QEX

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A Forum for Communications Experimenters

Issue No. 245

KW7CD introduces his ***Star-10*** transceiver, a high performance, fully synthesized, continuous coverage, coherent HF transceiver.

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About the Cover

Cornell Drentea, KW7CD introduces his *Star-10* transceiver and describes some of the criteria involved in designing a high performance, fully synthesized, continuous coverage, coherent HF transceiver.

Features

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THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interest in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of the radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

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The purpose of *QEX* is to:
1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,
2) document advanced technical work in the Amateur Radio field, and
3) support efforts to advance the state of the Amateur Radio art.

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and letters for publication in *QEX* should be marked Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX* or high-resolution digital images (300 dots per inch or higher at the printed size). Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Empirical Outlook

Larry Wolfgang, WR1B
lwolfgang@arrl.org

Never Stop Learning!

Although for many of us it has been a long time since we have seen the inside of a classroom (at least from the students' perspective), we should never say we are done with school, or have learned all that we can learn about a subject. Education is a life-long process. One of the things that has helped hold my interest in Amateur Radio for so many years is that there always seems to be something new to learn about. Whether it is a mode that I have not yet tried, construction techniques with new types of components or some different operating techniques, there always seems to be something else to learn.

In early August I had the opportunity to participate in an ARRL Teachers' Institute on Wireless Technology, conducted by Mark Spencer, WA8SME, at ARRL Headquarters. Mark taught the course at HQ for several years, and last year he also conducted several Institutes around the US.

The TI is a week-long, in-residence learning opportunity designed for motivated teachers and other school staff who want to learn more about wireless technology and bring that knowledge to their students. A variety of topics are covered during the 4 days of the TI, including basic wireless technology literacy, electronics, and the science of radio; bringing space into the classroom; ham radio operation; introduction to microcontrollers; and basic robotics. This is part of ARRL's Education and Technology Program, bringing wireless technology education to schools. It isn't teaching Amateur Radio licensing classes, and teachers who participate in the TI and/or who implement the program in their schools do not have to be licensed hams. Many of the participants are hams, however, or earn licenses after becoming involved in the ETP.

During this four-day program, Mark teaches how to instruct various age groups about electronics and wireless technology. His demonstrations and activities are focused on the ARRL Education and Technology classroom curriculum, but could be used in a wide variety of settings. In my case, I plan to use some of the techniques in my work with Scout groups, teaching Radio Merit Badge classes, and also adapting some of the techniques to license classes when I help a local radio club teach the Technician license material.

Early in the week, Mark does the "magnet through a copper or aluminum pipe" demonstration. I've read about this demonstration, and understand the principles upon which it is based, but that didn't quite prepare me for how dramatically slower the magnet falls through the conductive pipes than it does in free-fall, or through a PVC pipe, for example.

Mark has designed and built a number of "Exploration Boards" or "Experiment Boards" for teachers to use. For example, there is an Ohm's Law Exploration Board that includes an on-board digital VOM and various two and three-pin connectors. Mark even provides an envelope full of various resistors already wired into the mating connectors. By plugging in resistors and connecting jumpers, you can measure the voltage and current at various points in the circuit. By recording the values given on the digital display, you have a very effective demonstration of Ohm's Law.

Another board provides a circuit to allow teachers to explore basic resonant circuits and some DSP basics. Still another board provides various signals in a circuit that allows you to demonstrate frequency and wavelength relationships and even mix signals to demonstrate modulation (and demodulation).

A highlight of the course for me was learning to program a BASIC Stamp and seeing how the microcontroller could be used to sense some condition and then control an output. I've read many articles about using microcontrollers (and edited a few for publication), and I've done some BASIC programming over the years, but I had never actually written a program and copied it into the microcontroller. The hands-on learning approach is a lot more fun than simply reading about it, or listening to a lecture.

I guess that's the point. When we are trying to teach something about our hobby and share our passions with others, no matter their age, a hands-on, actually-doing-something approach beats just telling them about it every time. That's probably not news to any of our readers, but maybe it will help us to pause and think about it for a moment. What demonstrations or techniques have you used to teach a class or help a newcomer understand the concepts of your passions? What might you do to make them more effective? Why not share that information with others? We would welcome your ideas.

I'd like to close with a more direct plug for the ARRL Education and Technology Program. If you are interested in sharing the excitement of wireless technology and electronics in general, and Amateur Radio specifically, this program is worthy of your support. The ARRL Development Office would be happy to provide you with more information about how your contribution can help sustain this valuable program. *— 73, Larry Wolfgang, WR1B.*

The *Star-10*© Transceiver

In Part 1 of this series, we learn about some of the design criteria involved in high performance, fully synthesized, continuous coverage, coherent transceivers.

Cornell Drentea, KW7CD

Introduction

It has always been the dream of the technically inclined radio amateur to build his or her own equipment from scratch. Such has been the case in the first part of the 20th century, when Amateur Radio equipment was relatively easy to construct, allowing for simple home building. While single-band or band-switched equipment has been relatively easy to realize, RF design has recently evolved into a complex art of mixed technologies using new solid-state components, novel frequency generation techniques, microprocessors, digital logic and signal processing techniques that have exceeded the capabilities of the average amateur operator. This, in turn, has put the design of complex high performance radio equipment beyond the scope of the casual experimenter or equipment builder, forcing hams, more and more, to become appliance operators. Increasingly sophisticated equipment design has presented a steep learning curve even for the most capable home builder or the modern radio equipment manufacturer.

In the past, many construction articles have been published regarding simple radio projects. Dedicated single-band or limited band-switched, down-conversion superheterodyne receivers and transceivers have been published extensively in the literature. More recently, so-called "software defined radios," using old-fashioned zero IF direct conversions combined with new personal computer digital audio cards have evolved. Their performance has been controversial, only to be obscured by their perceived "flexibility."

Less published have been multi-band radios, due to their increased band switching complexity. Even less attempted have been full coverage high performance professional grade, up conversion / down conversion transceivers featuring fully synthesized, high resolution, coherent schemes used in conjunction with front-end automatically switched half-octave filter banks to ensure consistent high dynamic range performance over several octaves. Such designs have been left to the professional manufacturers, who can invest significant amounts of money and engineering resources over long periods of time for gain and profit.

This situation need not be so. With enough dedication, today's technically inclined ham is fully capable of developing full coverage, high performance transceivers that can compete in performance and features with their professional counterparts, and even outperform these designs.

This series of articles describes the development of just such a transceiver, the *Star-10,* a high dynamic range, fully synthesized coherent RF system that tunes continuously from 1.8 MHz to 30 MHz using microwave synthesis, coupled with true automatically switched (using miniature RF relays) half-octave filter banks, using a 10 Hz ultimate step resolution. This work is intended to inspire the radio amateur and the professional engineer alike, regarding modern, full coverage transceiver design.

The *Star-10* transceiver bears its name in good memory of an ambitious project I had been associated with in my youth — a product

The completed *Star-10* transceiver is designed to receive and transmit anywhere from 1.8 MHz to 30 MHz with a resolution of 10 Hz. It provides full cross mode, end-to-end RX/TX Split operation with a transmitter RF power of 100 W (125 W peak). The receiver has a composite spurious-free linear dynamic range in excess of 150 dB. It exhibits an MDS of –136 dBm (absolute), a third-order intercept point of +45 dBm. A coherent DDS-Driven PLL microwave synthesizer provides close in phase noise performance of –133 dBc/Hz. Receive spurious image rejection is –75 dB and transmit harmonic rejection exceeds –55 dBc. The transceiver uses a coherent up-convert down-convert superheterodyne approach.

The electrical and mechanical design features a modular approach using eighteen double sided, plated through printed circuit boards housed in machined, irradiated aluminum assemblies, all packaged in a custom made, hammer-tone finished cabinet measuring 15.5 x 6 x 11 inches.

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that has never happened. It is the culmination of several years of RF design and development and reflects a state-of-the-art approach to HF transceiver design. The implementation encompasses the many phases of engineering and development usually encountered in a complex commercial or military piece of equipment, from the system design through the circuit and software design, the multiple brass boarding, the complex testing, and packaging into a final form factor as shown above.

The *Star-10* transceiver was designed to receive and transmit seamlessly with an ultimate tuning resolution of 10 Hz anywhere in its four-octave frequency range, covering any HF ham band, past, present and future, and featuring high dynamic range performance exceeding even today's most modern professional transceivers. Its receiver features a composite spurious free linear dynamic range exceeding 150 dB over the entire frequency coverage. In addition, this transceiver is capable of transmitting in several modes, 100 W of RF (125 W peak) power from 1.8 to 30 MHz. Transmitter harmonic rejection and receiver image rejection meet or exceed commercial equipment requirements. (See the Specification section of the text.)

The transceiver's entire capability is slaved to a powerful 8 bit PIC microprocessor that runs approximately 10,000 lines of code continuously at 32 MHz (above the HF range to keep spurious products out of the receiver's input range) in a closed loop, only to be interrupted by its keypad or RS-232 commands. The *Star-10* has been designed with a flexible and friendly human interface that can only be compared with the feel of classic HP test equipment.

The Challenge

The *Star-10* project first evolved in the 1980s (see References 1 and 2) and has been recently upgraded using the previously designed half octave filter banks as combined with the latest state-of-the-art microprocessor, DDS-PLL and high dynamic range RF technologies. As such, the command and control system of the old *Star-10* design (see References 1 and 2) has been physically reduced from its old hardwired static logic implementation of over thirty integrated circuits, to a simple command and control board containing a single microprocessor and a minimum of additional control circuits. The complexity of the entire command and control functionality of the transceiver has been moved into software containing approximately 10,000 lines of code. See Figure 1.

Although the hardware command and control section was simplified from the previous design, the new design has a new front end, a new first IF, a new logarithmic/linear second IF, specially designed crystal filters, and a new microwave synthesizer (see References 3 and 4), which all have contributed to increased dynamic range. This will be described in detail later. In the interest of making this article series as short as possible, and because of the complexity of this project, block diagrams, simplified schematics and test specifications are used throughout the article. Consequently, there are no boards, parts or software available from this source.

The *Star-10* transceiver has been a unique research experience into understanding what can be done — from the point of view of the laws of physics — in receiver and transceiver dynamic range performance. This research has been performed over a period of five years with parts, technologies and packaging means available to me at the time. The transceiver has been implemented with some unique parts that may not be available anymore. The *Star-10* development has been a purely scientific endeavor, intended primarily to understand what could be done to achieve ultimate receiver performance. Although the results have been outstanding, slightly better results may be possible using newer technologies and parts. The *Star-10* project was not intended as a commercial product. Its duplication is not economically feasible.

Acknowledgements

This project has been an ambitious scientific undertaking, developed with a considerable investment, equivalent to the price of a top of the line transceiver and with additional help. Because of the complexity of this transceiver, there are no circuit boards or software available.

Many thanks are extended to several individuals and companies who made contributions toward this development. Special thanks go to Chris Sisemore, who took my complex system mathematics and developed approximately 10,000 lines of perfect code for the command and control section of the transceiver. His technical discipline and diligent work with me in testing the brassboards of the system over a period of better than a year in the command and control section of the transceiver have paid off in achieving a flawless system performance and a most friendly operator interface.

