	radians

Fig 2—The last worksheet of the project accepts user input that is not handled by the cover sheet. User input goes in the areas clear of shading.

method. It is readily verifiable. This data column draws the bottom of the Fresnel zone curve on the path-analysis graph.

60% Fresnel: This is simply the Fresnel data multiplied by 0.6. This data column draws the 60% Fresnel-zone curve on the graph.

Adjusted Ground Elev: This data column is simply the ground-elevation column plus the addend for Earth curvature plus any manual addends from the last column. This data column will draw the ground elevation line between each of the tower bases.

Foliage Elevation: This data column is the ground elevation plus the addend for Earth curvature plus a user-input constant for average foliage height. In this example, an average foliage height of 40 feet was used.

Earth Curvature: The formula for calculating earth curvature is $h = d1 \times d2 / 1.5\,k$. The values for $d1$ and $d2$ for each data point have previously been calculated. "k" is typically $4/3$ or 1.333; "k" should be set as user input with a default to 1.333.

Point Addend: This data column is for user input. The example shows 100 feet added to data point 11. Notice how this is depicted in the graph; there is a protrusion through the average foliage at a point approximately 6 miles from site "A." This could represent a tower or a building. The example proves that it does not penetrate the Fresnel zone.

This completes the explanation of the data table. All that remains is to define the graph. The process would depend upon the graph application used, but it should be straightforward. Be sure to let the graph adjust the Y-axis automatically. Once this is set up properly, the user only need enter the 20 elevations taken from the topo map. (Note: Spreadsheet programs with integral graphing capabilities are more convenient, but external graphing applications can be used if needed.)

Numeric Results

The last thing that needs to be done is to present the numeric results on the cover sheet. I used the area just above the graph for this purpose. The fade margin and the availability are the "bottom line" so I show them in bold type. As previously mentioned, the fade margin is equal to the system gains less the system losses. This is a good place to calculate and summarize those gains and losses. It is also nice to present the path distance in miles and kilometers.

A spin-off of availability is predicted downtime. This comes in handy when trying to explain costs; for example, $10,000 can buy you maybe 15 seconds a year. There are 31,557,600 seconds in a year.[5] One minus the availability times this number of seconds gives you the predicted downtime.

Conclusion

In summary, this paper provides the basic information required so that readers can create a line-of-sight path analysis application from a spreadsheet. The features and the presentation of your application are left to you. I strongly feel that a spreadsheet program, configured properly, can rival a commercial path-analysis application. A spreadsheet application similar to the one described in this paper has been tested repeatedly against commercial applications for the same path, and it consistently yields comparable results. If anything, the custom spreadsheet application makes more accurate and conservative predictions.

Notes

[1]Open Office, Gnumeric or KSpread for *Linux.*

[2]Barnett's model for outage or unavailability. "Engineering Considerations for Microwave Communications Systems," *AG Communication System*, 4th Edition, 1991, pp 68-69.

[3]The other rule is, "You can't push a rope."

[4]Topographic maps are available from several sources, overnight if necessary. Use the free Legend maps for each state to determine which topo or topos you will need for a path. I don't recommend using anything but the 7.5-minute series of maps. Other scales are too difficult to read. Topo map software is also available from several vendors and there are some topo maps available free of charge on the Internet.

[5]Do not forget leap years.

James received a Bachelors degree in Telecommunications from Texas A&M University, and is employed as a Principal Engineer in the Communications Department of a Fortune 100 corporation. He has been a licensed Amateur for almost 30 years.

A High-Level Accessory Front End for the HF Amateur Bands

Here's an antenna selector/preselector/attenuator/preamplifier accessory for every HF amateur transceiver. It can improve your receiver's IP2 for out-of-band signals and yield good flexibility.

By Sergio Cartoceti, IK4AUY

My father, I4FAF (an old-timer) and I both very much like Amateur Radio as a lifetime endeavor. We do not have backgrounds in electronic engineering, but we do have a lot of practice. My father is a fast builder of Amateur Radio projects, from printed-circuit artwork drawn by him with CAD software to working units in our home laboratory. Being retired now, he has more time and I help him from time to time.

Our goal is to get more from the Amateur Radio equipment available to us. Our gear is average, not top-priced. I want to improve my skills with DX or weak signals, while I operate in crowded bands during some international contests. Lately, I have discovered the low-frequency bands. They have added more fun.

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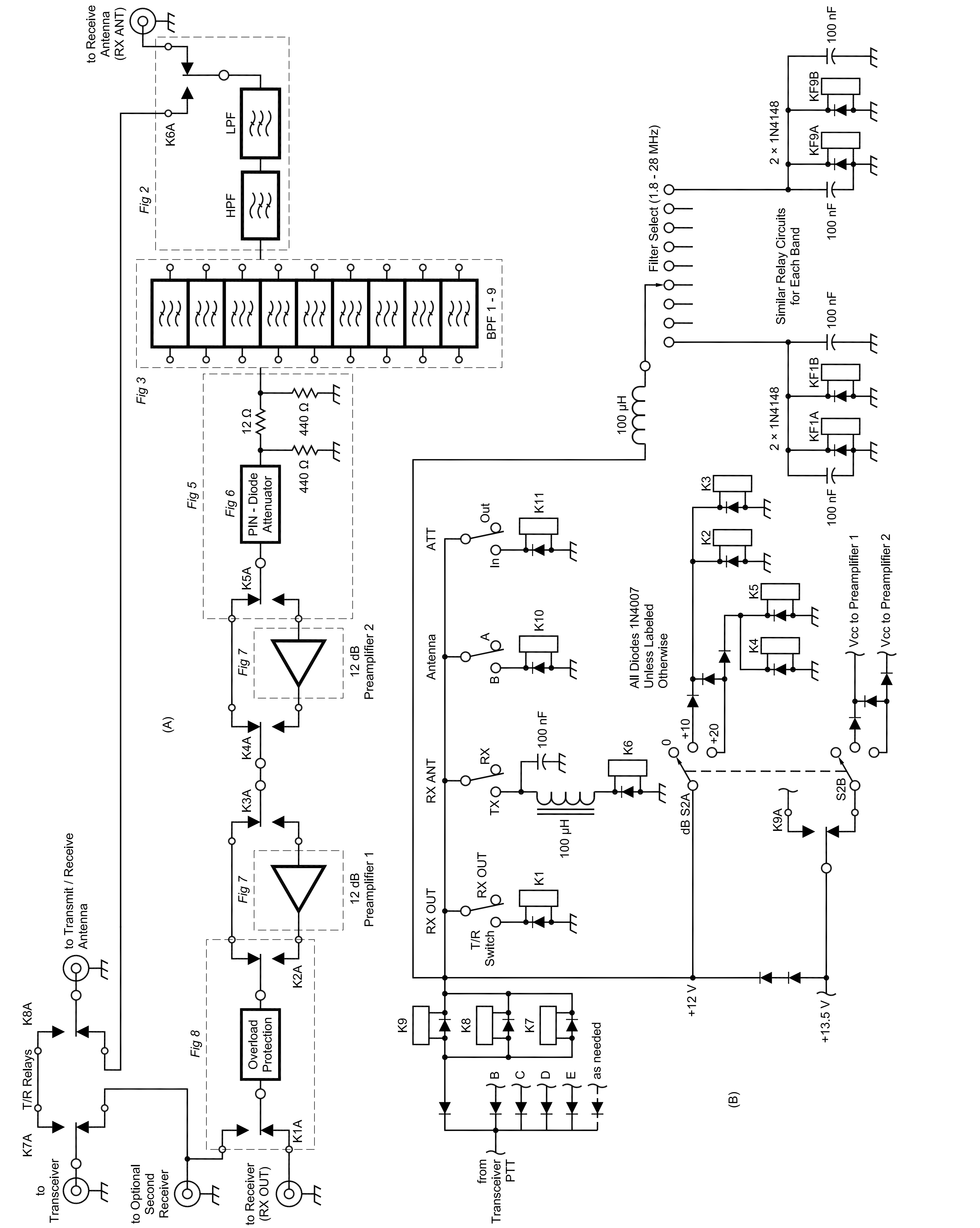
In 2000, we started to put up (at a flat, country location) a short vertical antenna by Butternut, the HF2V. It has the 160-meter coil kit, is top-loaded with four wires (each about 5 meters long) and has six ground radials about 40 meters long for 160, 80 and 40 meters. In winter 2001/2002, we started to test some receive-only antennas, with better signal-to-noise ratios and some directivity, in comparison to the 360° radiation pattern of a vertical antenna.

With the exception of Beverage and a four short vertical system presented by W8JI on his Web site, that still require much space for only –6 to –11dBi; most of the receive-only antennas we have considered are in the low-output category—in the range from –6 to –35 dBi. We have worked with Beverages, EWE, the delta-EWE by K6SE—a variation of pennant-flag antennas and K9AY loops.

We have seen them presented by our trusted teachers in recent articles in Amateur Radio publications. For example, ON4UN's *Low Band DXING* (third edition, don't miss reading the new chapter "Special Receiving Antennas"), *QST*, the *Antenna Compendium* series and K1ZM's *DXing On the Edge*—all published by ARRL. There has also been some follow-up on the Internet and on the top-band reflector from W8JI, WA2WVL (EWE antenna), K6SE (delta-EWE and other pennants), WA1ION (pennant with remote variable control of the resistive, in-line termination), K9AY (K9AY loops, now also with remote variable control), K3KY (his Web site has a full collection of



to Receive Antenna (RX ANT)
K6A
Fig 2
LPF
HPF
Fig 3
BPF 1 - 9
Fig 5
Fig 6
PIN - Diode Attenuator
12 Ω
440 Ω
440 Ω
K5A
Fig 7
12 dB Preamplifier 2
(A)
K4A
K3A
Fig 7
12 dB Preamplifier 1
to Transmit / Receive Antenna
K8A
T/R Relays
K7A
to Transceiver
to Optional Second Receiver
to Receiver (RX OUT)
Fig 8
Overload Protection
K2A
K1A
Filter Select (1.8 - 28 MHz)
100 µH
2 × 1N4148
KF9B
KF9A
KF1B
KF1A
100 nF
Similar Relay Circuits for Each Band
ATT
Out
In
K11
Antenna
A
B
K10
RX ANT
RX
TX
100 nF
100 µH
K6
RX OUT
T/R Switch
K1
All Diodes 1N4007 Unless Labeled Otherwise
K3
K2
K5
K4
0
+10
+20
dB S2A
S2B
K9A
Vcc to Preamplifier 1
Vcc to Preamplifier 2
+12 V
+13.5 V
K9
K8
K7
B
C
D
E
as needed
from Transceiver PTT
(B)

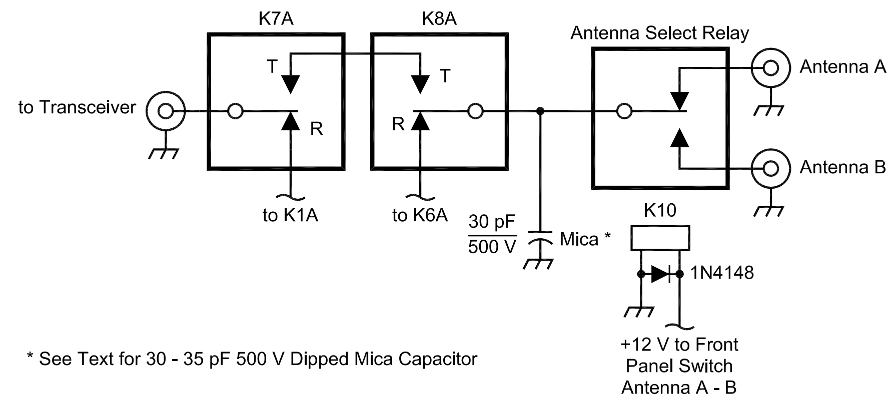


Fig 1—(left, A) A block diagram of the system. (B) PTT and front-panel controls. The band-pass filters are selected by relays and a front-panel band switch. Preamplifier gain is controlled by a front-panel switch. (C) An additional transmit antenna may be controlled with an added relay and switch.

Table 1—Preliminary Gain-Distribution

	Filters Loss (dB)	*Atten (dB)*	*Preamplifier Gain (dB)*
Stage Gain	–5	0	+12 dB
Total Gain	–5		+7
Stage *NF*	5	0	3*

*not measured. See Note 27.

contributions, links, about low band antennas), W7IUV (rotatable flag) and other well-known authors.[1-11]

There is a lot of interest and newcomers frequently ask, "What is the best receiving antenna for the low bands?" I like the Beverage very much, but my answer must be that I don't know, simply because, until now, I have not been able to test them all. Read K1ZM's book. He agrees that it's better for hams to have more types of antennas available on 160 meters. That is true because of the variable and peculiar propagation conditions on that band.[12-14]

Our Beverage and Other Receive-Only Antennas

We tested a 177-meter-long, unidirectional Beverage configuration (for USA), up about 2 meters above ground (rural terrain), with a 500-Ω end load (two 1000-Ω resistors parallel connected to a ground rod) and an input impedance ratio of 1:9. The transformer was made of seven quadrifilar turns in parallel using #20 AWG or 0.8-mm-diameter enameled copper wire. The core is an Amidon ferrite FT114-F with a permeability of 3000. Remember that this material, manganese-zinc, has a low bulk resistance, so it is best to cover the core with a thin layer of Teflon tape before winding the wire on it.