Additional thanks go to Randy Burcham, KD7KEQ, who laid out (sometimes twice) the complex double sided, plated through printed circuit boards. His special attention to multiple stitching of ground planes and to properly placing the RF components have made for well-behaved high gain amplifier circuits that did not oscillate, and quiet synthesizer circuits. He also fully documented the entire design in a true engineering fashion using D-size engineering drawings.

Equal thanks go to Constantin Popescu, KG6NK, who worked diligently with me, breadboarding and testing the system before layout in his well-equipped laboratory. He has also been very instrumental in the harnessing and troubleshooting of the final system implementation. Additional thanks go to George Cutsogeorge, W2VJN, of In-

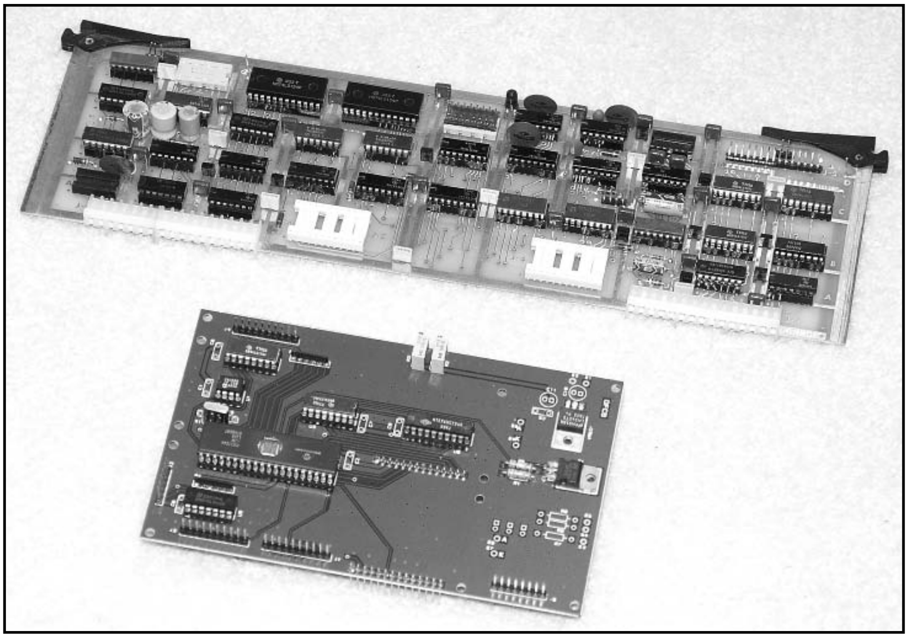

Figure 1 — The original *Star-10* command and control board (top) contained over thirty ICs. It has been replaced in the new design (bottom) by a single microprocessor controller chip and a minimum of additional circuits. Approximately 10,000 lines of software code have replaced the original hardwired logic functionality of the 1982 circuitry (See References 1 and 2).

ternational Filter Company who developed the high performance ultimate bandwidth IF crystal filters, Jerry Buckwalter of Alpha Components company, who developed the demanding high-intercept fundamental-type first IF roofing filters, the TEMEX Corporation who developed additional roofing filters, Ulrich Rohde, N1UL, of Synergy Microwave, who donated a low noise microwave VCO for the frequency reference unit (FRU) synthesizer and reviewed this article series, as well as to Peter Chadwick, G3RZP, of the former Plessey Company who donated samples of the high performance professional grade aerospace PLL chip used in the FRU. Additional thanks go to the Alinco Corporation, who made power transistors and other parts available for the power linear amplifier.

Finally, many thanks go to Phil Aide, KF6ZZ, who applied his switching power supply aerospace experience to develop a

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Specifications

General

- Frequency coverage
  - RX: 1.8 MHz to 29.99999 MHz in one band, continuous in 10 Hz
  - TX: 1.8 MHz to 29.99999 MHz in one band, continuous in 10 Hz
  - Split operation from 1.8 MHz to 29.99999 MHz, any mode (or cross mode)
  - Display: Composite 4 × 16 character, large green 320 × 240 dots LCD Twist Dot Matrix with green backlighting
- RX/TX Front end Filter system: Automatically switched independent half octave band-pass (RX) and high power low-pass (TX) filter banks:
  - 1.8 MHz to 3 MHz
  - 3 MHz to 4 MHz
  - 4 MHz to 6 MHz
  - 6 MHz to 8 MHz
  - 8 MHz to 12 MHz
  - 12 MHz to 16 MHz
  - 16 MHz to 24 MHz
  - 24 MHz to 30 MHz
- Modes: USB, wide, narrow; LSB, wide, narrow; CW, wide, narrow; RTTY/AFSK, PSK31, wide, narrow
- IF passband tuning (PBT): ± 1.5 kHz, all modes, RX and TX
- Receiver incremental tuning (RIT): ± 9.9 kHz
- Architecture: Coherent double conversion superheterodyne, first IF = 75 MHz, second IF = 9 MHz
- Roofing filter (75 MHz): 8 pole crystal filters in two banks, BW = 10 kHz
- Second IF (9 MHz) ultimate bandwidths: 32 pole cascaded, SSB 2.4 kHz, 1.8 kHz; CW, RTTY/AFSK, 0.5 kHz; crystal filter insertion loss compensated automatically
- Second IF (9 MHz) gain = logarithmic linear 100 dB with 2.4 kHz crystal filter output for noise reduction using Analog Devices AD603 amplifiers
- Total AGC range: 80 dB nominal, 120 dB total (AIPA + BIPA + 9 MHz AGC)
- AGC attack time <2 ms, Decay time 4 seconds (SSB), 1 second (CW)
- System warm-up time to 1×10^{-8} <30 seconds
- Tuning speed: <10 ms
- S-meter: Calibrated in dBm and S units (within 2 dB)
- Digital memory channels: 99 (2 scan edges)
- RF output power (continuously adjustable):
  - SSB/CW/AFSK/RTTY: 0 to 100 W (125 W peak)
- Modulation:
  - SSB/RTTY/AFSK: Class III — High level double balanced modulator used
  - KCW: Class III — High level double balanced modulator carrier insert
- Spurious emissions: Equal or better than –55 dBc
- Carrier suppression: Equal or better than –65 dBc
- Unwanted sideband suppression (16 pole filters used): Equal or better than –65 dB
- Phase noise RX/TX: –133 dBc/Hz close in
- Spurious RX/TX: –75 dBc or better
- RX sensitivity (500 Hz ultimate bandwidth): –136 dBm (absolute) with 32 pole filters cascaded (plus 8 pole roofing filter)
- RX IIP3: +45 dBm
- RX composite linear DR: Equal or better than 150 dB (500 Hz, Preamp on, all AGCs on)
- RX IP3SFDR: At least 130 dB (20 kHz tone spacing) (500 Hz, preamp on, all AGCs on)
- RX blocking dynamic range: Will receive a –110 dBm signal with 25 dB SNR in the presence of a –20 dBm signal located 5 kHz away (500 Hz ultimate BW, preamp on, no attenuators, no AGC action)
  - Advanced intercept point attenuator (AIPA): Programmable –3 dB, –6 dB, –10 dB steps
  - Preamplifier Gain: +10 dB
  - RF/IF gain PIN attenuator (BIPA): 30 dB front panel adjustable
- RX noise figure at MDS: 15 dB (no AGC action)
- Selectivity:
  - SSB — USB, LSB selectable: 2.4 kHz, 1.8 kHz, at –3 dB cascadable from 16 poles to 24 poles
  - CW, RTTY/AFSK/PSK31, USB, LSB selectable: 1.8 kHz and 500 Hz/–3 dB: composite cascadable to 32 poles (plus 8 pole roofing filter)
- Image and spurious rejection: Equal to or better than –75 dB
- AF output power: 2.6 W at 10% distortion with an 8 Ω load
- Spectrum analyzer output (75 MHz or 9 MHz ± 250 kHz)

Synthesizer — Frequency Reference Unit (FRU)

- DDS Driven PLL 0.75 to 1.05 GHz divided by 10, for 20 log 10 (20 dB) phase noise improvement
- FRU frequency resolution: 10 Hz (1 Hz at DDS frequencies)
- FRU phase noise RX/TX: –133 dBc/Hz close in
- FRU spurious rejection –90 dBc
- Tuning lock up time: continuous within <10 ms.

Master Reference Unit (MRU)

- 84 MHz — Phase locked to 10 MHz OCXO/WWV
- Aging 10 Hz in 20 years.
- Long term frequency stability over temperature: 1×10^{-8} provided by the 84 MHz master reference unit (MRU) — PLXO phase locked to a 10 MHz OCXO/WWV controlled on power up.
- MRU warm-up time to $1 \times 10^{-8} <= 30$ seconds from power up (system warm-up)
- MRU phase noise: –165 dBc /Hz close in or better

Power Supply

- Power supply in: 70 to 140 V ac, 50/60 Hz
- Power supply out (RX): 24.7 V dc at 2 A continuous, 3.5 A peak
- Power supply out 2 (TX): 13.7 V dc at 20 A max continuous
- Power supply spurious as seen by the receiver: –145 dBm at any frequency in the coverage or TBD
- Power consumption:
  - TX max dc power: 800 VA
  - RX standby dc power: 200 VA (typical)

Mechanical

- Dimensions (projections not included): 15.5(L) × 11(W) × 6(H) inches
- Weight: 30 lbs
- Antenna connectors: SO-239 (50 Ω) and BNC (50 Ω)
- Operating Temperature range: 0°C to +50°C

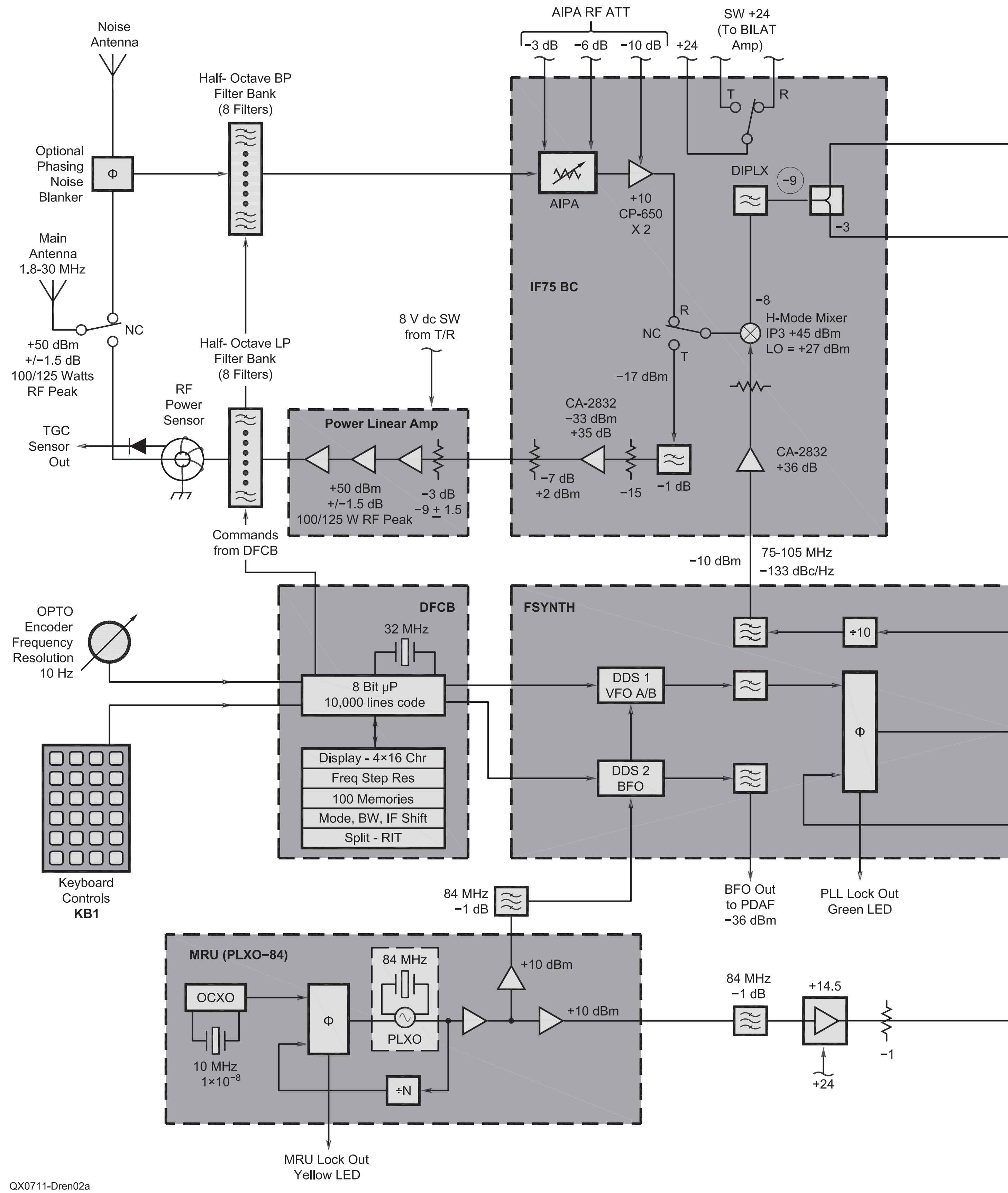
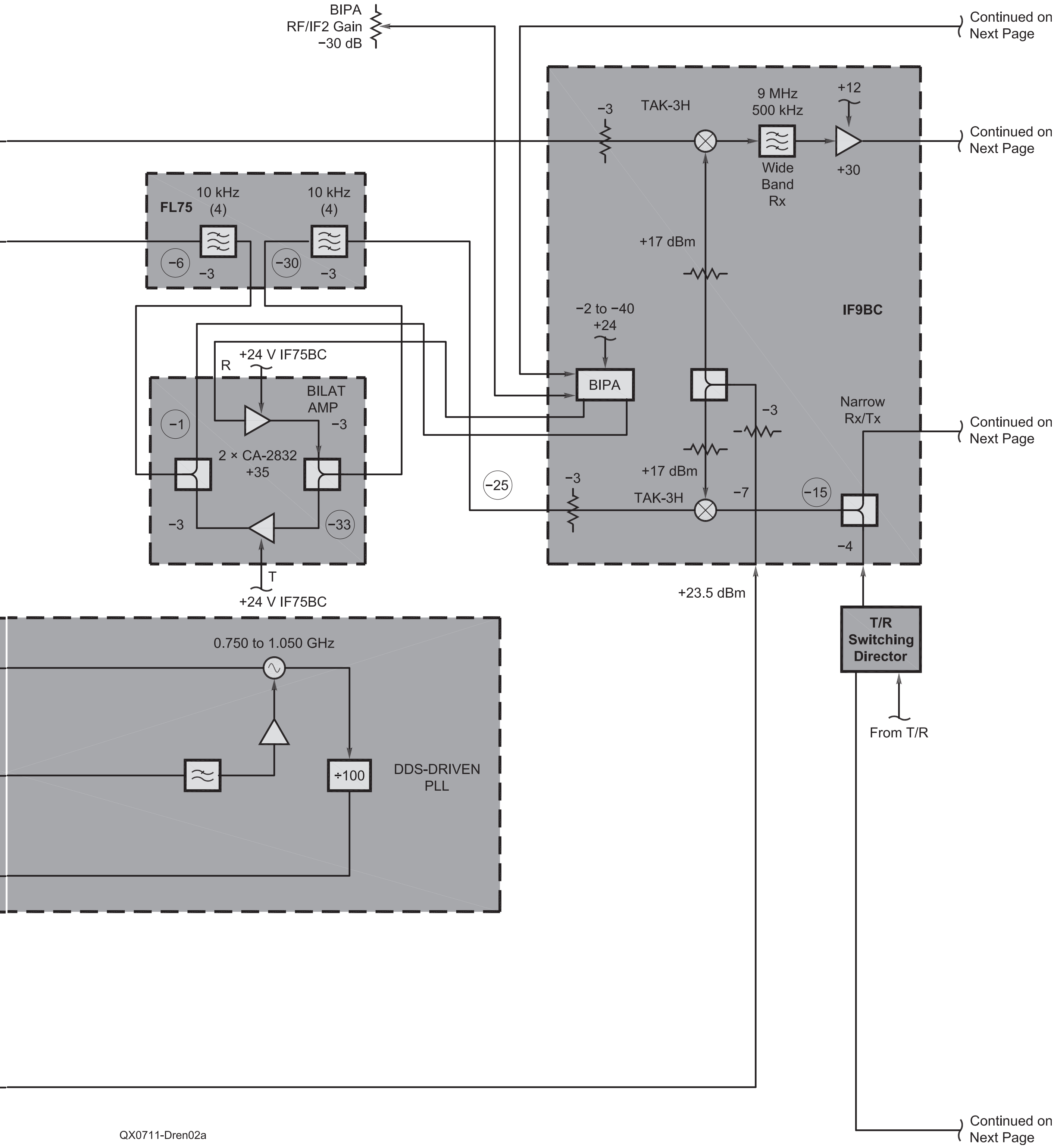


Figure 2 — This block diagram shows the *Star-10* transceiver circuit boards. Note that the diagram continues to three pages.


BIPA
RF/IF2 Gain
−30 dB
Continued on
Next Page
−3
TAK-3H
9 MHz
500 kHz
+12
Continued on
Next Page
+30
Wide
Band
Rx
FL75
10 kHz
(4)
10 kHz
(4)
−6
−3
−30
−3
+17 dBm
IF9BC
−2 to −40
+24
+24 V IF75BC
R
BILAT
AMP
−1
−3
BIPA
−3
Narrow
Rx/Tx
Continued on
Next Page
2 × CA-2832
+35
−25
−3
+17 dBm
TAK-3H
−7
−15
−3
−33
−4
T
+24 V IF75BC
+23.5 dBm
T/R
Switching
Director
0.750 to 1.050 GHz
From T/R
÷100
DDS-DRIVEN
PLL
QX0711-Dren02a
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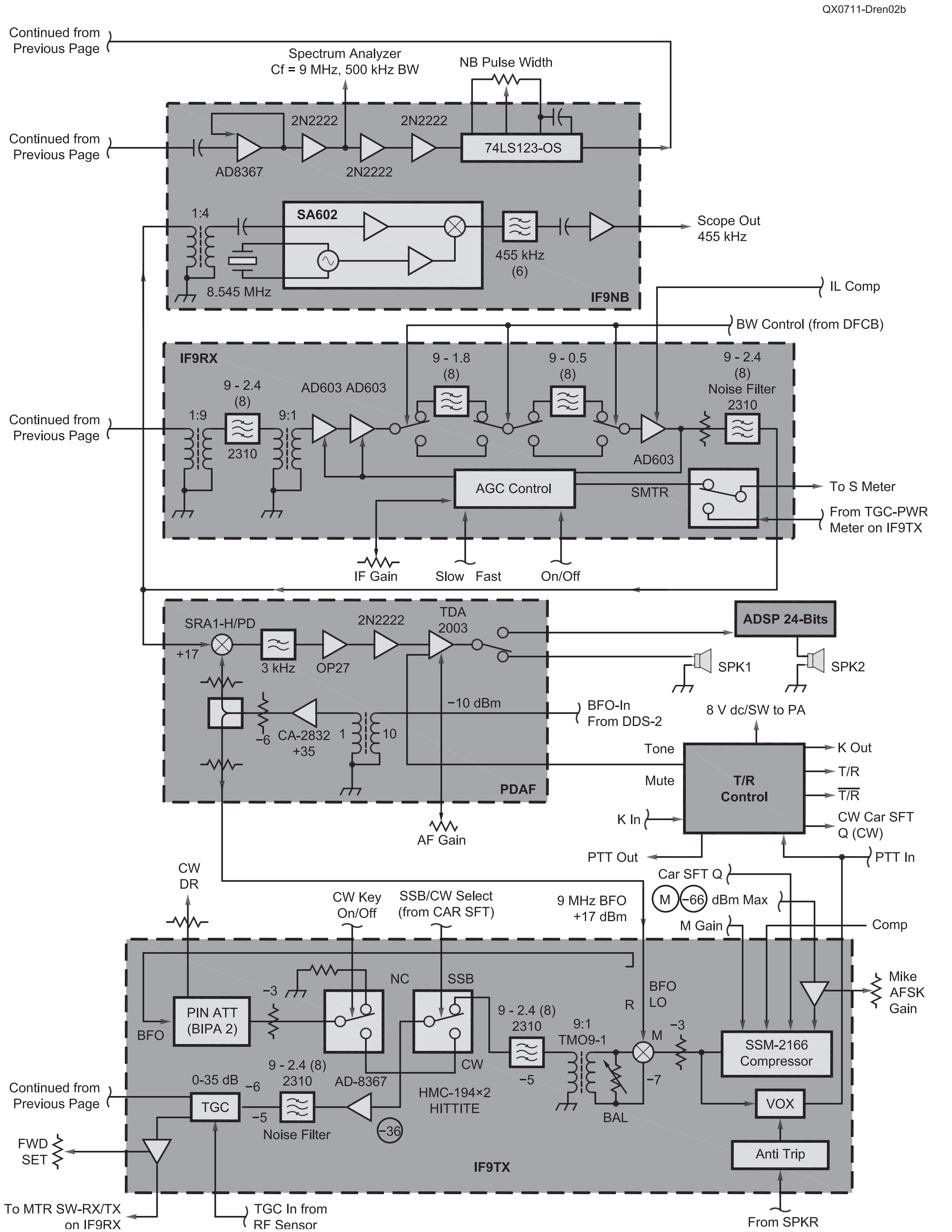

Figure 2 — (continued)

very quiet switching power supply specifically for the *Star-10.*

Design Goals

From the start, the *Star-10* transceiver had two key design goals. The primary goal was to produce a continuous HF coverage system with consistent high dynamic range receiver performance over the entire frequency range that rivals the performance of top of the line equipment. Many receivers today have different performance characteristics at different points in their frequency coverage. The focus of the *Star-10* design was on no-compromise wide-band architecture while maintaining the broadband approach and without accent on unnecessary bells and whistles.

The second goal was to maintain a rigorous and disciplined physical implementation to approach commercial or mil spec grade equipment. These efforts were realized through progressive packaging techniques against the self-imposed system and the circuit design. In addition, the *Star-10* design matured through using ample and gradual trade studies as well as comprehensive design verification techniques and tests consistent with standard engineering processes.