See John, ON4UN's, third-edition book for a transformer picture and photo on page 7-17, Fig 7-18: "modified transmission-line transformer." With our ground characteristics, we had better matching results with this ferrite mix than with the type 43 (permeability of 850) proposed in the book. The thing was tested by us with help of a new MFJ-269 and confirmed by our friend's laboratory-grade spectrum analyzer and tracking generator. ON4UN usually uses high permeability type MN-CX, which was not available to us. A very nice description of how to get a Beverage system to work properly is contained in K1ZM's book, on pages 12-1 to 12-6.

It had been up for only a few days of tests in the winter of 2001/2002 when a 160-meter CW-contest weekend came along. My impressions of performance were very favorable. I had a great time with this antenna. Around 10 US states were worked in one night and with a very clear copy over my short vertical as the transmitting antenna. I still remember those nice signals; of course, most of them were from well-equipped contest stations.

A simple Beverage antenna alone provides about –11 dBi. I don't know exactly the gain of my short vertical but the relative difference in level in decibels does agree quite closely. A phased system of two might have a higher output, at about –6 dBi, as per ON4UN. The other receive antennas' outputs can be, at worst, about –30 to –35 dBi for a delta-EWE loop; see more detailed data in the new chapter in ON4UN's book. We have tried the K6SE delta-EWE, a pennant receive-only antenna that has reasonable dimensions: total wire length is 72 feet, 28 feet on the base side and the high apex is about 17 feet from the base side. It is easy to make it rotatable. In that case, the transformer is differently placed in one lower corner; in the opposite corner is a 950-Ω series resistor and it matches 50 to 950 Ω. K6SE suggested a FT140-43 ferrite with primary and secondary wound at the opposite sides of the core. The primary is 8 turns and the secondary 34 turns with about 990µH to 1 mH using 20 AWG enameled wire. Remember that the directivity is in the opposite direction from the termination corner, toward the feed point, unlike the Beverage.

The Need for an Antenna Processor

We immediately realized the importance of making frequent checks on the receiving antenna and on the transmit/receive antenna to get more flexibility from the system. That is, to avoid overloading the inside equipment switches when the same functions (antenna 1, antenna 2 and receive-antenna selections) are already built into some recent radios. If needed, you can make maintenance of such switches easy if they are in an outside home-built unit. Now, we don't have the problem of switching among more than two antennas!

In practice, we felt immediately that we needed a complete independent accessory for our transceiver as an outboard tool to deal with the issue of better selectivity in the receive chain. So we stopped our antenna tests and started to think about the design of a complete HF front-end unit with band-pass filtering for our amateur bands only, not for general coverage. Just to simplify and avoid wasting time during contest activity with peaking controls, we decided as a practical tradeoff to choose fixed band-pass filters without variable controls. That's why we agree with G3RZP when he wrote in *QEX* May/June 2002, on page 40:

"… Are our receivers too sensitive? The answer is 'Probably, but…' There are some imponderables. On the LF bands especially, the use of separate receiving antennas producing much lower-level signals but also lower levels of noise means that requirements

[1]Notes appear on page 55.

may exist at times for the low noise-figure levels that are typically seen in modern receivers."

Obviously, the use of pre-mixer selectivity has a major effect on the performance requirements, although at 7 MHz, the proximity of the broadcast band offers little possibility of really effective filtering in conventional circuits. ". . . US conditions seem a lot quieter than those in UK." Thus a variable antenna attenuator has obvious advantages, but the attenuation steps need to be much smaller than the 6- or even 20-dB steps provided by commercial transceivers.

I must confirm that it is very hard for us in Europe. We work split on 40 meters to listen to DX and North America among very powerful broadcast AM stations in your portion of the band from 7.150 to 7.300 MHz. The test Peter performed was with a FT-102, which is not a general-coverage

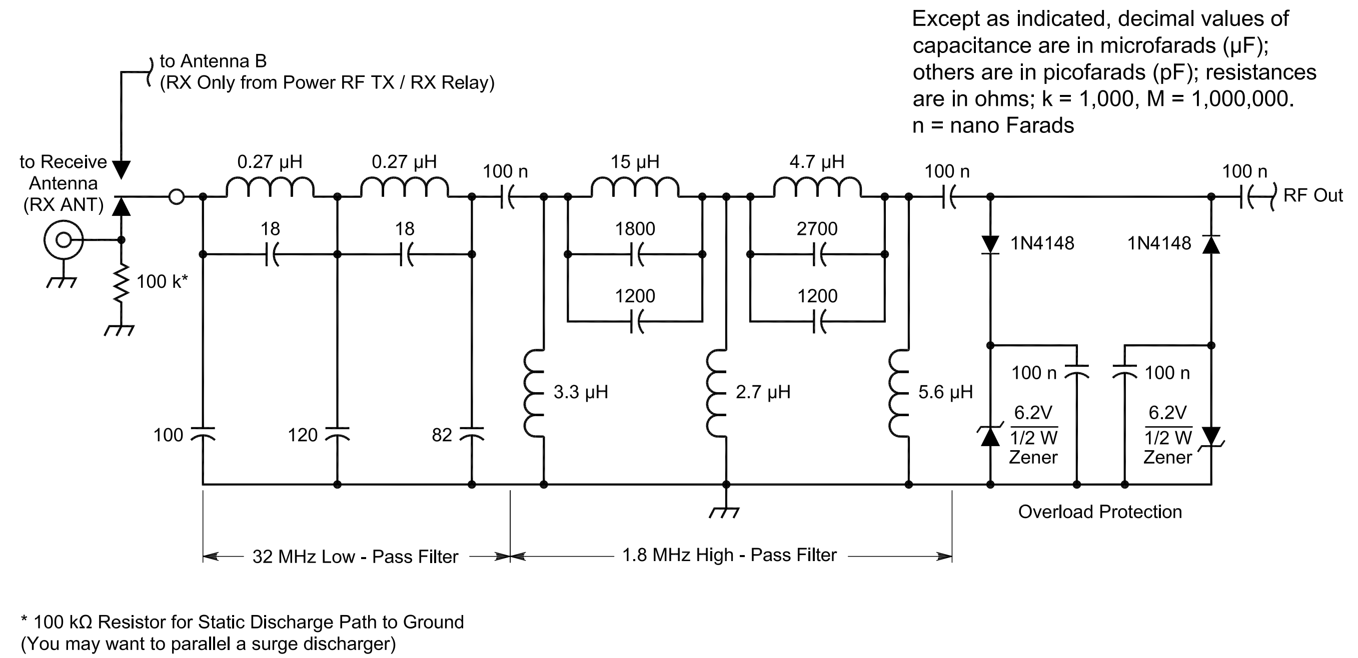


Fig 2—A schematic of the high- and low-pass filters. The 100-kΩ resistor near the RX ANT connector provides a path to ground for electrostatic discharges. You may want to add a surge supressor in parallel with it.

Table 2—Butterworth Band-Pass Filters with Three Toroidal Coils, Design Data, with In-Out Capacitive Divider for 50 Ω

Band (MHz)	*Toroidal Cores L1-2-3*	*Turns*	*Wire Diameter (mm)*	*μH*	*C1-C8 (pF)*	*C2-C7 (pF)*	*C (pF)*	*C4 (pF)*	*C5 (pF)*	*CR1-2-3 Var. C (pF)*	*CR Var C Diameter (mm)*	*Notes*
1.83	for 160 meter data, see the 1980 ARRL Handbook p 8-43 and refer to our dedicated board.											
3.7	T50-2	44	0.35	10	1000	150	5.6	150	10000	100	10	
7.07	T50-2	31	0.5	5	410	82	3.9	75	5000	65	10	
10.1	T50-2	24	0.6	2.8	300	68	3.3	56	2700	22	7	adapt board holes
14.2	T50-6	22	0.75	2	200	44	2.2	40	2400	22	7	
18.1	T50-6	20	0.8	1.6	150	33	1.5	33	1800	22	7	
21.2	T50-6	18	0.8	1.4	135	30	1.7	28	1250	22	7	
24.9	T50-6	16	0.8	1	120	27	1.5	22	1000	22	7	
28.5	T50-6	14	0.8	0.8	100	22	1.5	18	560	22	7	

Notes
- -CR1-CR2-CR3 Variable capacitors are small Philips film or Teflon dielectric.
- -Fixed capacitors are dipped mica (or good ceramic). Some USA suppliers: RF Parts, Surplus Sales of Nebraska. In cases where you don't have a value available, make a parallel combination to get as close as possible to the design value. In any case, check the real filter shape.
- -In and out relays +V dc line control must be connected together and to panel band switch for each band with some RF decoupling.
- -Use enameled copper wire of closest AWG to the metric size shown; see Table 4.
- -Toroidal cores are standard AMIDON at **www.amidon-inductive.com/** or Palomar types at **www.palomar-engineers.com/**.
- -The filters are constructed on glass epoxy, double sided circuit board.

receiver but has some ham-band preselectors in it. I don't want to use too much attenuation first if the rig used is even poorer.

First, I would like to try a band-pass filter in front of it with a moderate insertion loss, narrower than the internal one. Then, eventually, I will add more attenuation if needed.

A resistive or PIN-diode attenuator is by nature broadband. Insertion loss in a band-pass filter is already an attenuation of RF signals. Outside the filter passband, attenuation increases on both sides. A practical preselector is desirable in the front end of a receiver to protect all the following stages of the receive chain. In-band insertion loss shouldn't be too high, but 4-6 dB is acceptable since in many cases, you don't need the full sensitivity of your modern receiver. Only when band conditions permit can you switch in one preamplifier to compensate insertion loss.

I do believe that a good receiver must be designed for low IMD in all stages and should have narrow filters from the beginning of the chain so all the following stages are protected. If not, you need a better following chain. Some system gain-distribution consideration could be done, with one preamplifier switched in, attenuator off, as shown in Table 1.

To calculate cumulative *NF* and cumulative input intercept, please look at Chapter 4, "Receiver Design," in W. E. Sabin and E. O. Schoenike, *Single Sideband Systems & Circuits*, 2nd Ed., McGraw Hill, now also "HF Radio System & Circuits."[15]

ARRL laboratory tests[16] have reported about the good performance of the Elecraft K2 receiver with respect to 5-kHz-spacing, two-tone IMD test in comparison with some higher priced commercial equipment. Thereby arises a question: Why? A first answer could be that it has a narrower first IF filter and maybe a better first mixer as well. Most up-conversion, general-coverage receivers for amateurs have all-mode capability and one roofing filter around 70 MHz, and wide enough for FM. A switchable first IF filter to narrow the bandwidth while on CW or SSB is desirable, but that adds to the cost. In addition, you might need to change the whole architecture since such VHF first-IFs are not compatible with narrow band-pass filters.

Maybe a secondary effect must be considered: problems in the area of signal delay to synchronize a conventional noise gate for an effective noise blanker. I do remember, some years ago, someone complaining about less effective blanking action with a Drake R4C receiver after the replacement of the

Table 3—Cauer Band-Pass Filters Design Data

	L1-L4					L2				L3				C2		C3	
Band (MHz)	Core	Turns	Wire Diameter (mm)	µH	C1-C4 Var C (pF)	Core	Turns	Wire Diameter (mm)	µH	Core	Turns	Wire Diameter (mm)	µH	Var Cmax	Fixed C	Var Cmax	Fixed C
28.5	T80-6	34	0.75	6	1.5-5.5 (7mm)	T50-10	10	0.75	0.3	T50-10	7	0.75	0.2	100	56	65	56
24.9	T80-6	36	0.75	6.8	1.5-5.5	T50-10	11	0.75	0.4	T50-10	10	0.75	0.3	100	68	100	—
21.2	T80-6	38	0.75	7.4*	1.5-5.5	T50-10	17	0.75	1	T50-10	16	0.75	0.9	65	—	65	—
18.1	T80-6	39	0.75	8*	1.5-5.5	T50-6	20	0.75	1.8	T50-6	18	0.75	1.5	65	—	65	—
14.2	T94-2	46	0.75	18*	1.5-5.5	T50-6	16	0.75	1.2	T50-6	15	0.75	1.0	100	56	65	47
10.1	T94-2	60	0.5	30*	1.5-5.5	T50-2	23	0.5	3.3	T50-2	22	0.5	2.7	100	—	65	—
7.05	T94-2	64	0.5	34*	3-13(7mm)	T50-2	20	0.75	2	T50-2	16	0.75	1.5	100	280	100	220
3.7	T94-2	64	0.5	34	65(10mm)	T50-6	18	0.5	1.7	T44-10	14	0.5	0.8	—	2200 poly	—	1100 poly
1.83	T94-2	100	0.35	85	100(10mm)	T50-2	25	0.5	3.3	T50-2	18	0.5	1.7	—	3900 poly	—	2300 poly

Notes

-Variable capacitors are small Philips film or Teflon dielectric. To achieve values of 1.5-5.5 pF, connect a fixed capacitor (3-10pF) in parallel with the variable.