An effort was made to make all interconnecting RF interfaces between assemblies 50 Ω. All RF connectors are gold plated SMA types. RG-194 Teflon and low loss semirigid coaxial cables have been used throughout.

As previously mentioned, the *Star-10* covers seamlessly and continuously the entire HF range of 1.8 MHz to 30 MHz in a single band with a 10 Hz frequency resolution, and with an ultimate receiver composite linear dynamic range of 150 dB or better. (Note: composite linear dynamic range is defined as the ability to funnel a given large input RF signal range into a final transducer without compressing and using multiple AGCs.)

The transceiver is a dual conversion up convert / down convert design that features automatically switched (using miniature RF relays) half-octave filter banks in the front end and a high first IF for superior image, spurious and harmonic rejection over the entire frequency range. Again, there is no channelized single-band-only coverage, like in some of the so-called “high performance” limited coverage 9 MHz IF transceivers found on the market today. The bells and whistles have been limited to the essentials, but plenty of software functionality has been provided throughout. The requirements and specifications for the *Star-10* transceiver are listed below. Dynamic range numbers represent goals as well as final tested results.

STAR-10 ULTRA - HIGH DYNAMIC RANGE COMMUNICATIONS RECEIVER ANALYSIS

System P Input (dBm)	Analysis Temp	AGC 1		AGC 3		
-132.00	25 °C	ON 0	10	ON 0	80	Analyze
System P Out (dBm)	VIEW 600 X 800	AGC 2		AGC 4		
6.00	SET SIZE TO 75%	ON 0	30	ON 0	0	Reset
						FALSE Processor Gain 1.00

MDS (dBm)	Linear DR (dB)	ULT BW (Hz)	N Floor (dBm)	Noise Fig (dB)	ICP1dB (dBm)	IIP3 (dBm)	IIP2 (dBm)	System OIM3	SFDR (dBm)
-132.2	34.3	500	-146.9	14.7	-98.0	-93.0	-88.3	-72.0	9.6
Pout (dBm)	**Comp (dB)**	**Gain (dB)**	**N Floor (dBm)**	**SNR (dB)**	**OCP1dB (dBm)**	**OIP3 (dBm)**	**OIP2 (dBm)**	**SFDR3 (dB)**	**SFDR2 (dB)**
6.0	0.0	138.0	5.8	0.3	40.0	45.0	49.7	26.2	22.0

	INPUT DEVICE PARAMETERS @ 25°C											CUMULATIVE PERFORMANCE									
	Device Description	Device Gain (dB)	Delta Gain (dB/°C)	Device CP1dB (dBm)	Delta CP1dB (dB/°C)	Device OIP3 (dBm)	Delta OIP3 (dB/°C)	Device OIP2 (dBm)	Device NF (dB)	Delta NF (dB/°C)	Device BW (MHz)	Gain (dB)	Comp Level (dB)	Signal Power (dBm)	Noise Floor (dBm)	NF (dB)	Cum DR (dB)	IIP3 (dBm)	IIP2 (dBm)	Cum SFDR3 (dB)	Cum SFDR2 (dB)
1	FL HO	-1.00	0.00	99.00	0.00	99.00	0.00	99.00	1.00	0.00	30	-1.00	0.00	-133.00	-146.87	1.00	198.23	100.00	100.00	132.15	99.11
2	AIPA AGC1	0.00	0.00	99.00	0.00	99.00	0.00	99.00	0.00	0.00	30	-1.00	0.00	-133.00	-146.87	1.00	195.22	96.99	93.98	130.15	96.10
3	PREAMP	10.00	0.00	40.00	0.00	45.00	0.00	65.00	5.00	0.00	30	9.00	0.00	-123.00	-131.87	6.00	124.23	36.00	55.95	86.15	74.59
4	HMIX	-8.00	0.00	40.00	0.00	45.00	0.00	65.00	8.00	0.00	30	1.00	0.00	-131.00	-139.20	6.67	122.92	35.36	53.05	85.28	72.80
5	DIPLEX	-1.00	0.00	99.00	0.00	99.00	0.00	99.00	1.00	0.00	30	0.00	0.00	-132.00	-140.01	6.86	122.73	35.36	53.01	85.15	72.69
6	SPLITTER	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	1.00	0.00	30	-3.00	0.00	-135.00	-142.78	7.09	122.50	35.36	53.02	85.00	72.58
7	FL75A	-3.00	0.00	10.00	0.00	15.00	0.00	20.00	3.00	0.00	0.01	-6.00	0.00	-138.00	-144.36	8.51	141.33	20.84	25.62	97.55	75.56
8	BI-SPL1	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	3.00	0.00	30	-9.00	0.00	-141.00	-145.43	10.44	139.41	20.84	25.62	96.27	74.59
9	BIPA AGC2	0.00	0.00	20.00	0.00	15.00	0.00	35.00	0.00	0.00	0.01	-9.00	0.00	-141.00	-145.43	10.44	139.20	19.13	24.63	95.13	74.10
10	BI AMP	36.00	0.00	40.00	0.00	45.00	0.00	65.00	5.00	0.00	30	27.00	0.00	-105.00	-105.36	14.51	130.60	15.52	22.95	90.01	71.22
11	BI-SPL2	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	3.00	0.00	30	24.00	0.00	-108.00	-108.36	14.51	130.60	15.52	22.92	90.01	71.21
12	FL75B	-3.00	0.00	20.00	0.00	25.00	0.00	30.00	3.00	0.00	0.01	21.00	0.00	-111.00	-111.36	14.51	118.23	3.70	7.41	82.13	63.45
13	PAD	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	3.00	0.00	30	18.00	0.00	-114.00	-114.36	14.51	118.23	3.70	7.41	82.13	63.45
14	MIX TAK3 H	-7.00	0.00	14.00	0.00	24.00	0.00	30.00	7.00	0.00	30	11.00	0.00	-121.00	-121.35	14.52	116.84	3.22	5.38	81.80	62.43
15	SPLITTER	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	4.00	0.00	30	8.00	0.00	-124.00	-124.33	14.54	116.82	3.22	5.38	81.79	62.42
16	FL9 2.4	-3.00	0.00	40.00	0.00	45.00	0.00	50.00	3.00	0.00	0.0024	5.00	0.00	-127.00	-127.31	14.56	122.99	3.22	5.29	85.90	65.46
17	IF 9 AGC3	0.00	0.00	99.00	0.00	99.00	0.00	99.00	0.00	0.00	0.0024	5.00	0.00	-127.00	-127.31	14.56	122.99	3.22	5.29	85.90	65.46
18	AD 603 x 2	84.00	0.00	40.00	0.00	45.00	0.00	50.00	5.00	0.00	0.0024	89.00	0.00	-43.00	-43.20	14.67	76.53	-44.00	-39.05	54.36	43.24
19	FL9 1.8	-3.00	0.00	40.00	0.00	45.00	0.00	50.00	3.00	0.00	0.0018	86.00	0.00	-46.00	-46.20	14.67	76.02	-45.76	-43.68	54.01	41.55
20	FL9 0.5	-6.00	0.00	40.00	0.00	45.00	0.00	50.00	7.00	0.00	0.0005	80.00	0.00	-52.00	-52.20	14.67	81.23	-46.11	-45.31	57.49	43.52
21	AD 603 FIXED G	20.00	0.00	40.00	0.00	45.00	0.00	50.00	5.00	0.00	0.0005	100.00	0.00	-32.00	-32.20	14.67	71.82	-55.53	-53.99	51.21	39.18
22	PAD	-12.00	0.00	99.00	0.00	99.00	0.00	99.00	12.00	0.00	30	88.00	0.00	-44.00	-44.20	14.67	71.82	-55.53	-53.99	51.21	39.18
23	FL9 2.4	-3.00	0.00	40.00	0.00	45.00	0.00	50.00	3.00	0.00	0.0024	85.00	0.00	-47.00	-47.20	14.67	71.70	-55.65	-54.91	51.13	38.72
24	PD SRA1H	-7.00	0.00	10.00	0.00	23.00	0.00	30.00	7.00	0.00	0.0024	78.00	0.00	-54.00	-54.20	14.67	63.61	-58.35	-58.15	49.33	37.10
25	AF TDA 2003	60.00	0.00	40.00	0.00	45.00	0.00	50.00	5.00	0.00	0.0024	138.00	0.00	6.00	5.80	14.67	34.34	-93.00	-88.28	26.23	22.04
26	BLANK	0.00	0.00	99.00	0.00	99.00	0.00	99.00	0.00	0.00	100000	138.00	0.00	6.00	5.80	14.67	34.34	-93.00	-88.30	26.23	22.02

Figure 3 — Part A: *Star-10* system composite linear dynamic range analysis results anticipate an absolute MDS performance of –132 dBm (–136 dBm was the measured actual result).

System Design

The *Star-10* transceiver features a double conversion approach using a first IF of

75 MHz for good receiver image rejection and a second IF of 9 MHz for achieving ultimate bandwidths of 2.4 kHz, 1.8 kHz and 0.5 kHz. The design allows for baseband DSP to be used after the second conversion. Provisions are made for external spectrum analysis over 0.5 MHz bandwidth at the 75 MHz and 9 MHz IFs. An outboard spectrum analyzer unit can be used for viewing band activity. The entire transceiver's block diagram is shown in Figure 2. This block diagram closely represents the finished product.

As can be seen in Figure 2, the block diagram encompasses both transmit and receive functions. I will focus mainly on the receiver, since the system is bilateral. As can be seen, the receiver system can employ as many as three AGC loops. (Note: A single AGC loop was implemented so far in the hardware, withAIPA and BIPA manually operated). Its behavior was modeled using actual component gains, compression parameters, and ultimate bandwidths, using my specially developed composite dynamic range software entitled Victoria Falls©. This software has the proven capability to ramp up like in real life the input RF at the antenna port all the way from the MDS, up to the system's compression point, turning on all three AGC stages progressively, in reverse order, and graphically displaying the actual dynamic range behavior on a spectrum-analyzer-like color display, proving the entire compression-free composite linear dynamic range performance of over 150 dB.

The results of this analysis are shown in Figure 3A and B. They take into consideration all component parameters shown in the system block diagram from Figure 2. The bottom line composite linear dynamic range results of the analysis are shown graphically in Figure 3C. They anticipate the system's receiver performance from the input to the output as funneled through the system, using the three AGC stages, without compression. (Note: The system's MDS was tested at –136 dBm absolute.) The vertical bands show the three AGC actions necessary to keep the receiver uncompressed over the entire range. Please note how the system's noise figure increases as the RF input is ramped up and the composite AGCs enter the picture. This is normal, as any receiver's noise figure is depreciated by the AGC action, while the signal level is always higher than the receiver's noise figure at any given point on the dynamic range. What is important is that reception is possible with increased noise figure because the signal to noise level is always maintained higher as the signal goes up through the uncompressed dynamic range.

The system design modeling process is usually the most important phase of an entire transceiver design and especially of the receiver design. It is a very tedious process and

STAR-10 ULTRA - HIGH DYNAMIC RANGE COMMUNICATIONS RECEIVER ANALYSIS

System P Input (dBm)	Analysis Temp	AGC 1		AGC 3		
21.00	25 °C	ON -10	10	ON -80	80	Analyze
System P Out (dBm)	**VIEW 600 X 800**	**AGC 2**		**AGC 4**		
38.08	SET SIZE TO 75%	ON -30	30	ON 0	0	Reset
						FALSE Processor Gain 1.00

MDS (dBm)	Linear DR (dB)	ULT BW (Hz)	N Floor (dBm)	Noise Fig (dB)	ICP1dB (dBm)	IIP3 (dBm)	IIP2 (dBm)	System OIM3	SFDR (dBm)
-26.9	47.4	500	-146.9	120.0	20.4	25.4	26.5	27.4	18.3
Pout (dBm)	**Comp (dB)**	**Gain (dB)**	**N Floor (dBm)**	**SNR (dB)**	**OCP1dB (dBm)**	**OIP3 (dBm)**	**OIP2 (dBm)**	**SFDR3 (dB)**	**SFDR2 (dB)**
38.1	0.9	18.0	-8.9	47.1	38.4	43.4	44.5	35.0	26.7

	INPUT DEVICE PARAMETERS @ 25°C											CUMULATIVE PERFORMANCE									
	Device Description	Device Gain (dB)	Delta Gain (dB/°C)	Device CP1dB (dBm)	Delta CP1dB (dB/°C)	Device OIP3 (dBm)	Delta OIP3 (dB/°C)	Device OIP2 (dBm)	Device NF (dB)	Delta NF (dB/°C)	Device BW (MHz)	Gain (dB)	Comp Level (dB)	Signal Power (dBm)	Noise Floor (dBm)	NF (dB)	Cum DR (dB)	IIP3 (dBm)	IIP2 (dBm)	Cum SFDR3 (dB)	Cum SFDR2 (dB)
1	FL HO	-1.00	0.00	99.00	0.00	99.00	0.00	99.00	1.00	0.00	30	-1.00	0.00	20.00	-146.87	1.00	198.23	100.00	100.00	132.15	99.11
2	AIPA AGC1	-10.00	0.00	99.00	0.00	99.00	0.00	99.00	10.00	0.00	30	-11.00	0.00	10.00	-146.87	11.00	187.81	99.59	97.61	125.21	92.92
3	PREAMP	10.00	0.00	40.00	0.00	45.00	0.00	65.00	5.00	0.00	30	-1.00	0.01	19.99	-131.87	16.00	124.23	46.00	65.83	86.15	74.53
4	HMIX	-8.00	0.00	40.00	0.00	45.00	0.00	65.00	8.00	0.00	30	-9.00	0.01	11.99	-139.20	16.67	122.92	45.36	62.97	85.28	72.76
5	DIPLEX	-1.00	0.00	99.00	0.00	99.00	0.00	99.00	1.00	0.00	30	-10.00	0.01	10.99	-140.01	16.86	122.73	45.36	62.92	85.15	72.64
6	SPLITTER	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	1.00	0.00	30	-13.00	0.01	7.99	-142.78	17.09	122.50	45.36	62.94	85.00	72.54
7	FL75A	-3.00	0.00	10.00	0.00	15.00	0.00	20.00	3.00	0.00	0.01	-16.00	0.28	4.72	-144.36	18.51	141.33	30.84	35.62	97.55	75.55
8	BI-SPL1	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	3.00	0.00	30	-19.00	0.28	1.72	-145.43	20.44	139.41	30.84	35.62	96.27	74.59
9	BIPA AGC2	-30.00	0.00	20.00	0.00	15.00	0.00	35.00	30.00	0.00	0.01	-49.00	0.28	-28.28	-146.87	49.00	110.84	30.84	35.59	77.23	60.29
10	BI AMP	36.00	0.00	40.00	0.00	45.00	0.00	65.00	5.00	0.00	30	-13.00	0.28	7.72	-105.87	54.00	105.83	30.83	35.52	73.89	57.76
11	BI-SPL2	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	3.00	0.00	30	-16.00	0.28	4.72	-108.87	54.00	105.83	30.83	35.52	73.89	57.76
12	FL75B	-3.00	0.00	20.00	0.00	25.00	0.00	30.00	3.00	0.00	0.01	-19.00	0.30	1.70	-111.87	54.00	105.63	30.63	33.85	73.75	56.92
13	PAD	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	3.00	0.00	30	-22.00	0.30	-1.30	-114.87	54.00	105.63	30.63	33.85	73.75	56.92
14	MIX TAK3 H	-7.00	0.00	14.00	0.00	24.00	0.00	30.00	7.00	0.00	30	-29.00	0.30	-8.30	-121.86	54.01	105.54	30.60	33.38	73.73	56.69
15	SPLITTER	-3.00	0.00	99.00	0.00	99.00	0.00	99.00	4.00	0.00	30	-32.00	0.30	-11.30	-124.83	54.03	105.52	30.60	33.38	73.71	56.67
16	FL9 2.4	-3.00	0.00	40.00	0.00	45.00	0.00	50.00	3.00	0.00	0.0024	-35.00	0.30	-14.30	-127.81	54.06	111.69	30.60	33.36	77.83	59.75
17	IF 9 AGC3	-80.00	0.00	99.00	0.00	99.00	0.00	99.00	80.00	0.00	0.0024	-115.00	0.30	-94.30	-146.87	115.00	50.75	30.60	33.36	37.20	29.28
18	AD 603 x 2	84.00	0.00	40.00	0.00	45.00	0.00	50.00	5.00	0.00	0.0024	-31.00	0.30	-10.30	-57.87	120.00	45.75	30.60	33.33	33.87	26.76
19	FL9 1.8	-3.00	0.00	40.00	0.00	45.00	0.00	50.00	3.00	0.00	0.0018	-34.00	0.30	-13.30	-60.87	120.00	47.00	30.60	33.30	34.70	27.37
20	FL9 0.5	-6.00	0.00	40.00	0.00	45.00	0.00	50.00	7.00	0.00	0.0005	-40.00	0.30	-19.30	-66.87	120.00	52.56	30.60	33.29	38.41	30.15
21	AD 603 FIXED G	20.00	0.00	40.00	0.00	45.00	0.00	50.00	5.00	0.00	0.0005	-20.00	0.30	0.70	-46.87	120.00	52.56	30.60	33.16	38.41	30.09
22	PAD	-12.00	0.00	99.00	0.00	99.00	0.00	99.00	12.00	0.00	30	-32.00	0.30	-11.30	-58.87	120.00	52.56	30.60	33.16	38.41	30.09
23	FL9 2.4	-3.00	0.00	40.00	0.00	45.00	0.00	50.00	3.00	0.00	0.0024	-35.00	0.30	-14.30	-61.87	120.00	52.56	30.60	33.14	38.41	30.07
24	PD SRA1H	-7.00	0.00	10.00	0.00	23.00	0.00	30.00	7.00	0.00	0.0024	-42.00	0.30	-21.30	-68.87	120.00	52.55	30.60	33.04	38.41	30.03
25	AF TDA 2003	60.00	0.00	40.00	0.00	45.00	0.00	50.00	5.00	0.00	0.0024	18.00	0.92	38.08	-8.87	120.00	47.42	25.43	26.48	34.96	26.75
26	BLANK	0.00	0.00	99.00	0.00	99.00	0.00	99.00	0.00	0.00	100000	18.00	0.92	38.08	-8.87	120.00	47.42	25.43	26.47	34.96	26.74

Figure 3 — Part B: *Star-10* system composite linear dynamic range analysis results anticipate linear performance using all AGCs to +20 dBm. Final performance varied somewhat from this performance, as actual component data changed throughout the design cycle. (See the specifications section in the text.) The program actually ramps up the RF (like in the real life receiver) from the MDS and up through the three AGC ranges (turning them on progressively) until compression occurs.

can take a considerable amount of time. The software used, no matter how sophisticated, can help the designers, but not design the radio for them. This phase of the design is a key part of the design verification methodology mentioned earlier. It sets the system's initial performance goals as close to the final design as possible, so a minimum of modifications will be necessary in the circuit design. Although some designers just string along circuits, no RF system should be pursued without doing this important homework.

With the system's receiver behavior calculated and proven using the software modeling tool, the performance of the design has been further verified and tested (at a final MDS of –136 dBm) in the initial brassboards, and in total concert with the synthesizer's phase noise analysis, brassboarding and tests, which were done separately. A concurrent analysis and breadboarding of the command and control system took place in parallel. Finally, several brassboards and integration of the entire system took place before the final packaging, using all finalized components in progressive order.

A similar analysis was performed in reverse for the transmitting chain, but is not shown here for reasons of simplicity.

With the composite linear dynamic range analyzed, the *Star-10* frequency plan (architecture) was analyzed next, for in-band IF intermodulation distortion (IMD) using my specially designed IMDWEB software. The results of the spurious free performance over the entire 1.8 to 30 MHz range (including receiver image, and higher order spurious products) and using the automatically switched half octave filters are shown in Figure 4A and B. They prove that no significant in-band IF intermodulation distortion products occur at any frequency in the RF frequency coverage with the proper half octave filter switched in, as seen inside the first IF of 75 MHz, and as carried through the second IF of 9 MHz.

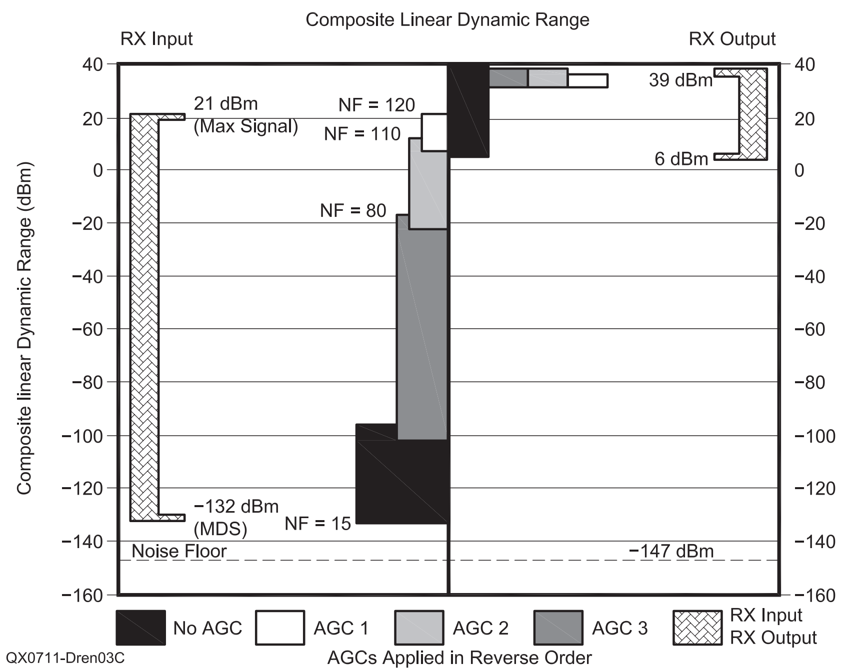

Figure 3 — Part C: Spectrum-analyzer-like graphic results using the software of composite ramping of the RF input over the entire composite linear dynamic range behavior for the *Star-10* system, showing the action of the three AGC stages, and proving the receiver's linear composite dynamic range.