-Fixed capacitors are dipped mica (good ceramic, or polyester when indicated). Some dipped-mica capacitor suppliers: RF Parts, Surplus Sale of Nebraska. In some cases where you don't have a standard value available, make a parallel combination to get as close as possible. In any case, check the real filter shape.

-In and out relays +V dc control line must be connected together and to panel band switch for each band with some RF decoupling.

-Use enameled copper wire of closest AWG to the metric size shown; see Table 4.

-Toroidal cores are standard AMIDON at **www.amidon-inductive.com/** or Palomar types at **www.palomar-engineers.com/**.

-Board used: glass epoxy single sided.

Attention: The lower µH values are prone to errors. See also text "practical world paragraph." Fine tuning of L2-C2 and L3-C3 is required if you don't get a well shaped 60-65 dB attenuation response.

*Indicates a large difference from data on the OK1RR Web site.

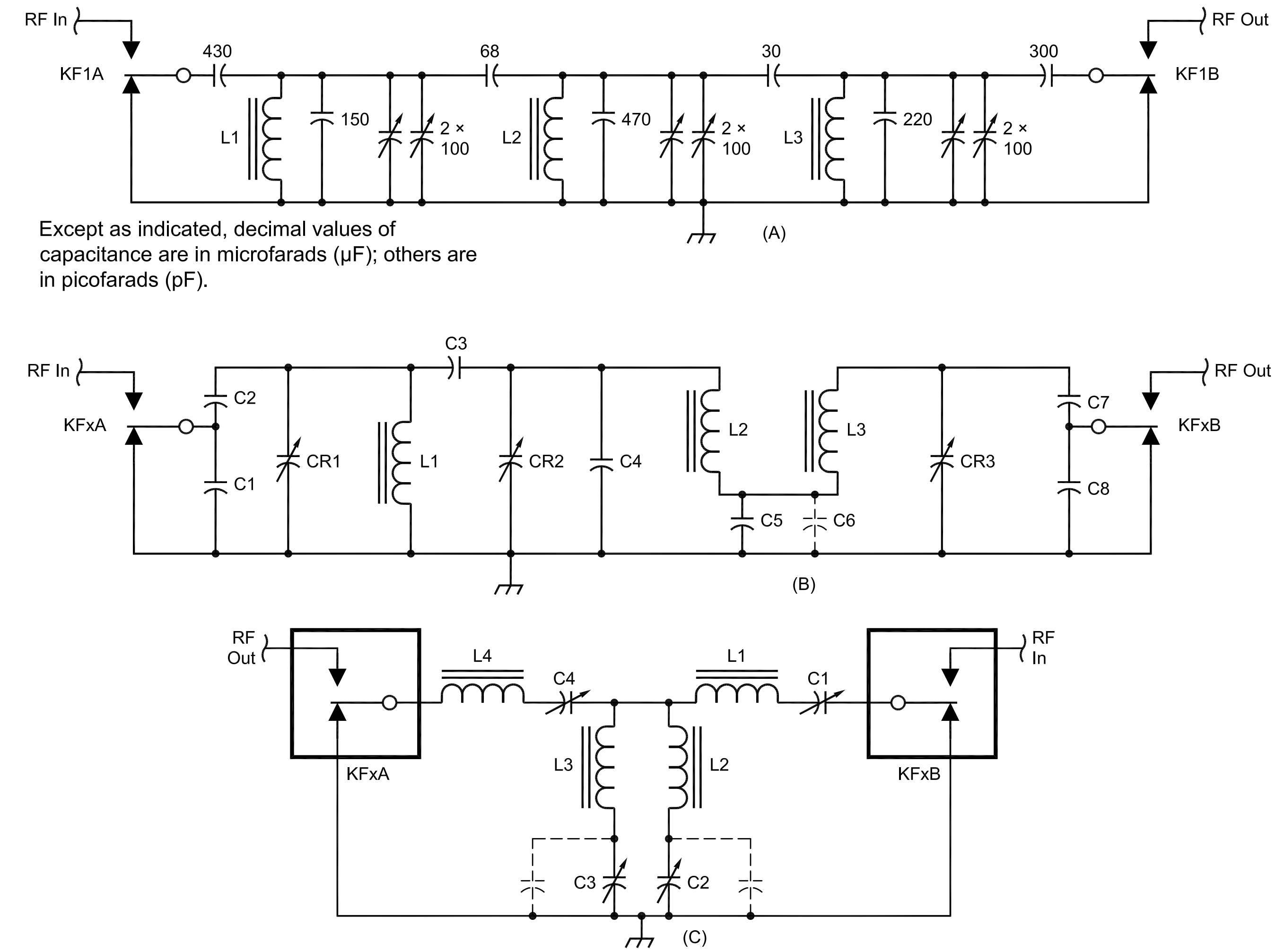


Fig 3—Schematics of the band-pass filters. (A) shows a 1.8-MHz band-pass filter from an old *Handbook.* L1, L2 and L3 are T68-2 powdered-iron toroid cores with 40 turns of 0.5-mm (#24 AWG) enameled wire. Capacitors are dipped mica parts. Variable capacitors are Philips film/Teflon units that are 10-mm in diameter. (B) shows 3.5 to 28 MHz Butterworth filters with in/out capacitive dividers. See Table 2 for each band's design values. Connect the relay 12-V control lines together, and control them with circuit in Fig 1B. Decouple the control lines for RF at both ends. (C) shows 3.5 to 28 MHz Cauer filters. See Table 3 for each band's design values.

first-IF crystal filter with a narrower one.

A well-designed front end with band-pass filters around our band segments is an added bonus to improve our equipment's *IP2*. A preamplifier is not always needed.

Our Front End

Now, our accessory needs to be an external independent front-end unit with its own filtered power supply. It should be easily connected to any transceiver (new or old) with:

1. A variable attenuator from 1-20 dB with a bypass switch;
2. Modular ham-band-only band-pass filters with relay switching (no diodes to avoid IMD), the inputs of unused filters are shorted to ground;
3. Two stages of preamplification;
4. A push-pull, broadband medium quiescent-current amplifier configuration with low IMD and a reasonably low noise figure with some kind of RF feedback.

The variable attenuator and preamplifiers are protected by the band-pass filters, since they are placed after them in the receiver chain.

This is a reinterpretation of a high-level receiver front end, as we see it, adapted for our use. It is a system made with well-known circuits as building blocks. You can modify what you want, since every unit is modular. Improvements are welcome.

The unit must be capable of some switching among different antennas: **ANT 1 RX/TX**, **ANT 2 RX/TX**, **RX-ONLY ANT**. For each antenna selected, the receive-only signal path is always routed through the band-pass filters. See K5AM's article (*QEX*, Nov/Dec 2001, p 40) in which he pointed out different *IP2* performance when measured at the main receiver terminal or at **ANT RX ONLY** input, leaving some hope to the home builder for better performance.

The front end should be useful in

Table 4—Metric versus AWG Wire Sizes

Wire Diam. (mm)	*Equivalent AWG*
0.35	*27*
0.5	*24*
0.6	*22*
0.75	*20*
0.8	*20*

casual DX operating, in single-operator contests and in multi-operator contests with the receiver signal parallel routed to two receivers. These would be a main receiver and a secondary receiver with a second operator who can tune independently. It should be capable of work in low-frequency amateur bands but with modular construction that can be upgraded to cover all HF bands including WARC bands.

When used with full-sized antennas, it should provide benefits as well, since the band-pass filters are designed and aligned with sharp bandwidth and excellent shape factor. We use more space than most embedded band-pass filters. Equipment manufacturers must tradeoff cost, dimensions and the *Q* of components.

I'm thinking now of an Amateur Radio system composed of one antenna with multiband coverage with only one feed line to the rig that covers 10, 15, 20, 40 meters and the WARC bands, like log-periodics. Friends with such antenna systems told me about more IMD problems in their receivers during evening hours because of the high-level signals present in the broadcast bands around 40 meters. The situation is a bit better for those who use monoband antennas and separate antennas with separate feed lines for 40 and 80 meters. Again, this problem seems to be worse in Europe, as pointed out by G3RZP.

Think about radios with receiver general-coverage capability. If the number of band-pass filters is 10, they must be around 3 MHz (30 MHz/10) wide at –3 dB, and of course much more at –60 dB. These so called "sub-octave-width band-pass filters" are a limited form of preselector filtering[17] but they are still helpful. We have tried to select band-pass filters with bandwidth of around 400 kHz, except on 28 MHz because the band we can use is 1.7 MHz wide. The filters are centered at the middle of each amateur band with an acceptable insertion loss. In that way, the improvement in bandwidth we achieve is about 6 to 1 (3 MHz/0.4 MHz) and we believe that everything before the first IF roofing filter is a bit better protected from strong, out-of-band signals.

We have data for all nine amateur bands, WARC included, of two different band-pass filters types. First is the Butterworth response (Fig 3B, Table 2) with three toroidal inductors, based on a study previous published in *QST* by Bill Sabin, WØIYH.[18, 19] Second is

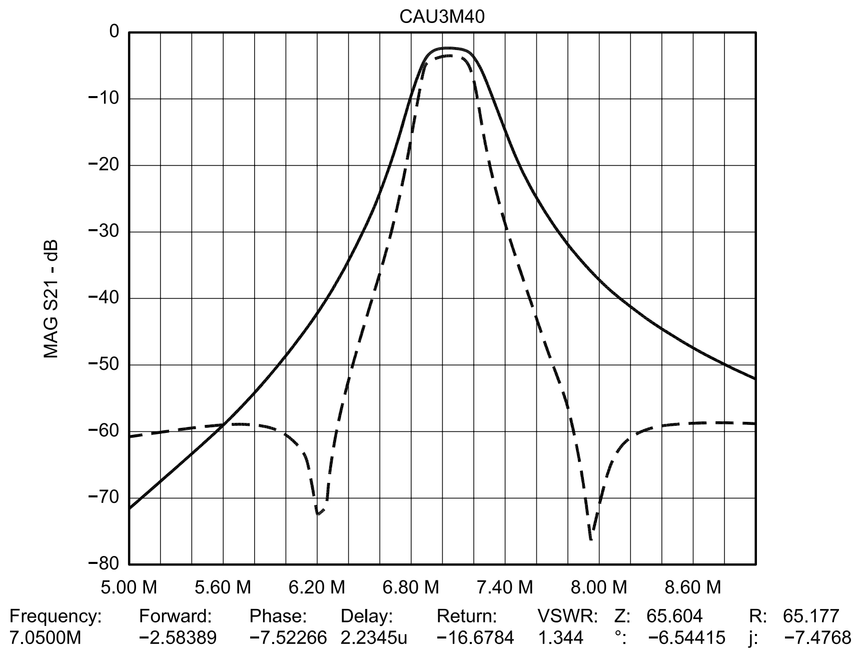


Fig 4—Response curves for the Butterworth and Cauer band-pass filters.

W8JI's Point of View

I would like to report here a message by Tom Rauch, W8JI, that recently appeared from him in the top-band reflector because I think it is a very clever summary:

"This question comes up frequently, and a brief summary might be useful. Magnetic loops can offer improvement in S/N if you have local noise from one primary direction. For sky-wave noise or QRM, they are somewhat useless. The 'shield' is meaningless, except [that] a properly implemented shield can sometimes improve balance of the loop. They are really *not* magnetic antennas, except immediately next to the antenna. At a distance of about 1/10th wavelength or so, a small 'magnetic' loop's response is primarily electric; and at about a half-wave or farther, they are no different than any other antenna type you might use. In a location free of noise or interference coming from a well-defined single direction in the loop's null, they will not improve S/N ratio.

"K9AY loops, flags, pennants and EWE's [sic] all work on the same principles, since they are all small terminated loops. They behave like small two-element vertical arrays, with an internal phasing system. The termination insures each 'vertical' element has equal current; phase is inherent in the design and comes from the horizontal component of the wires. They are primarily useful when noise or QRM is directly off or near the rearward direction.

"These loop antennas (even the single unterminated loop) are all moderately low-impedance antennas, and despite rumors, you *can* use a metallic mast with any of them as long as the mast is isolated from the element, not much taller than the antenna and non-resonant.

"Snake antennas are really just 'random luck' antennas. There isn't any science, reliability, or planning to successful installations. Sometimes you'll find a wire or antenna that helps under some conditions. It might be an antenna for another band (like an 80-meter dipole) that just happens by random chance to work, or it might be a wire strung out on the ground like a snake.