Those versed in the art will recognize that this analysis was carried over to a 16th order (8 × 8 harmonics) of products to ensure further reliability (a 7th order analysis is usually sufficient). For a more in-depth explanation regarding how to read IMDWEB charts, see References 1, 5, 6, 7 and 8.

System Description

I will discuss how the *Star-10* transceiver system works. As mentioned before, this will encompass both transmit and receive functions, but I will focus mainly on the receiver, since the system is generally bilateral. Looking at the system block diagram in Figure 2, the antenna is switched between the receiver and transmitter by the T/R control. An optional phasing type noise-canceling noise blanker unit (such as the ANC-4) can be inserted if needed ahead of the receiver to protect against nearby QRN. As expected, the transceiver is always in the receive mode by default.

Figure 4A

IMDWEB Program Input for the *Star-10* System

Corresponding Numbers on Graph of Part B

		1	*2*	*3*	*4*	*5*	*6*	*7*	*8*	*9*	*10*
Spurs in Band											
MHz/GHz	*Frequencies*	*Band 1*	*Band 2*	*Band 3*	*Band 4*	*Band 5*	*Band 6*	*Band 7*	*Band 8*	*IF2*	*IF3*
F2 (RF)	Cf	2.5	3.5	5	7	10	14	20	27	75	9
	BW 1	1	1	2	4	4	4	6	6	0.5	0.1
	BW 2	0.25	0.25	0.5	1	1	1	1.5	1.5	0.1	0.01
	BW 3	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.01	0.01	0.003
F1 (LO)	Cf	77.5	78.5	80	82	85	89	95	102	84	8.545
F OUT (IF 1)	Cf	75	75	75	75	75	75	75	75	9	0.455
	BW 1	2	2	2	4	4	4	5	6	0.5	0.1
	BW 2	1	1	1	1	1	1	1	1	0.1	0.01
	BW 3	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.01	0.003

Figure 4 — Part A: IMDWEB program inputs for the *Star-10* system. The data is coded and keyed with the graphic results at Part B.

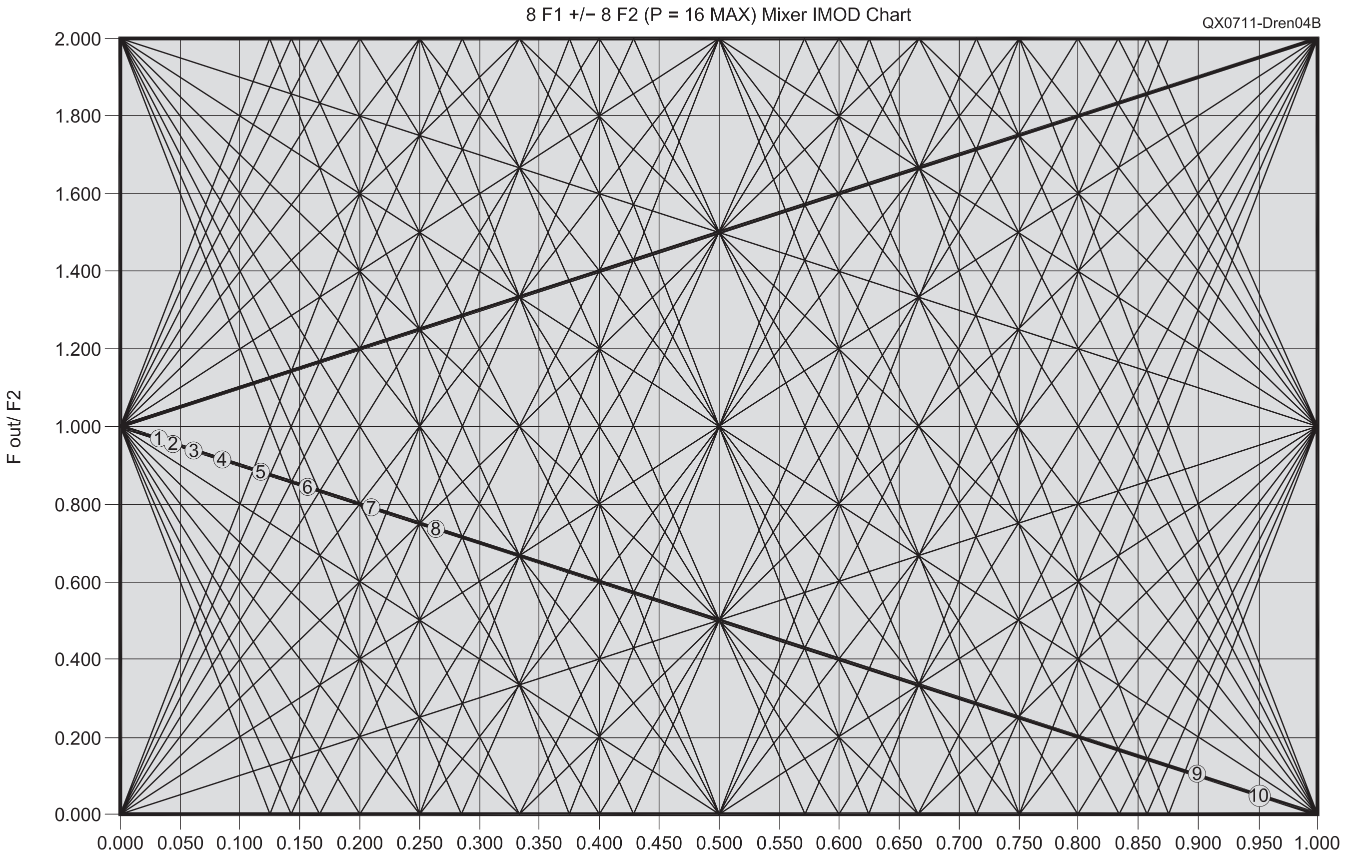

Figure 4 — Part B: The IMDWEB program final results clearly prove that no in-band IF intermodulation product (IMD) lines cross the IFs (the small numbered circles) anywhere in the frequency coverage. The complex IMD results are carried to a 16th order, and show good performance throughout the system's conversions. The frequency ranges are coded to Part A to show the entire system performance in a single plot.

As previously mentioned, the *Star-10* features independent, automatically switched half-octave filter banks for receive and transmit. This is shown at the top left of Figure 2. In the receive mode, RF signals from 1.8 to 30 MHz are automatically selected by the command and control mechanism in the half-octave receiver band-pass filter bank by the DFCB (Command and Control) board as shown, depending on the frequency of operation. The same commands are presented in parallel to a set of high power half-octave low-pass filter banks that have corner frequencies matching exactly the receiver's half-octave filters.

The actual implementation of the automatically switched half-octave receive and transmit filter banks will be discussed in greater detail later.

Automatic frequency selection is achieved anywhere in the frequency range of 1.8 to 30 MHz, providing equal image and spurious rejection in receive, as well as equal harmonic and spurious rejection in transmit anywhere in the frequency coverage.

The 75 MHz first IF puts the receiver image away by 150 MHz at any frequency between 1.8 to 30 MHz. With the proper half-octave filter selected in the banks, the amount of rejection provided is uniform throughout the coverage. Conversely, the proper half-octave low-pass filter selected in the transmit chain insures equal spurious and harmonic rejection throughout the frequency range. (See References 1, 2 and 5.)

Consequently, both receiver and transmitter filter functions are exactly identical, unless operating split over a wide range, in which case appropriate switching between the selected frequencies occurs over the range upon T/R switching. The selection is automatically achieved from the command and control board (DFCB), which also controls the synthesizer (FRU) commands.

The filtered received RF signals from the half-octave bandpass filter bank enter the receiver circuits in the IF75BC board assembly through the advanced intercept point attenuator (AIPA) and the +10 dB push-pull preamplifier located on this board. This combination allows for the programmable AIPA attenuators (part of the AGC control system) to be inserted in the receiver front end. Because of the tremendous dynamic range capability of the *Star-10,* the preamp can be always on. The AIPA functions are implemented via miniature RF relays. Conversely, the transmit chain, when activated, outputs RF signals to the power linear amplifier and further to the high power automatically switched half octave low-pass filter banks, through the RF power transmitter gain control (TGC) and further through the T/R switch, to the antenna.

The front end of the *Star-10* transceiver, IF75BC sets the dynamic range of the entire system as mentioned before. Class A amplifiers operating at 24 V are used in conjunction with a low-noise, high-intercept-point FET push-pull preamplifier using the CP 650, the programmable front end attenuator switched with RF relays, a special high level (class III) H-mode mixer and other hardware. The IF75BC assembly dissipates about 30 W of dc power to insure the high dynamic range for the receiver. Two brushless miniature fans extract heat from the amplifiers through heat sinks. The board is housed in a machined aluminum box with cutouts for command

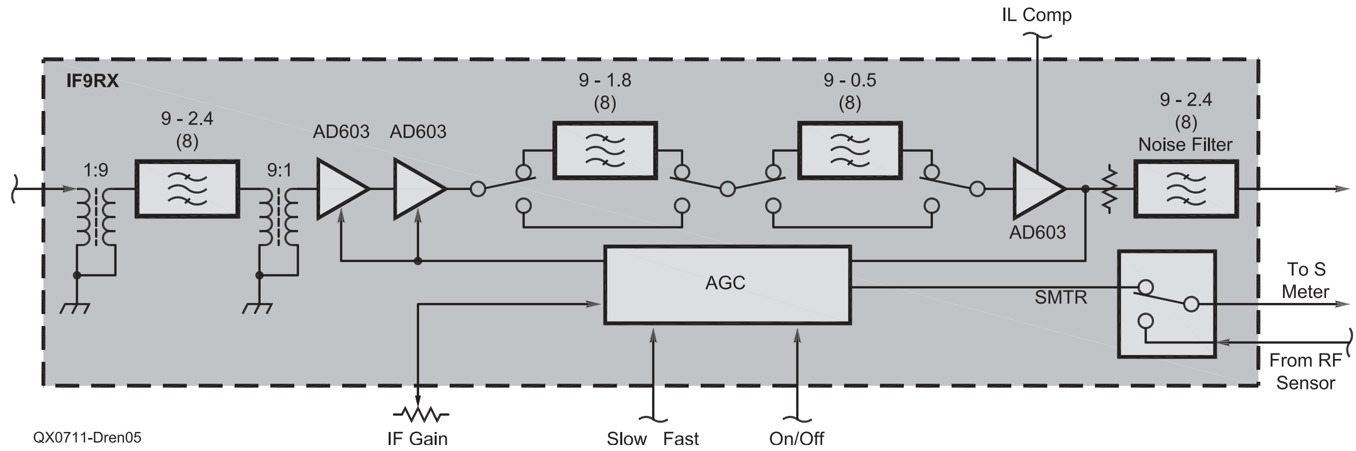

Figure 5 — IF9RX uses a cascaded filter selection of the ultimate receiver bandwidth, depending on mode and bandwidth (wide or narrow) choice, as commanded by the command and control (DFCB) and keypad assemblies.

and control connectors and SMA connectors for the RF ports.

The IF75BC assembly creates the first IF at 75 MHz and works in receive as well as transmit. It also contains a low-pass filter and diplexer-splitter circuitry. Part of the 75 MHz IF information is directed to the crystal-roofing filter for further processing. The other half, which is 500 kHz wide, is directed to the IF9BC board for further conversion and to IF9NB for spectrum analyzer and noise blanker functions.