"One thing all these antennas (like most smaller 160-m[eter] receiving antenna systems) have in common is [that] results will vary greatly with each individual application.—*73, Tom, W8JI*" the Cauer response with four toroidal inductors (Fig 3C, Table 3). Our Cauer (elliptical) is a little different from that presented in *QEX* by W3NQN.[20] That version emphasized maximum attenuation in adjacent amateur bands. Ours is in the same family, but it is easier to wind the toroidal cores—on a single layer—since it is already calculated to match 50 Ω.

We found the initial idea and data for this version of Cauer filters from the OK1RR Web site and we have verified them.[21] We tried larger-diameter toroidal cores—T94 instead of T68—as was used by W3NQN in his version.

Our Cauer filters are optimized third-order ellipticals with steeper skirts than traditional Butterworths. We compared ours with three-toroidal-inductor Butterworth versions (Fig 4) and you can see the different shapes yourself with the following procedure.

We thank OK1RR and W3NQN who first mentioned the *ELSIE* program to us in *QEX*. We would like to go further inside it with you and discover together how it is a friendly and powerful tool.

ELSIE Software by WB6BLD for RF Filter Design

For both versions, we have winding data on Amidon toroidal cores. They may be analyzed with *ELSIE*[22] by WB6BLD, James Tonne, of Trinity Software, who now lets you download freely a fully functional student/amateur version from his Web site. I would like to thank James for that—very well done.

You can easily superimpose three different curves, that is, 3-MHz normal filter bandwidth, the Butterworth and the Cauer (both with a sharper 3-dB bandwidth like 0.4 MHz), to see the different behaviors in shape factor. It is also easy to simulate changes in values with instant impact on the curve shown. And you can even print to an HP laser or compatible ink jet printer.

Download *ELSIE* software from WB6BLD's Web site. This demo version is fully working and the main limitation is the upper filter order of seven. You can store up to 15 work files in the same folder. To be able to recall some files for overlay purposes, keep the working files created and saved in the same folder. Move the ones you don't immediately need to another folder.

After you learn how to enter a filter design into *ELSIE*, you get a beautiful display of the filter's response. With relatively wide frequency spans, you see the filter's behavior away from the passband. Change the frequency span to check more precisely the –6-dB bandwidth.

It is easier to do this at the computer than to explain the process here. The learning curve with this software is short, and it is as WB6BLD states: "After a bit, you will not need any manual."

You have a lot more to see and discover with this software and you can print the output. A wonderful option is the capability to overlay more curves when you have already created all the files you want to compare. You can look at not only the transmission curve but also the return loss curve. The capability of fine-tuning the component values with real-time display of effects is amazing.

With all L-C data for the band of your interest, you need some help before starting to wind the toroidal inductors. You may need to go from one core size to another with some trade-off in *Q*. I have recently found a very nice and useful *Windows* program, free, made by some Italian amateurs, IK2JSB and friends. It lets you calculate simply the number of windings on most Amidon toroidal cores (you just select the type) from inductance value input. I remember a similar calculator in the well-known VE3ERP DOS suite of programs. Download the program from IK2JSB's site.[23]

When you run the program, choose option AMIDON. Some label descriptions about color codes for Amidon mixes are in Italian, but it is very simple and complete.

Practical Considerations and Alignment Tips

Of course, you need to test and carefully tune each filter with small variable capacitors in parallel with the

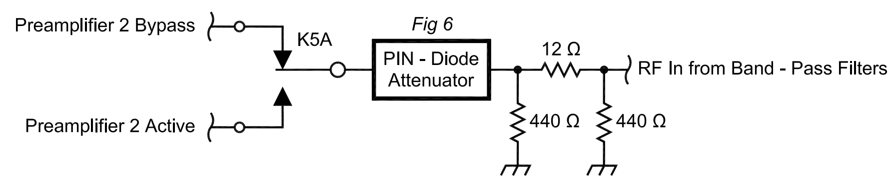


Fig 5—Schematic of the preamplifier switching logic and matching.

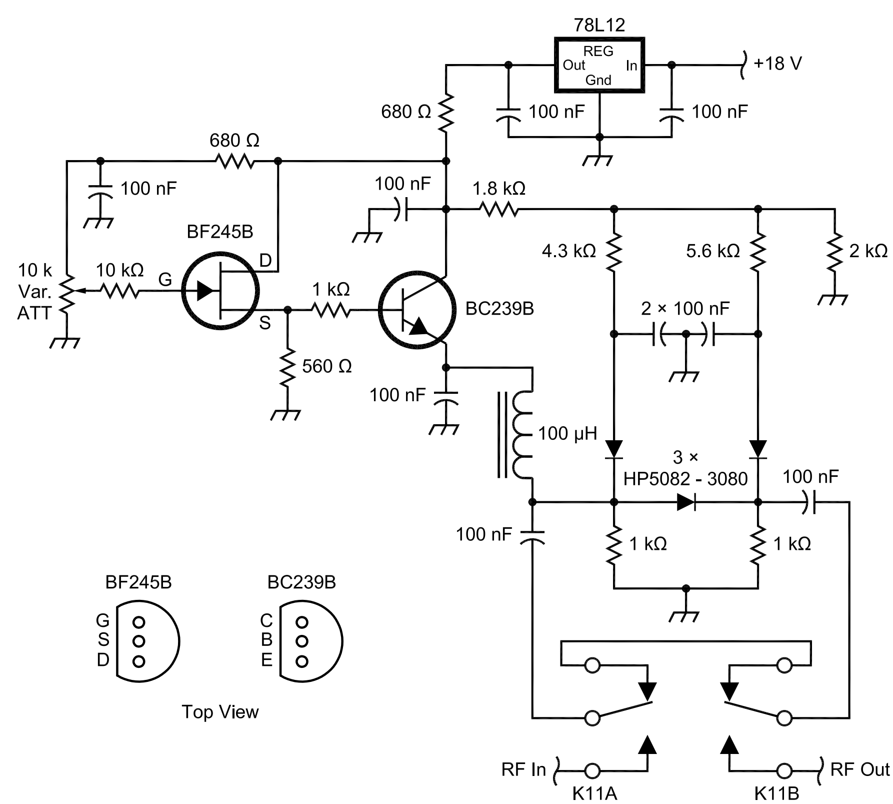


Fig 6—Schematic of the PIN-diode attenuator circuit. HP5082-3081 work better than those shown (see text). The board layout allows for an NAIS TX-2 12-V relay.

fixed values so the capacitances match the design data. We strongly suggest you use high-quality variable capacitors for easier alignment and the best filter shape. We used small Philips, Teflon-film, 300-V variable capacitors. For fixed capacitors, you can use ceramic disks; but we preferred the dipped-mica types.

Relaxed values of *Q* are realistic—maybe 120 for *L*s, 900 for *C*s. We suggest you mount the series elements C1-L1 and L4-C4 first and check with a signal generator the insertion loss. If the attenuation is more than 10 dB, even with adjustment of the variable capacitors, you need to change the coil by one turn and try again. Then mount L2-C2 and L3-C3. Some fine-tuning of these values may be necessary for better shape factor and bandwidth settings versus attenuation. Check the whole filter shape and insertion loss. If you do not get top results in some higher bands, accept them or try a slightly wider bandwidth: 500 kHz to 1 MHz is normal. Again, some changes in L2-C2, L3-C3 may be necessary to trade off for better attenuation. In the lower microhenry values, the errors tend to be greater. So a core made from the winding by data can be quite far from the proper value. In the lower bands, the results we got are much closer to predictions.

We must report that in two or three bands we got some practical data that were a lot different from those reported by OK1RR; otherwise it couldn't work in practice. We use a small π attenuator to always get a proper 50-Ω load.

All RF switching is done by simple small SPDT relays of reasonably good RF properties. Better ones have thick, gold-plated contacts; avoid palladium-alloy contacts. Precision inductance and capacitance meters might help. To get an idea about real performance, a network analyzer is the preferred instrument. Since most of us simply cannot afford one, I suggest you read the *QEX* article about RF network analyzers by Steve Hageman.[24]

Cauer filters seemed to work better when assembled on single-sided epoxy board for a bit less insertion loss. Butterworth seemed to work well also on double-sided copper epoxy board and they are easier to align.

Variable Attenuator

Our attenuator has a variable 1-20 dB range with a panel-mounted resistive control. It is built around traditional three-PIN-diode circuit (see Fig 6). It can easily be modified for automatic AGC for a receiver homebrew project. The most important thing is to select PIN diodes with minority carrier lifetimes longer than 1µs to improve IMD characteristics at low frequencies, as originally stated in HP Application Note 936 (now Agilent).

Suitable diodes in glass packages are the HP 5082-3080; or better, the 5082-3081 or MA47600. SMD versions are HSMP 3810; HSMP 3814 is a dual, common-cathode device. I suggest you read the very interesting HP/AGILENT Application Note 1048, "A Low-Cost Surface Mount PIN-Diode π Attenuator," with schematics for a three-diode, 5082-3081 design with 15-V dc attenuation control. There is a discussion about an even more symmetrical configuration with four diodes and a test report over the frequency range of 300 kHz to 3 GHz. At 10 MHz, a two-tone, third-order intermodulation distortion input intercept point over +30 dBm for attenuation settings in the range 10-20 dB is claimed. *IP3* is better with less attenuation.

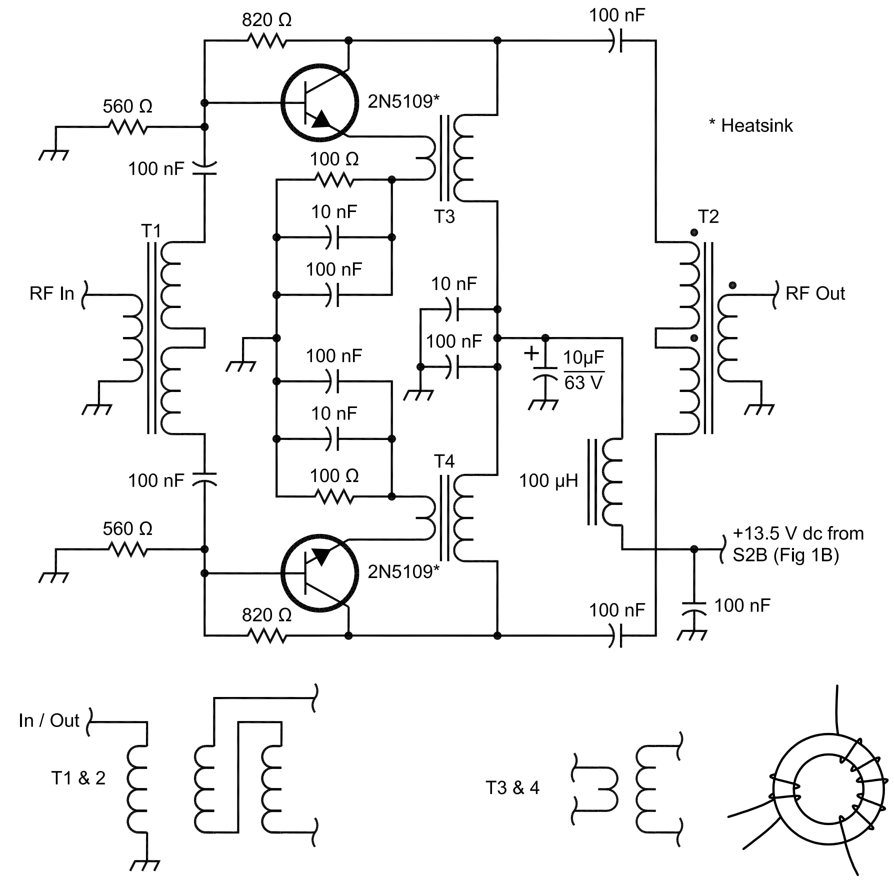


Fig 7—Schematic of the preamplifier circuit used in two stages. T1 and T2 are each 12 trifilar (twisted) turns of 0.35-mm (AWG #27) enameled wire on a FT50-43 ferrite toroid core. T3 and T4 are each a 9-turn primary and a 2-turn secondary (0.35-mm AWG #27 enameled wire) on an FT37-43.

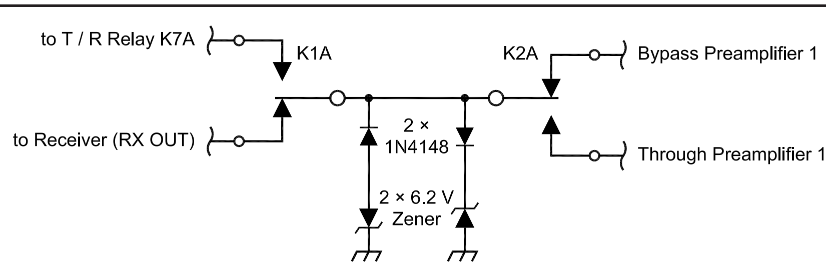


Fig 8— Schematic of overload-protection and switching circuits.

Push-Pull Broadband Preamplifiers

We have chosen medium-current 2N5109 RF transistors in a push-pull, common-emitter broadband configuration (see Fig 7). It is a low-noise version with transformer collector-emitter feedback and all home-built transformers. The basic design is from Ulrich Rohde, KA2WEU's article presented in *ham radio* magazine[27] (now available on CDROM from ARRL) with our own printed-circuit layout. In that article, there is also an IMD/dynamic range graphic as Fig 13, p 17, *ham radio*, Oct 1976. We have raised the quiescent current a bit (about 15 mA more) to 55-60 mA for each 2N5109, with our resistance value of 820 Ω from collector to base, for even less IMD. We measured *OIP3* at around +41. Maybe we will achieve a slightly higher noise figure than the 2 dB indicated in that article. The rugged 2N5109s in TO-39 cases, with a rated power dissipation of 2.5 W, have performed very well with small heat sinks.

This preamplifier unit is simple, stable and reliable. We have had one in use since 1995 in I4FAF's homebrew HF transceiver front end for amateur bands only. There is a front-panel rig photo on our Web site; **www.qsl.net/ik4auy**. We have had over 10,000 QSOs without any problem. Our rig architecture is triple-conversion with IFs at 10.7 MHz, 9 MHz with PBT and 455 kHz. We selected a Mini-Circuits high-level passive doubly balanced mixer—a TAK-3H, +17-dBm LO-power unit with improved IMD (Level 17S) following our inside Butterworth band-pass filters.

We have two preamplifier units, so we can get 0/+10/+20 dB of gain if needed. The net gain depends on band-by-band filter insertion losses. Most of the time, I use only the first preamplifier stage. The gain per stage is around 12 dB. The sequence of our chain is shown in Fig 1. The filters are always in line in the receiver channel; then the variable attenuator if needed; then the amplifiers, if needed.

The unit includes a well-filtered power supply, with a small 1-A ac RF filter, capable of delivering about 240 mA for the preamplifiers, switching relays and the variable attenuator (see Fig 9). A single metal LM317 is more than enough for very-long-time operations, mounted on the rear aluminum case of the unit with a mica insulator and silicon grease.

One more odd thing: **ANT 1/ANT 2** and **TX/RX** channels input and output switching is obtained with three SPDT RF relays. Since we have a transmitter with maximum output of 120 W, we made our own SPDT relays with two SPST RF reed relays (see Fig 10), avail-

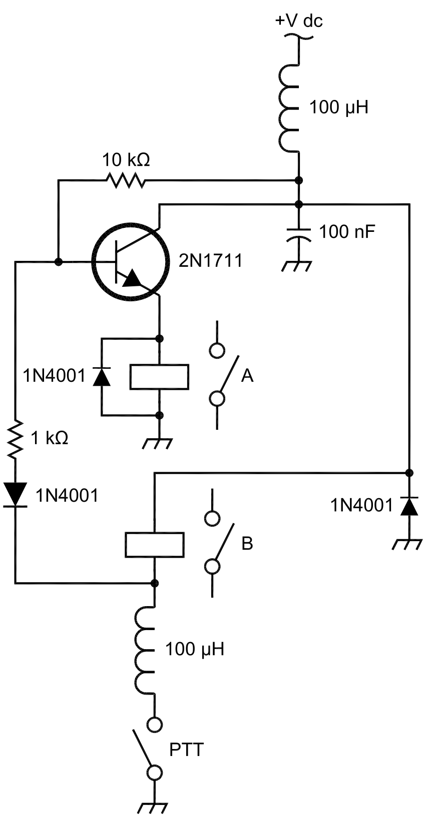
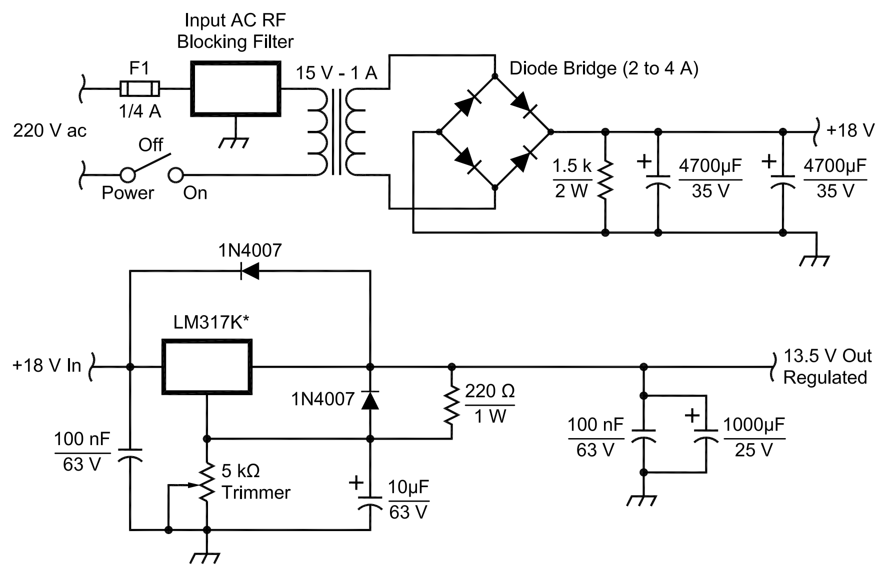


Fig 10—Schematic of the PTT activated RF-power relay constructed from two reed relays.



* LM317K (Metallic Case) is Mounted on Back Chassis with some Silicone Grease and a Mica Insulator.

Fig 9— Schematic of the regulated power supply. Although a 220-V ac input is shown, only the transformer need change for 120-V operation.

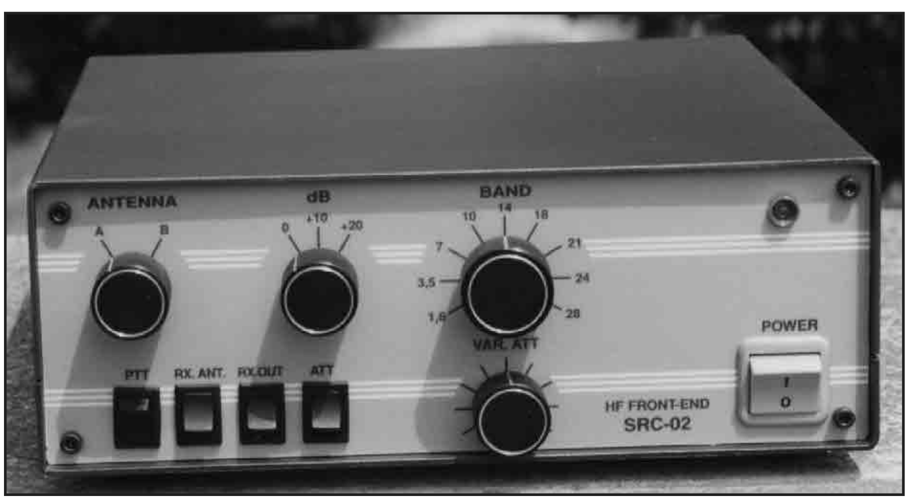


Fig 11—A front view of the project.

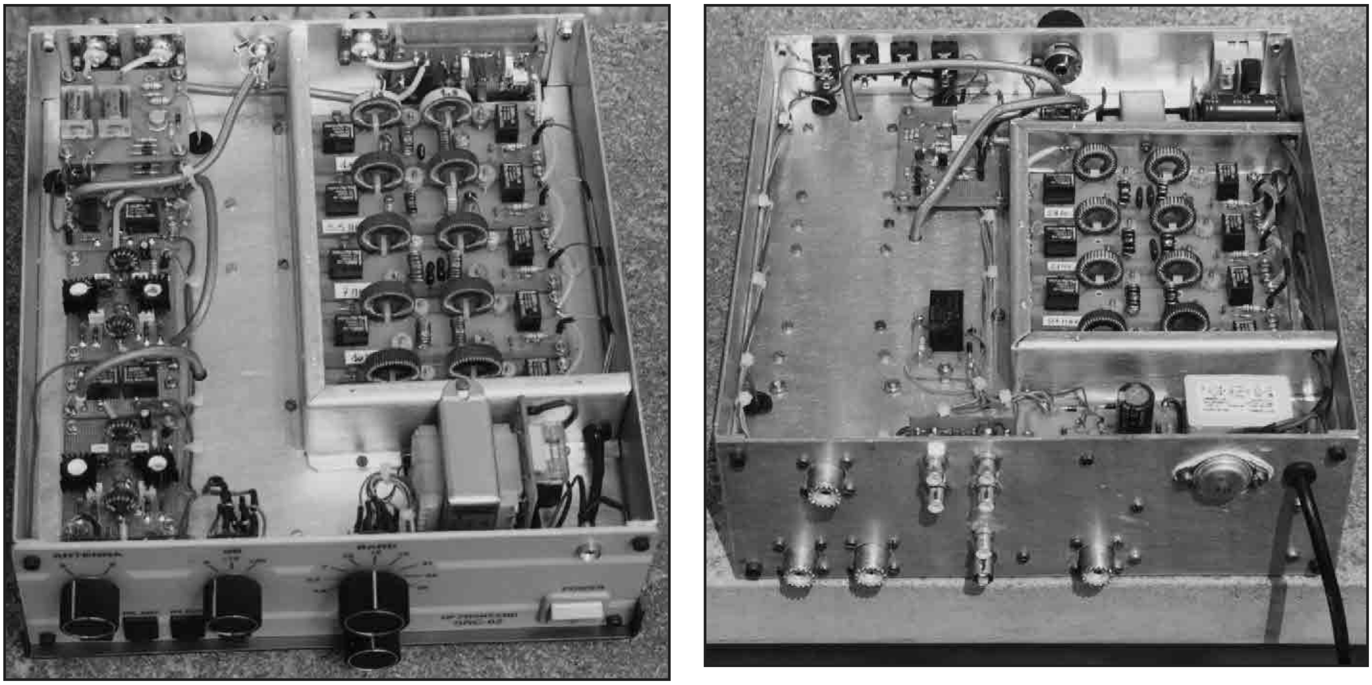


Fig 12—(A) front-top and (B) rear-bottom views of the project.

able at much lower cost than a good SPDT RF relay. They work very well and fast enough. For power levels up to 250 W, you can use any relay with contacts rated at more than 2 A and suitable for RF use. You might use surplus coaxial RF relays or a new Tohtsu CX120A (rated at 60 dB of isolation at 50 MHz in the RF PARTS catalog).

We used a 500-V dip-mica capacitor from the input of the last output relay to ground, around 30 pF to get an SWR very close to 1:1. Remember to check also, with a sensitive wattmeter to a dummy load, to avoid any dangerous RF coupling from the transmit antenna to the receiver antenna while transmitting.

Acknowledgement

This article is dedicated to my father, I4FAF, because it wouldn't have been possible without his valuable material work and many hours of bench tests.

Notes

[1]J. Devoldere, ON4UN, *Low-Band Dxing*, third edition (Newington, Connecticut: ARRL).

[2]J. Briggs, K1ZM, *Dxing on the EDGE*, (Newington: ARRL).

[3]E. Cunningham, K6SE, "Flags, Pennants and Other Ground Independent Low band receiving Antennas," *QST*, Jul 2000, page 36.

[4]G. Breed, K9AY, "The K9AY Terminated Loop-A Compact, Directional Receiving Antenna," *QST*, Sep 1997, p 43. Hum problems when switching the K9AY loops (in *QST* Technical Correspondence).

[5]F. Koontz, WA2WVL, "Is this EWE for you?" *QST*, Feb 1995, p 31; feedback in Apr 1995, p 75.

[6]F. Koontz, WA2WVL, "More EWE for you," *QST*, Jan 1996, p 32.

[7]Antenna Compendium Vol. 5 (Newington: ARRL), "Ewe 'Four' Me," by J. Smith, VK9NS, p 32.

[8]Visit **www.contesting.com** to get access to Top Band reflector.

[9]**www.w8ji.com**, Tom Rauch's personal Web site for receiving-antenna theory and ideas (see also four phased receiving only short vertical under "small-vertical arrays").

[10]**www.angelfire.com/md/k3ky/page37.html** K3KY Web-link collections about low-band receiving antennas (among them K6SE pennant, W7IUV rotatable pennant (**www.qsl.net/w7iuv/**), WA1ION variable termination flag/pennant and more in **www.qsl.net/wa1ion/**).

[11]**www.aytechnologies.com** for K9AY loop info and distributed by ARRAY Solutions, **www.arraysolutions.com**, or Wellbrook Web site in UK, **www.wellbrook.uk.com/K9AY.html** also **www.hard-core-dx.com/nordicdx/antenna/index.html**, antennas, choose loops.

[12]R. Brown, NM7M, "On the SSW Path and 160-meter Propagation," *QEX,* Nov/Dec 2000, pp 3-9; for top-band propagation characteristics.