The 75 MHz roofing filter is bilateral and is made of two, four-pole sections with a 3 dB bandwidth of ~10 kHz. These filters have been expressly designed for the *Star-10*. They exhibit high intercept points and can withstand the RF levels (up to +5 dBm) present in the system at this point in the receiver system over the entire dynamic range. For better signal handling, the roofing filters have been distributed before and after the bilateral amplifier assembly BILAT AMP, as shown.

The 75 MHz roofing filter assembly is followed by the bilateral amplifier (BILAT AMP) assembly. This assembly allows 75 MHz signals to pass automatically either way (receive or transmit) by merely switching the 24 V power distribution to it from IF75BC, which in turn is performed by the T/R assembly. The circuit is unique because there is no need for hard switching of RF inputs and outputs, due to the automatic rejection of unwanted RF paths provided by the natural isolation of the unused sides of the splitters/combiners at the input and output of the amplifiers. The BILATAMP amplifiers are high gain (+36 dB), high intercept, class A types similar to those used in the IF75BC assembly. A similar brushless miniature fan is used for cooling here. Only one amplifier is on at a time, allowing for cooler operation. The *Star-10* makes ample use of passive splitters and combiners for its bilateral circuitry. The paths not used provide some 30 dB of natural isolation to the used paths. Additional switching has been found necessary in addition to this isolation to provide complete muting in the IF9RX circuitry.

The FL75 roofing filters and BILAT AMP assemblies are followed by the IF9BC assembly. This assembly is equipped with the second AGC loop, called BIPA, which allows for 30 dB of adjustable gain action from the front panel RF/IF gain control. The adjustable BIPA attenuator as well as the transmit CW drive circuits use a classic PIN attenuator circuit, which will be discussed later. The 75 MHz IF signals coming from the BILAT AMP assembly are coverted here to the 9 MHz IF (500 kHz wide IF) for the IF9NB and for the main 9 MHz receive IF assembly called IF9RX.

In addition, the 9 MHz transmitter IF, IF9TX, is also input to the IF9BC assembly using a similar passive splitter/combiner technique as previously used in the BILATAMP. The IF9BC assembly uses high-level class II mixers to perform the conversions. A 500 kHz wide, 9 MHz IF filter is used to condition the spectrum analyzer and noise blanker functions of the IF9NB. The ultimate bandwidth for the receiver is established through the crystal filter bank in the IF9RX assembly. The bandwidth for the SSB/AFSK transmit functions is established through a similar filter bank in the IF9TX assembly.

As can be seen from Figure 2, the IF9BC main receiver output is further input to the IF9RX assembly. This IF achieves the ultimate receiver bandwidth selection and amplification as commanded by the command and control assembly DFCB. The IF9RX board provides 100 dB of gain (80 dB AGC) using three high dynamic range (high IP3) logarithmic/linear IF blocks from Analog Devices. The IF bandwidth selection is provided by four 8-pole crystal filters that were specially designed for the *Star-10*. Instead of selecting individual filters like conventional IF designs, the *Star-10* IF9RX filter assemblies are combined in a cascaded AND function (rather than an OR function) for a total of 32 poles (plus the 8 pole roofing filter) of superb selectivity. This cascaded architecture makes the IF9RX a unique design that works in tandem with the system's command and control software. This is shown in Figure 5.

The eight pole crystal filters are configured in a cascaded configuration for increased selectivity for a minimum of 16-pole and a maximum of 32-pole selectivity (in addition to the 8 pole roofing filter). The first and last 2.4 kHz filters set the maximum IF bandwidth of 2.4 kHz while the 1.8 kHz and 500 Hz filters set narrow selections for different modes depending on the mode selected from the command and control. Three AD603 logarithmic linear amplifiers are used to provide ~100 dB of gain (80 dB AGC), with the third amplifier used to compensate for narrow filter insertion loss and equal AGC/S-meter indications regardless of the filter combination. The last 2.4 kHz filter is used to clean up noise from previous amplifiers.

As shown in Figure 5, two 8 pole crystal filters with a bandwidth of 2.4 kHz are always used at the beginning and the end of the 9 MHz IF chain for good noise management. Then, additional 8 pole crystal filters of narrower bandwidths are inserted or removed between the gain stages (for a maximum of 32 poles in CW Narrow) depending on the mode selection and as commanded by the DFCB. The selection is achieved with miniature RF Teledyne relays, just as in the front end of the radio (no PIN diode switching for RF paths in this radio). Automatic insertion loss compensation control is achieved depending on the diverse filters configurations so there is no difference in signal amplitude when changing filters and bandwidths.

Because of the 100 dB gain provided by this important board, and the limited board size of 5.5 × 4.5 inches, the IF9RX board had to be specially laid out to prevent possible oscillation. The first layout did oscillate. A special effort was made by KD7KEQ to provide a new layout, using hundreds of plated through ground stitches in the double-sided board ground planes, to make the system as quiet as technically possible. Additional effort was put into its execution of this board by KG6NK, making this demanding board perform, as it should.

Conversely, the IF9TX assembly provides for processed SSB signals and all other transmit functions supplied to the IF9BC. Microphone amplification and compression are provided together with SSB mixer, CW drive, carrier insertion, transmitter gain control (TGC) feedback, and switching functions using Hittite solid-state RF switches. In addition, SSB transmit bandwidth and on the air "sound character" are set by two crystal filters for a total of 16 poles of 2.4 kHz bandwidth, similar to those in the IF9RX. This allows for a true and clean communication sounding SSB transmission.

Finally, the receiver chain is completed by feeding the IF9RX output to the product detector PDAF assembly. Here, the 9 MHz coherent BFO signal coming from the synthesizer (FRU) enters the product detector mixer using a high level class II device, to be demodulated by the mode commands received from the DFCB assembly. The BFO frequencies used by the product detector are shown in Table 1, along with all other LO interactions for setting up the system in all the available modes of operation. The functions are entered through the KB1 keyboard, processed through the DFCB and output by the FSYNT assembly. As can be seen, the commands change with the T/R functions, so the net result is that we transmit exactly on the same frequency as we receive, regardless of the operating mode. The functions are selected automatically by DFCB and are subject to the modes selected.

The filtered and AGC conditioned IF9RX output is finally presented to the PDAF product detector assembly. Here, the 9 MHz signals from the IF are mixed in a high-level class II mixer (the product detector) with the high level (another class A amplifier is used here) BFO LO signal coming from the coherent synthesizer FSYNT. An audio low-pass filter further cleans up the resulting audio signal before being amplified and output to the DSP and/or speaker. Additional audio beeps corresponding to commands coming from the command and control assembly DFCB are audio mixed and presented to the audio stages. Muting signals from the T/R assembly are fed concurrently to PDAF and IF9RX as well as to other points in the receiver chain.

In addition to the muting function, the T/R assembly combines conditioned keying and PTT/VOX signals received through the command and control assembly DFCB, to perform the total transceiver control functions. For example, the T/R uses a unique method of digitally generating slight delays (the Morse code is shifted through a shift register) in the CW keying path, to allow the synthesizer to settle and lock in QSK, before Morse code characters are shifted out through the transmitter. This helps to stop possible chirping when using the extra wide split frequency capability of this transceiver and also between Morse code elements, making for clean CW if operating on two separate frequencies, and even between the elements of CW signals. The T/R assembly is fully digital and will be discussed later.

We will now discuss the command and control DFCB assembly in conjunction with the frequency reference unit (FRU) FSYNT assembly, along with the master reference unit (MRU) assembly.

The heart of the *Star-10* is the command and control assembly DFCB, which works in conjunction with the keypad assembly and the RS-232 interface. These assemblies are physically installed together with the displays behind the front panel of the transceiver as shown in Figure 2. The transceiver's entire capability is slaved to a powerful 8-bit Microchip PIC-17C44 microprocessor controller that runs approximately 10,000 lines of code continuously at 32 MHz (Note: chosen above the HF range to keep possible spurious RF products out of the receiver's input range) in a closed loop, only to be interrupted by its keypad or RS-232 commands.

I initially used the UV erasable PIC-17C44 version, offered in the in-line package for optimal prototype development. This allowed for multiple UV erasable reprogrammed software versions (at least 100 revisions) with the latest V.3.1.

A Few Words About The Chosen Microprocessor

The PIC-17C44 microprocessor operates at up to 33 MHz with full interrupt capability. I am using it at 32 MHz to put any possible spurious problems above the HF band. The PIC-17C44 has an instruction cycle of 125 ns. It is equipped with 33 I/O ports (all have been used), and 16 levels deep hardware stack plus 64K × 16 addressable program memory space.

The PIC-17C44 microprocessor is a high speed CMOS, fully-static protected, 8-bit microcontroller employing an advanced RISC architecture. It has enhanced core features, 16-level deep stack, and multiple internal and external interrupt sources. The separate instruction and data buses of the Harvard architecture allow a 16-bit wide instruction word with a separate 8-bit wide data word. The two-stage instruction pipeline allows all instructions to execute in a single cycle. A total of 55 instructions (reduced instruction set) are used. Additionally, a large register set gives this microprocessor some new architectural innovations used to achieve a very high performance. For mathematically intensive applications such as used in this application, the device has a single cycle 8 × 8 Hardware Multiplier.

PIC-17C44 microcontroller typically achieves a 2:1 code compression and a 4:1 speed improvement over other 8-bit

Table 1

Effects of Mode Selection on System Frequency Compensation of Local Oscillators

This includes the BFO (LO3) in the PDAF provided by the FSYNT, and as commanded by the Command and Control board, DFCB and the Keypad.

MODE	*LO1 RX*	*LO1 TX*	*LO2 RX/TX*	*LO3 RX*	*LO3 TX*	*SHIFT/RX-TX*
USB	UP 1500 Hz	UP 1500 Hz	84 MHz	8.9985 MHz	8.9985 MHz	1500 Hz
LSB	DWN 1500 Hz	DWN 1500 Hz	84 MHz	9.0015 MHz	9.0015 MHz	1500 Hz
CWU	UP 800 Hz	UP 800 Hz	84 MHz	8.9992 MHz	9.000 MHz	1500 Hz
CWL	DWN 800 Hz	DWN 800 Hz	84 MHz	9.0008 MHz	9.000 MHz	1500 Hz
FSKU	UP 2210 Hz	UP 2210 Hz	84 MHz	8.99779 MHz	8.99779 MHz	NA
FSKL	DWN 2210 Hz	DWN 2210 Hz	84 MHz	9.00221 MHz	9.00221 MHz	NA

microcontrollers. The PIC17C44 has up to 454 bytes of RAM and 33 I/O pins. In addition, the PIC17C44 adds several peripheral features useful in high performance applications including (not all utilized in this application):

- Four timer/counters
- Two capture inputs
- Two PWM outputs
- A universal synchronous asynchronous receiver/ transmitter (USART)

These special features reduce external components, thus reducing cost, enhancing system reliability and reducing power consumption. We will discuss the command and control DFCB assembly later.

The *Star-10* has been designed with a very flexible and friendly human interface that has been compared with the feel of classic HP test equipment. This functionality did not come easy and has taken a considerable amount of time and dedication to design and prove. I spent approximately a year and a half in the complex system and command interface requirements and implementation, investment that resulted in a "bug free" design.