[13]P. Marino, IT9ZGY, "Impariamo a capire gli spot WWV," *Radio Rivista* (ARI), Jun 2001, p 47.

[14]E. Barbieri, I2BGL, "Il rumore atmosferico e l'ascolto ottimale dei segnali radio," part 2, *Radio Rivista* (ARI) Jun 2001, p 34.

[15]W. Sabin and E. Schoenike, *HF Radio System & Circuits* (Newington: ARRL, Order No. 7253).

[16]*QST,* Jul 2001, p 80: "Table 2—Dynamic Range Measurements at 5-kHz Spacing for Several Current HF Transceivers,"

[17]W. Sabin and E. Schoenike, *Single Sideband Systems and Circuits*, second edition (New York: McGraw Hill) or the new edition in Note 15: W. D. Hart, Chapter 9 "Preselectors and Postselectors," p 359. "Sub-octave Bandpass Filter," p 357, summarizes the properties of preselectors in reducing: IM, cross-modulation, reciprocal mixing, desensitization, spurious and image responses and circuit overload damage.

[18]W. E. Sabin, WØIYH, "Design Narrow Band-Pass Filters with a *BASIC* Program," *QST,* May 1983, pp 23 to 29.

[19]F. Cherubini, IØZV, "Filtri preselettori per HF," *Radio Rivista* (ARI) Sep 1989, p 38, from 80 meters to 10 meters. A fine 160-meter band-pass filter from the *1980 ARRL Handbook*, p 8-43. There are more tips about this front-end from Rockwell-Collins KMW-380 Service Manual.

[20]E. Wetherhold, W3NQN, "Receiver Band-Pass Filter Having Maximum Attenuation in Adjacent Bands," *QEX* July/Aug 1999, pp 27-33.

[21]For the slightly different Cauer band-pass filter to which we refer in this article, see OK1RR's Web site under "Tech Stuff, More front-end selectivity for your receiver" at **www.qsl.net/ok1rr/.** There, he also makes a comparative test with traditional well-known W3LPL filters, originally presented by W3LPL, now at his Web site.

[22]WB6BLD *ELSIE* software at **www.qsl.net/wb6bld/**.

[23]IK2JSB has an Amidon toroidal coils calculator at **space.tin.it/computer/grgandin/**. Carefully follow the installation procedure described in the text.

[24]S. Hageman, "Build a 250-MHz Network Analyzer," *QEX*, Mar/Apr 2002, pp 3-10.

[25]The idea to use an AD8307 as a logarithmic detector to show the filter curve on an oscilloscope occurred after the reading of W. Hayward, W7ZOI, and R. Larkin,

W7PUA, *QST,* Jun 2001, p 38 and W. Schneider, DJ8ES, "Logarithmic Amplifier up to 500 MHz with AD8307," in *VHF Communications* Feb 2000, pp 119-124.

26A very interesting application article "10.7 MHz, 120 dB Logarithmic Amplifier," by Barrie Gilbert, Eamon Nash, from *Microwaves & RF* for Mar 1998 is on-line in the Analog Devices Web site under Logarithmic Amplifiers technical articles. With an AD603 logarithmic variable amplifier in front of an AD8307, this one is a very fine upgrade for any oscilloscope display panadaptor project like that in *QEX*, May/June 1999, "A calibrated Panoramic Adapter," by Bob Dildine, W6FSH, pp 9-22, that uses a conversion architecture at 10.7 MHz IF.

27U. L. Rohde, DJ2LR (KA2WEU), "Optimum Design for High-Frequency Communications Receivers," Oct 1976 *ham radio*, p 10, includes a broadband HF preamplifier built around two CATV transistors on p 18, version C. *ham radio* is available on CD ROM from ARRL.

IK4AUY, Sergio*: Amateur Radio license at 16, active since 1980 as his father I4FAF's second operator. He got his own call at 18. He is now 38, and still enjoys DX and Contests. Sergio holds a degree in economics and commerce from Bologna University and works in a local bank as credit analyst. He started young and read a lot of good amateur-related books and magazines in his spare time.*

He holds S79AU from the Seychelles Islands and Amateur Extra US license AC7PC (my Italian address only is good for mail), 5BDXCC and is on the DXCC Honor Roll as well. In some contests he uses special call IR4B, and he is one of Marconi Memorial Station (IY4FGM) official operators from his summer home where he started, young, his first wireless radio experiments, now mausoleum and museum, in Pontecchio Marconi (Bologna). Sergio and other IY4FGM operators work in the organizational team of the annual Italian HF DX Convention at the end of September.

I4FAF, Romano, *is an old timer in Amateur Radio; his age is 69. He started young at 18 by repairing radios in the navy. He has been an ARRL member since 1980.He is on the DXCC Honor Roll. I4FAF mainly enjoys home-building since retirement, having built HF transceivers, power amplifiers and test equipment.*

Both Romano and Sergio are members of ARI, the Italian national association, and the ARRL. They have published four technical articles in Radio Rivista *since 2000. They have a Web site at* ***www.qsl.net/ik4auy****.*

RF

By Zack Lau, W1VT

An Optimized 6-Meter Yagi

This 6-meter Yagi was first designed for QRP portable use—I wanted a simple beam that could be quickly put up and taken down, while being easy to transport in a compact sedan. I decided on a permanently assembled 2×4 foot center section and four thin 3-foot tubes. This easily fits on the back seat of most cars. Assembly is just a matter of sliding in the tubes, adjusting the precise lengths, and tightening down the element clamps. I published a computer model of the design in the 19th Eastern VHF/UHF Society *Proceedings*.

The dimensions shown are different from what I used 10 years ago—I decided to tweak the dimensions for a better front to back ratio in the DX portion of the amateur band. While the gain is 0.5 dB less than the original, the F/B is over 19 dB, much better than the typical 8 to 12 dB one associates with two-element Yagis. The free space gain is still a respectable 6.5 dBi according to Roy Lewallen's *EZNEC* and Brian Beasley's *YA*. Fig 1 shows a *YA* file—the extra non-resonant element is a limitation of the program. It expects at least three elements, so I detuned it and placed it 833 feet away from the other elements, so its effect is negligible. *YA* was once bundled with the *ARRL Antenna Book*. Fig 2 shows the *YW* file—*YW* is Windows program bundled with the current *Antenna Book*. The *YA* antenna element lengths are a little shorter—Brian calibrated his program to match *NEC*. Both programs list half element lengths—assuming symmetry allows programs to work faster and handle more antenna elements. The director element in the model is 0.1-inch shorter than the actual hardware—to account for the shortening effect that occurs when an element is electrically attached to a conductive boom. The driven element, being attached with an insulating plate, does not need this correction factor. Figs 3 through 6 show the dimensions of the various parts.

A Moxon rectangle could be used for even more F/B—at the expense of

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```
50MHz yagi
50.100 50.150 50.250 50.200 MHz
3 elements, inches
0.500 0.375
0.000 24.000 20.000
10000.000 24.000 35.400
10019.500 23.750 31.500
```

Fig 1—*YA* file

```
206-02H.YW, 2-ele., 2' boom, 6.56 dBi midband gain
50.0 50.125 50.25 MHz
3 elements, inches
0.500 0.375
0.000 23.750 20.000
10000.000 23.750 35.900
10019.250 23.750 31.900
Match frequency: 50.125 MHz
Driven-element tip: 35.9 inches
Cable Z0: 25.0 ohms
Original file name: C:\ANTBK19\YAGIS\206-02H.YW
```

Fig 2—*YW* file

a more complicated design that often requires one to bend aluminum tubing. It folds back the elements so the tips are closer to each other to optimize the coupling for best F/B.[1] One source of pre-bent aluminum is old lawn chairs—Dick Stroud converted one into a 6-meter Squalo.[2] Instead, I optimized the element diameter to optimize the coupling. On 6 meters, the electrically optimum thickness also works quite well mechanically.

Frugal amateurs will be delighted to discover that the lengths are also optimized for six-foot tubing stock. Cutting a driven element tip from a six-foot length leaves enough stock for a director tip. Two six-foot lengths of tubing can be used to make the center sections of the beam—you will have a four-foot section left over. It may be used for making another 6-meter beam. If you want to stack Yagis, I suggest reading the online tutorial by Ian White, G3SEK, at **www.ifwtech.co.ukk/g3sek/stacking/stacking2.htm**.

The feed-point impedance is very close to 25 Ω, resistive. This is preferable to lower impedances, which typically result in designs with greater losses. You wouldn't want a design with worse F/B and greater theoretical gain, if the practical implementation resulted in a net gain that was no higher. I chose to split the element and feed it with a λ/4 matching section that doubles as a choke balun. The optimum coax impedance is √(25 Ω × 50 Ω), or 35 Ω. RG-83 coax is ideal—if you can find this specialty 35-Ω coax. It may be possible to obtain it from the Wireman.[3] This cable is occasionally produced by Times Microwave Systems; there isn't much demand for this oddball impedance. A cheaper design is shown in Fig 5—you can parallel two 75-Ω cables to come pretty close to 35 Ω. I've also had success with this method, as shown in Fig 7.

You may wish to account for the pigtail leads at either end—the velocity factor is higher when the dielectric is mostly air. The calculation requires some algebra. If *L1* and *L3* are the pigtails and *L2* is the coax cable:

[1]Notes appear on page 60.

$$L_{total} = \frac{984\,\text{ft MHz}}{4f} = \frac{984}{4 \times 50.1\,\text{MHz}} = 4.91\,\text{ft} \qquad \text{(Eq 1)}$$

$$L_{total} = \frac{L1}{Vf1} + \frac{L2}{Vf2} + \frac{L3}{Vf3}$$

$$4.91\,\text{ft} = \frac{0.75"}{0.95\left(\frac{12"}{\text{ft}}\right)} + \frac{L2}{0.78} + \frac{1"}{0.95\left(\frac{12"}{\text{ft}}\right)}$$

$$= 0.066\,\text{ft} + \frac{L2}{0.78} + 0.088\,\text{ft}$$

$$4.76\,\text{ft} = \frac{L2}{0.78}$$

$$L2 = 3.71\,\text{ft} \qquad \text{(Eq 2)}$$

I used clamps with wing nuts to easily adjust the exact element lengths. Ideally, you would adjust the director lengths for the maximum F/B—while listening to a convenient beacon. There are many beacons between 50.0 and 50.1 MHz. G3USF keeps a list of 6-meter beacons at **www.keele.ac.uk/depts/por/50.htm**.

Construction

I recommend reading my Jan/Feb 1998 "RF" column that describes a 4-element 6-meter beam. It has many photographs. It describes in detail how to machine your own custom 1/2-inch tubing clamps that can be finger tightened. Ordinary stainless-steel hose clamps could also be used. I used a 21.75-inch length of 1-inch square tubing for the boom. The tubing wall thickness is 0.080 inches. The director is clamped to the boom with a bent

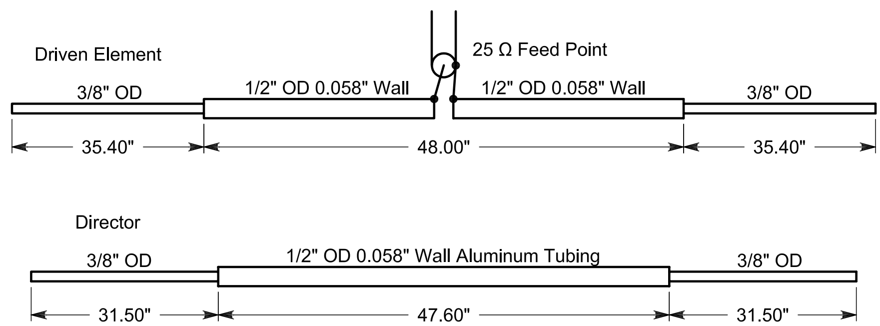

Fig 3—6061-T6 aluminum element dimensions for the 6-meter beam.

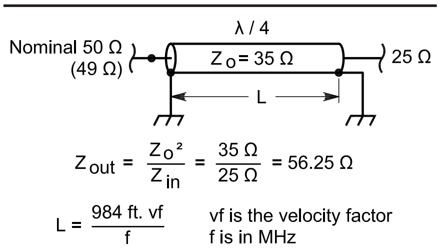

Fig 4—25-Ω to 50-Ω transformer using an electrical λ/4 of 35-Ω cable.

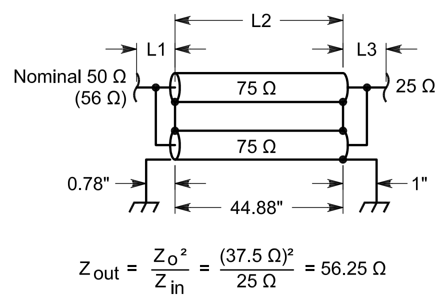

Fig 5—25-Ω to 50-Ω transformer using 75-Ω cable.

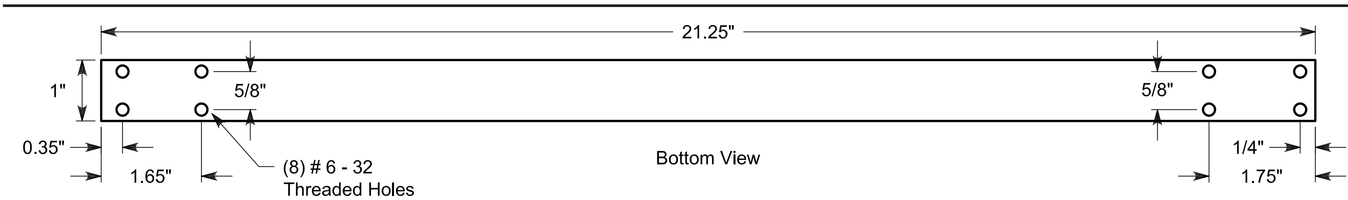

Fig 6—Bottom view of the boom. The top is nearly identical, except that the holes clear #6-32 screws. Use a #36 bit to drill the tapped holes and a #28 bit to drill the holes on top.

aluminum strap. I put a pair of 0.06×0.5×1.0-inch aluminum plates between the screw heads and strap to keep the strap from distorting. The plates fit snugly against the **U** of the strap, as shown in Fig 8.

The driven element is attached to the boom with four aluminum straps screwed onto a 1/4-inch thick Lexan plate. I used #6-32×1 1/4-inch long stainless-steel screws to hold down the plate and straps. These screws are just the right size if you tap the bottom of the aluminum boom. You need longer screws if you intend to use nuts and lock washers. I've not had any trouble with the aluminum threads stripping, but this could be a problem with thin-wall tubing. This insulated element-mounting technique could also be used with the director, but the length of the director should be shortened by 0.10 inches.

Don't forget to drill mounting holes on the side of the boom for the mast clamp: Two holes for a **U** bolt and saddle work fine for this little antenna. Drill them near the center of the boom—the exact size and spacing depend on your choice of **U**-bolt. I've had

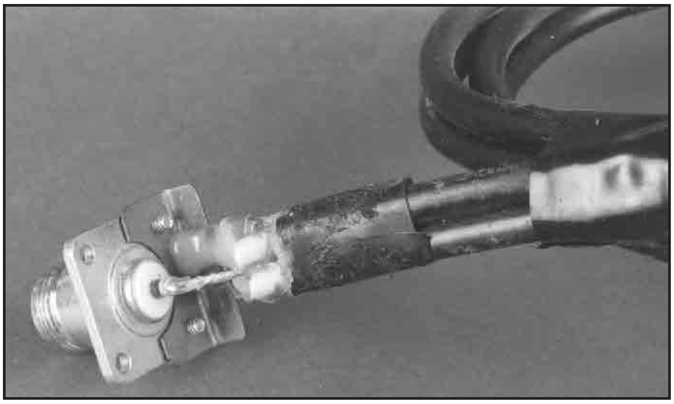

Fig 7—An N connector attached to a pair of 75-Ω cables in parallel.

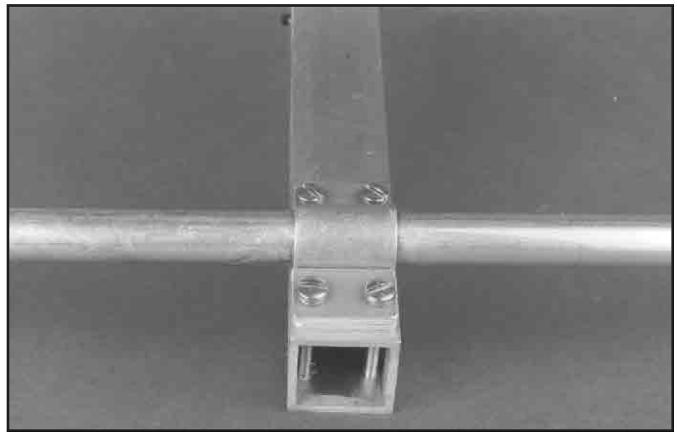

Fig 8—Strapping the director to the square boom.

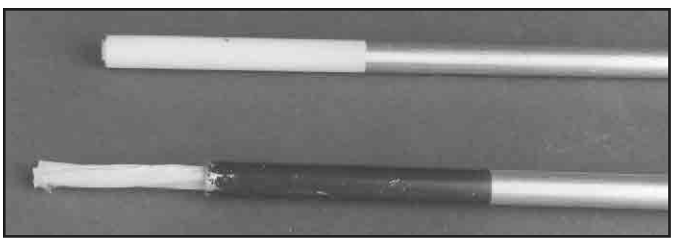

Fig 9—The element tips are painted for color-coding. The rope used to damp vibrations is partially pulled out of the lower element.

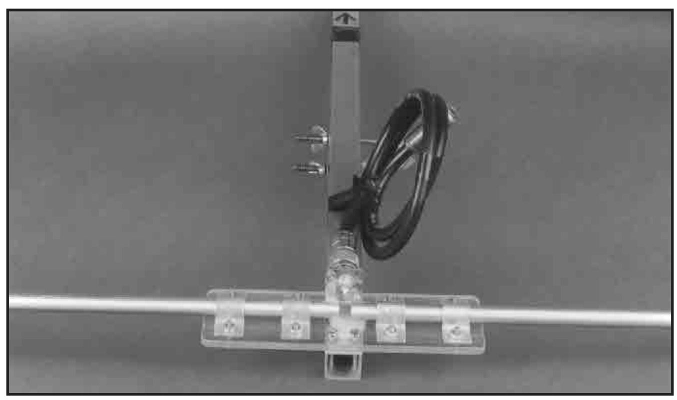

Fig 10(right) —2-turn balun made out of coiled RG-83.

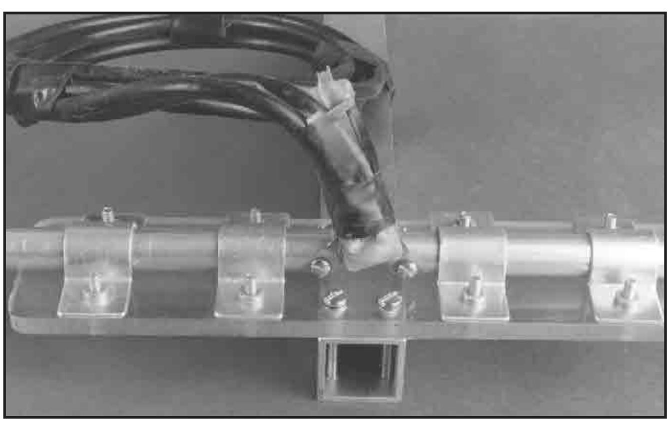

Fig 11—Terminals are swaged to a G-10 insulator and brass strips.

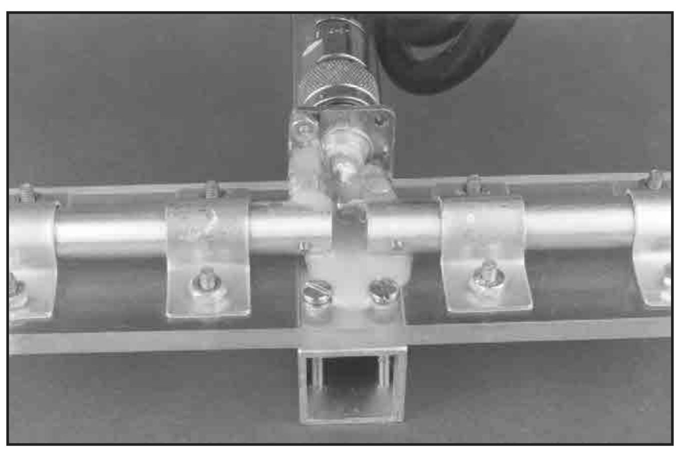

Fig 12—Dow 3140 RTV conformal coating is used to protect the feed-point connection.

good luck with stainless U-bolts with 1/4-20 threads. Instead of the usual hex nuts, I use stainless-steel wing nuts.

The ends of the 1/2-inch aluminum tubing are slotted with a hacksaw or band saw; this allows the clamp to easily compress it against the 3/8-inch tubing. The 1/2-inch tubing should have a wall thickness of 0.058 inches. This size telescopes nicely with 3/8-inch tubing. I cut the element tips to 38 and 34 inches—there will be about 2 1/2 inches of overlap between the tubes. I painted the element tips for color-coding, as shown in Fig 9. It is a good idea to stuff some rope in the element tips—this damps out vibrations and prevents the antenna from whistling in the wind. I used scrap polypropylene cord from electrical cable.

Standard N connectors are easily installed on RG-83 coax. I mounted an N female on one side and an N-male on the other. The coax is coiled into two turns with an inside diameter of 3.75 inches, as shown in Fig 10. It is taped to the boom with electrical tape. Making a choke balun out of two 75-Ω RG-59 coax cables is a little more difficult. If possible, I'd look for RG-59 coax with stranded copper braid and center conductor. I used Belden 9259. Copper is easy to solder—don't make the mistake of trying to solder aluminum with techniques designed for tinned copper. A stranded center conductor is less likely to break. It helps even more to add strain relief to the soldered connections. At the antenna feed point, I swaged a pair of terminal posts to a 5/8×2 1/8-inch piece of unetched circuit board, as shown in Fig 11. The terminal posts also riveted two thin brass strips to the board—the strips are screwed to the elements with #4-40 hardware. Strain relief is obtained by taping the coax cables to the board—black electrical tape works fine. I made a similar strain relief at the other end—a brass strip attached to an N connector also clamps around the shield braids for good electrical and mechanical contact. Just like the RG-83 coax, the parallel RG-59 pair is coiled into two turns, for totaling four coils of coax.

For protection against corrosion, I coated the exposed feed point and braided coax shields with Corning 3140 RTV, as shown in Fig 12. Since the material is transparent, it is easy to see that the connections are still good after a decade of intermittent use.

I included an indicator arrow made of black electrical tape, so it is easier to remember which way to point the antenna.

Notes

[1]L. B. Cebik has a 10-meter design on his Web site: **www.cebik.com/mox.html**.

[2]R. Stroud, W9SR, "Six Meters from Your Easy Chair," *QST*, Jan 2002, pp 33-34.

[3]The WireMan Inc, 261 Pittman Rd, Landrum, SC 29356; tel 800-727-WIRE, Orders 864-895-4195, fax 864-895-5811; **www.thewireman.com/**.

Letters to the Editor

New Faces for Old Meters (*Tech Notes*, Nov/Dec 2002)

I read with interest the article by Tom Cefalo, W1EX, on generating new meter faces and indeed doing this is important to making projects look good. However, I coughed when he wrote he was using *AutoCAD*. For those not familiar with that program, it is one of the premier three-dimensional drawing/drafting packages. The rub is the price: around $3400 at last check. Even *AutoCAD LT* goes for around $600 before adding the symbols package.

A better choice for ham use would be *AutoSketch*, a more basic two-dimensional drawing package at about $100. Perhaps there are other programs out there that would work for this and other drawing and drafting projects [and that would] not drain the wallet. Suggestions, please, for those of us who do not have access to such powerful and expensive software.—*Tom Cook, WA2BPE, 4375 Bellinger Hollow Rd, Corning, NY 14830;* **wa2bpe@infoblvd.net**

A Software-Defined Radio for the Masses, Part 3 (Nov/Dec 2002)

I am just loving Gerald Youngblood's articles, and how much better *QEX* looks to me now than it did several years ago. Congratulations to everyone involved there. One of the most interesting things I have seen in years is the Tayloe design. A "subset" of this was familiar to many in the early days of tinkering with DSP cards when you multiplied every other sample of an incoming signal by –1 and this "inverted" the sideband.

I feel a truly serious drawback to Gerald's articles is the fact that he does not do the "nitty-gritty" work himself and relies on Intel Primitives. This may, in fact, be the fastest code around, but it is also the most useless for our purposes, in my opinion, since as an experimenter (1) I can get at most a 30-day evaluation license without paying $199, and (2) I cannot download the files from you, do work on enhancing the code, and then redistributing it without purchasing the license.

Another design problem is using 4096-bin FFTs for everything and every application. The group delay through the filters is 2048 [sample times] unless he is careful in the design, and that seems too much for SSB!—*Bob McGwier, N4HY, 64 Brooktree Rd, East Windsor, NJ 08520-2438;* **rwmcgwier@comcast.net**

Doug,

Yes, the 4096-bin FFT creates a 46-ms group delay. From on-the-air experience, it works just fine for my taste and is easy to tune. I use the mouse wheel as the tuning knob and can set the tuning step size. One could easily change my example code to a smaller FFT (2048, 1024, 512, etc) for a lower group delay. A traditional FIR filter design is also an option for those who wish to experiment.

I believe that this issue is a matter of personal preference. If an experimenter wants to write FIR filters, FFTs and so forth from scratch, they are encouraged to do so. In fact, before I discovered the Intel SPL, I actually wrote an FIR filter and FFT in *Visual Basic* (quite slow in high-level code). I am not a *C++* programmer, so I did not want to go to the trouble of learning a new language in addition to all the other things I was working on.

When I wrote Part 1 of the series, the Intel Signal Processing Library was still free and available for download. See the short sidebar to my article in this issue for more information on the Intel IPP.

My opinion about starting from libraries is much like using integrated circuits. I can, if I so desire, build some functions in discrete logic that are available in a standard IC, but why do it if I am focused on the larger scope of a project? My goal in the article series was to make SDRs more accessible to the masses. Individual amateurs are free to experiment however they like. Your article series, "Signals, Samples, and Stuff...." [*QEX*, 1998], as well as other publications cover the math to allow one to write one's own DSP library; so I felt there was no reason for me to duplicate those excellent efforts.—*Gerald Youngblood, AC5OG,;* **AC5OG@arrl.net**

Understanding Switching Power Supplies, Part 2 (Jan/Feb 2003)

It appears that all of my delta-degree references were translated into absolute temperature references. 40 C degrees is a difference between two absolute temperatures. 40 degrees C is an absolute temperature that is three degrees above body temperature. This is a subtle but important difference!—*Ray Mack, WD5IFS, QEX Contributing Editor;* **rmack@arrl.org**

Theory of Intermodulation and Reciprocal Mixing: Practice, Definitions and Measurements in Devices and Systems, Part 2 (Jan/Feb 2003)

I finally made it through [this article] by Ulrich Rohde. [I have] some comments.

There is no "t" in Schwarz (look at the name on the spectrum analyzer in Fig 34). People like to put a "p" in the SGS Thomson (pronounced tom-son, not tomp-son) name.

The quantity of logarithmic power (dBm) does not add. In Fig 47 is the statement "90 dBm – 35 dBm = 55 dBm". This is incorrect. 0 dBm + 0 dBm does not equal 0 dBm. You have to convert to linear power, add, then convert back to logarithmic power. 1 mW + 1 mW = 2 mW. 2 mW is twice 1 mW so the answer is 3 dBm.

It makes it tough to get through an article when mistakes like this are made. Don't feel too bad as the trade journals also publish mistakes like this. And it is difficult to get through those articles. I have written to authors and magazines before on this.—*Larry Joy, WN8P, 2116 E Mohawk Dr, Olathe, KS 66062-2432, Life Member ARRL;* **lawrence_joy@yahoo.com**

Hi Larry,

And thanks for your comments. You are right about the error in the caption of Fig 47. We should have printed 55 dB, not dBm. The author is discussing a ratio of powers, in which case it is correct to subtract logarithmic units. In addition, part of the caption should read "...drop to about –90 dBm...."—Doug Smith, KF6DX, QEX Editor; **kf6dx@arrl.org**

Brainteaser (Jan/Feb 2003)

The brainteaser is a very good one that can teach many of the principles of transmission lines, SWR, directional couplers, directional wattmeters and the principle of superposition. It contains a subtle twist (a zinger!) that will stump many college-educated engineers. [The brainteaser] could serve as an educational tool—even at the college level—to illuminate many of these dark areas.

Many amateurs have trouble relating to purely symbolic letters; they feel more comfortable with real values. Thus, I have assumed that P = 1 W and R = 1 Ω. One volt RMS on a 1-Ω line equals 1 W. The more difficult concepts of the brainteaser can be illuminated by using the principle of superposition, which could be stated as follows: In a linear, reciprocal network, if applied signal *A* causes a response *a*, and applied signal *B* causes response *b*, then the simultaneous application of signals *A* and *B* will cause a response that is the vector voltage sum of *a* and *b*.

Since source *A* is a generator that is impedance-matched to the line, it cannot experience any signal-reflection effects. Therefore, it will act as a good dummy load for any external signal that is applied to it. Wattmeter *A* always indicates only the power that flows into the line from source *A*. Thus, wattmeter *A* will always indicate 1 W.

The Thevenin impedance of 3 Ω at source *B* will cause an SWR of 3:1. That causes a voltage reflection coefficient of $\rho = 0.5$. Since power is a function of voltage squared (ρ^2), that means that source *B* will reflect 1/4 of the power that's externally applied to it, although that may not be what wattmeter *B* indicates while source *B* is active.

At $t = 0$, each generator supplies 1 W to the line. If $t = 0$ is interpreted as "during the first picosecond," then even the mismatched source *B* is supplying that power. The internal generator of source *B* will have to generate 4 V RMS to accomplish that. The internal generator of source *A* will be set to 2 V. The difference in the voltages will determine the direction of current flow, and that will become important in the next steps.

The equation of operation of the 1-Ω impedance directional wattmeter of this brainteaser is as follows: Indicated Power = $([E + I] / 2)^2$, where E and I are the RMS vector quantities of voltage and current—meaning that they each have a polarity, a magnitude, and in some cases, a phase angle. The proper use of that equation requires calculating the voltages and currents—and the phase angles—that are present at each wattmeter. Many amateurs and engineers have trouble doing this when there are multiple sources and transmission lines present. Therefore, some simplifications are in order.

The "one wavelength of cable" means that the 1-V signal from source *A* differs by 360° when it arrives at source *B* and it is in-phase with the signal that is being generated by source *B*. In the steady-state response, a 360° difference is essentially the same as a 0° difference. Therefore, the one wavelength of cable can be replaced by a length of zero for these calculations. Next, if this circuit is truly linear, then the frequency of operation could be lowered to 0 Hz and the sources represented by batteries and resistors. That would allow source *A* to be replaced with a 2-V battery in series with a 1-Ω resistor, and source *B* will be a 4-V battery in series with a 3-Ω resistor. Both source batteries have their negative leads grounded, making them "in-phase." Notice that each of these new sources would, again, deliver exactly 1 V (or 1 W) into a 1-Ω load resistor, as the brainteaser requires. Bear in mind these simplifications are only valid for calculating the steady-state conditions at this one frequency, and the simplified circuit would not represent the real system's response if the frequency were changed.

By Ohm's Law and with those simplifications, we find the current between the sources (caused by a 2-V difference across a total of 4 Ω) flows to the left at 1/2 A, and the line voltage is 2.5 V. Wattmeter *A* sees 2.5 V and a current of minus 1/2 A (since it flows opposite to the favored direction). Putting this information into the equation above yields an indicated power of $([2.5\text{-}0.5]/2)^2 = 1$ W.

Wattmeter *B* sees 2.5 V and a current of plus 1/2 A (since it flows in the favored direction). This causes a reading of $([2.5+0.5]/2)^2 = 2.25$ W.

Source *A* and source *B* were each set up to deliver 1 V (1 W) into a 1-Ω line. You could say that the source-*A* signal reached the source-*B* impedance of 3 Ω where it realized a 3:1 SWR, which created a co-phased reflected signal of +1/2 V. Source *B* was set up to generate a 1-V signal into a 1-Ω line, therefore wattmeter *B* senses the +1/2 V of reflection in its favored direction, plus the 1 V generation—a total of 1.5 V. Thus, it reads 2.25 W (1.5 V squared). These readings will be disquieting to many operators since neither generator (nor their sum) is producing 2.25 W. Also, the real system energy budget is different, again.

When looking at the simplified circuit, notice that the 4-V battery is supplying 1/2 A, thus it is generating 2 W. The 3-Ω resistor is dissipating 3/4 W. The 1-Ω resistor is dissipating 1/4 W, and the 2-V battery is being charged at the rate of 1 W; thus the wattages all add up in a way that seems reasonable, but they are different from the wattmeter readings. By the way, the simplified circuit line voltages and currents, the internal generator wattages and the dissipations would all be the same if the sources were real RF generators and the 1 wavelength of line were present. It is disquieting to see that wattmeter *B* gives a reading that has more than one interpretation, particularly when multiple sources and a mismatch are both present. In that simultaneous situation, the individual components of its reading (voltage and current) must be found to predict the response.

When source *B* has a 1/3-Ω Thevenin impedance, its internal signal generator has a magnitude of 4/3 V. With this change to the simplified circuit, the new current is caused by a 2/3-V difference across a total of 4/3 Ω, and 1/2 A flows to the right. The new line voltage is 1.5 V. Wattmeter *A* will read $([1.5 + 0.5] / 2)^2 = 1$ W. Wattmeter *B* will read $([1.5 - 0.5] /2)^2 = 0.25$ W. From a transmission-line point of view, the 1-V signal from source *A* reaches the 1/3-Ω impedance and again realizes a 3:1 SWR, or a reflection of 1/2 V.

But this time, the less-than-1-Ω termination causes the 1/2-V reflection to undergo a phase reversal upon reflection. Thus, the 1-V signal that source *B* was set up to generate is out of phase with the reflected signal and wattmeter *B* senses (1 – 0.5) V, or 0.25 W. The new dc circuit has 1/2 A flowing to the right, thus the 2-V battery is generating 1 W, the 1-Ω resistor dissipates 1/4 W, the 1/3-Ω resistor dissipates 1/12 W, the 4/3-V battery is being charged at 2/3 W, and thus the principle of conservation of energy is being satisfied.

I hope my explanations of wattmeter operation will make users feel a little more comfortable when using this very useful and slightly troublesome instrument.—*Richard T. Knadle, K2RIW, 316 Vanderbilt Pkwy, Dix Hills, NY 11746-5856;* **rknadle@suffolk.lib.ny.us**

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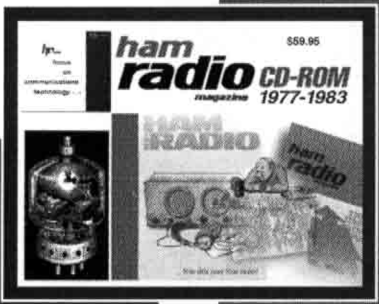

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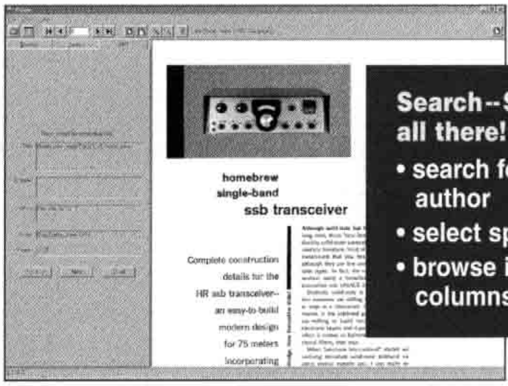

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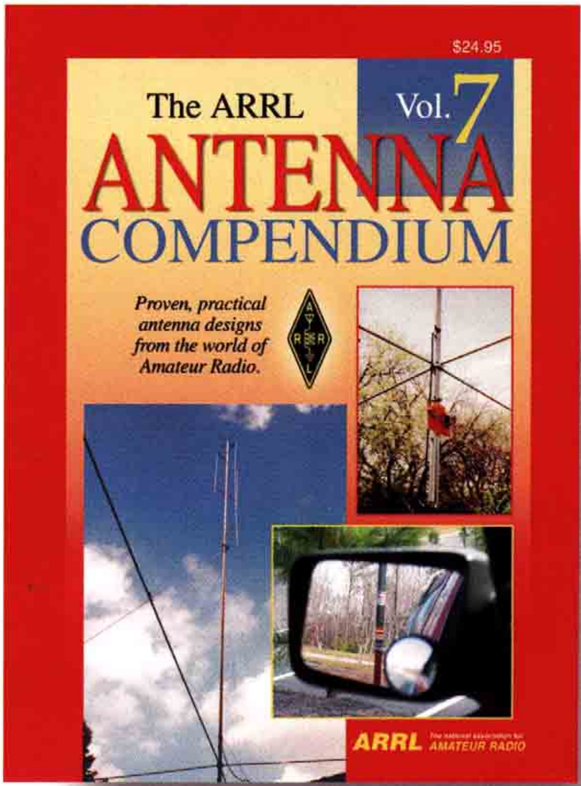

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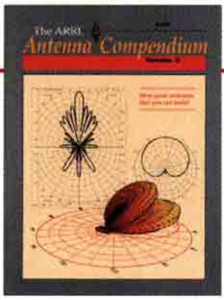

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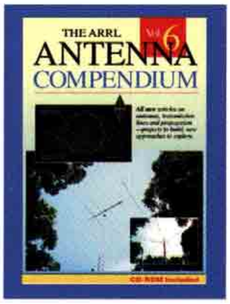

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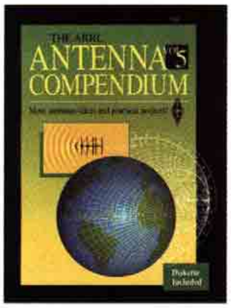

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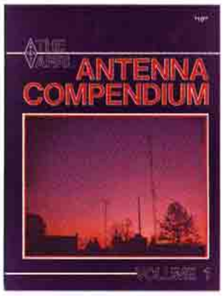

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