The command and control interface DFCB was designed using on board EEPROM

Command and Control Calculations

One of the main governing equations for the front panel display is:

$$f_{Display} = \frac{(f_{LO} - f_{IF})}{10}$$

As an example, for an f_{IF} of 75,000,000 Hz (75 MHz) and a local oscillator output of 89,240,110 Hz, the control software will calculate the operating frequency to display:

$$f_{Display} = \frac{(89{,}240{,}110 - 75{,}000{,}000)}{10}$$

$$f_{Display} = 14{,}240{,}110$$

In this case, the *Star-10* display will show 14.240.11, representing an operating frequency in the 20 m band. Figure A shows the relationships between these various signals as they are processed.

There are, of course, many other calculations taking place in the Command and Control assembly. The following are a few examples of calculations related to how the microprocessor deals with the tune frequencies as related to the DDS.

$$f_n = \left(\frac{2^{32}}{100\times10^6}\right) \bullet \left(\frac{75\times10^6 + Display}{10}\right)$$

$$f_n = \left(\frac{2^{32}}{100\times10^7}\right) \bullet 75\times10^6 + \left(\frac{2^{32}}{100\times10^7}\right) \bullet Display$$

Where f_n is the Tune Value.

Then:

$$f_n = 322122547.2 + 4.294967296 \bullet Display$$

For example, a Display value of 29.999990 × 10^6 yields:

f_n = 450971523

$$DDS_{OUT} = \frac{f_{Desired} + f_{IF}}{10}$$

And

$$DDS_{OUT} = \frac{DDS_{Word} \bullet f_{REF}}{2^{32}}$$

Let

$$f_{KBD} = \frac{f_{Desired}}{10}$$

$$\frac{DDS_{Word} \bullet f_{REF}}{2^{32}} = f_{KBD} + \frac{f_{IF}}{10}$$

$$DDS_{Word} = \frac{2^{32}}{f_{REF}} \bullet f_{KBD} + \frac{2^{32}}{f_{REF}} \bullet \frac{f_{IF}}{10}$$

Let

$$f_{REF} = 84\times10^6$$

$$f_{IF} = 75\times10^6$$

Then

$$DDS_{Word} = 51.13056304762 \bullet f_{KBD} + 383479222.8571$$

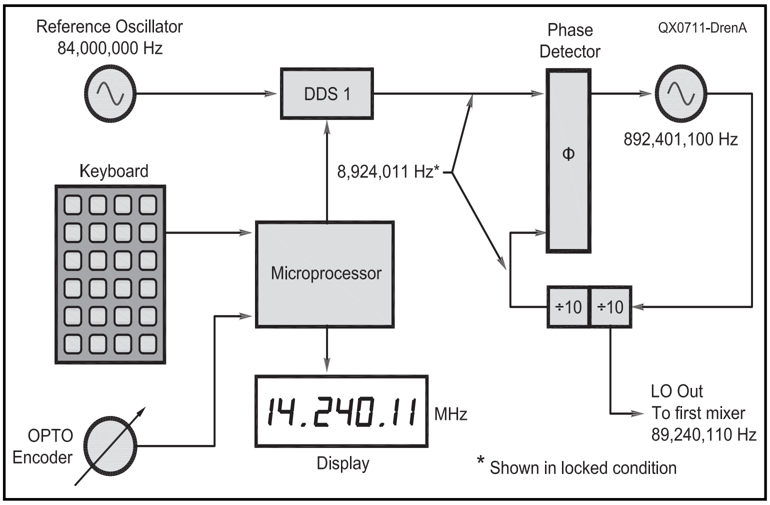

Figure A — This partial block diagram shows a simplified version of the frequency generation section of Figure 2. This diagram illustrates how the microprocessor and 84 MHz reference frequency control the DDS1 signal going into the PLL phase detector as well as displaying the radio operating frequency. The PLL output is the LO signal to the first mixer.

memory. I/Os were implemented through the custom keypad (which was built from scratch) as augmented by an opto-encoder. Up, down and direct frequency inputs, mode select, bandwidth select, split, RIT and IF Shift memory functions are addressed directly from the keypad and the opto-encoder. The opto-encoder (main knob) also has a push-push function to select two brightness levels for the integrated display and the sound feedback signals are audible through the audio amplifier.

The *Star-10* system power up default is 10.00000 MHz — WWV-USB mode, for zeroing the entire system's accuracy from the MRU. Upon turning on the power, the display shows 10.00000 MHz along with the *Star-10* logo and the software version (V.3.1). The receiver is set up by default to USB and if an antenna is connected, WWV signals are heard for initial calibration. After this imposed calibration on the system (at each power up), the operator inputs a new operating frequency to the last 10 Hz via the keypad. See the lead photo for a picture of this keypad. There are no "bands" on the keypad. This transceiver is programmable from 1.8 MHz to 30 MHz in one band, with 10 Hz resolution. The front end filtering selection is executed seamlessly and follows automatically behind the scene through the automatic switching half-octave filter banks.

The main display is facilitated through two 32-character, back-lit, green-blue integrated LCD-Twist dot matrix displays. (Note: this provides a total of 64 characters on 4 lines.) Slight frequency changes and RIT/IF-SHIFT intervention are achieved via the opto-encoder in conjunction with the keypad as the human interface. In addition, the main knob activates brightness and sound functions through the microprocessor.

One of the key functions of the command and control assembly is to control the frequency synthesizer (FRU), FSYNTH. This unit uses two coherent loops in conjunction with the MRU reference operating at 84 MHz, which is also used as a fixed second LO. (See References 9 and 10.) There are two AD-9850 DDS circuits to be controlled. The first DDS is used in a microwave DDS-Driven PLL (see Reference 10) system that was previously described in References 3 and 4. This loop operates in 1 Hz increments from approximately 0.7 GHz to 1.05 GHz and is divided by 10 for improved phase noise performance for the first variable LO. The highly filtered second DDS is used as a 9 MHz BFO providing the various product detector frequencies from Table 1. The 84 MHz MRU LO serves as both, a reference for the two DDSs as well as a fixed LO for the second conversion. Thus, a fully coherent system results.

The command and control DFCB system is capable of addressing either the main loop DDS or the BFO DDS. The main synthesizer loop (the DDS-Drive PLL microwave loop) is controlled through direct keypad entry as taken over by the opto-encoder. When in the mode select mode, the keypad controls the microprocessor such that the BFO/DDS-2 follows a fixed programmed function/frequency offsets from the nominal 9 MHz and as changed by the USB, LSB, CW, CWN, AFSK command requirements. This programmability along with the entire transceiver's frequency sources programmability was previously shown in Table 1.

Up/Down arrow commands are used on the keypad to enter RIT and PBT offsets. The passband function (marked "SFT" on the keypad) allows selected TX or RX offsets to vary ± 1.5 kHz moving the IF BW and other sources in either side of the zero in either transmit or receive. Once set, the IF PBT remains memorized during the power-on session, to be reset to its nominal values by the power "off" switch until the next power-on session begins and a new "SFT" entry is input. The RIT function allows for ± 9.9 kHz received frequency offset from nominal and gets reset to nominal zero with transceiver power off.

After the power up and the 10 MHz calibration mode appears, the operator enters the frequency of interest via the keypad in VFO A. This frequency is the receive and transmit frequency for the transceiver, unless choosing to operate split. It can be fine tuned with the main tuning knob or the up/down arrow buttons, with addressable resolution as well as by the RIT and PBT keypad inputs. If choosing to hold either the UP or the DWN buttons for more than a second, a scanning function from the nominal displayed frequency is achieved. Touching any other keypad button can stop the scanning. By pushing the split button on the keypad, a second frequency, VFO B can be entered within the transceiver's entire frequency coverage. The new frequency (VFO B) shows up on the second row of the frequency display as shown in the lead photo. The R>T and A>B buttons change/reverse the addressability of the two VFOs.

VFOs A and B are virtual VFOs, since the same synthesizer is used to generate them. Hams are usually taught to think that there are actual separate VFOs in synthesized radios. In reality, this is far from truth. The virtual VFO functionality using a single synthesizer is much the same here as in most synthesized transceivers on the market today.

Additional keypad inputs MODE and W/N keys select the mode and bandwidth requirements. These functions automatically correct the LO settings, so bandpass frequency centers and appropriate bandwidths are selected in the IF9RX and IF9TX to provide seamless operation on the exact same frequency with the station at the other end.

More flexibility is provided through a linear scale frequency indicator showing on the last row of the display as shown. This is visible in Figure 2. The DFCB also provides for scanning functions as well as 99 memories. Finally, the keypad can be totally locked up through the LCK function button.

The frequency synthesizer (FRU) DDSs are commanded by DFCB through serial communication lines. The serial communication speed is sufficient to allow for a proper human interface. Access time is in the microseconds after the microprocessor has been speeded up to the 32 MHz closed loop operation. It should be noted that initial microprocessor clock speeds were progressively increased from the initial 4 MHz to the current 32 MHz as the system grew in complexity. As we did not initially know what size software we would end up with, running approximately 10,000 lines of code in a continuous loop proved to be too slow for interrupt interaction compatible with human operator reactions. Thus, the 32 MHz resulted. The DFCB uses its own crystal oscillator. Some of the governing formulas for the system's interface are shown in the sidebar.

The DFCB commands are presented to the synthesizer, FSYNTH. The synthesizer translates these commands into variable and fixed frequencies using the DDSs and the microwave loop operating from 0.75 GHz to 1.05 GHz as locked to the fixed 84 MHz crystal reference of the MRU (master reference unit). The MRU is a separate assembly. It uses a tight tolerance, 0.001% quartz crystal in a Colpitts PLXO (phase locked crystal oscillator) circuit to provide a close-in phase noise performance of better than –165 dBc/Hz. The 84 MHz crystal is further locked in the MRU to a 10 MHz oven controlled crystal oscillator (OCXO), which provides the long-term stability of 1×10^{-8} for the entire radio, after a 30 second warm-up time. The 10 MHz source is for high stability, while the 84 MHz source is to insure good phase noise performance. The MRU is powered up once and could be left on even during transceiver power off. In the present implementation, it is powered together with the rest of the radio. Its warm-up time (which is the radio's warm-up time to 1×10^{-8}) is 30 seconds. A more detailed description of the MRU will be presented later.

The MRU frequency is used to reference the two DDSs in the FRU, as well as serves as the second fixed LO for the radio. A single high purity microwave VCO is used in the microwave PLL of the FRU. The synthesizer description and operation has been presented in References 3 and 4. After continued improvements in the loop bandwidth versus lock-up trade offs and additional dc filtering, a –133 dBc/Hz close in phase noise performance has been realized at the LO1 injection point. This performance will be discussed further in Part two of the series. As phase noise translates directly into the system IFs on a dB per dB bases, this performance is fully compatible with the MDS and dynamic range expected of the receiver. Comprehensive DR tests against top of the line transceivers were made in the KG6NK laboratory using state-of-the-art test equipment. More on this performance will be presented later.

This concludes Part 1 of this article. In Part 2, I will discuss major assemblies and circuit design/development issues for the *Star-10* transceiver blocks. Pictures and operational discussions of the blocks will also be introduced. Total system integration and performance tests will also be presented in this series.

Cornell Drentea, KW7CD, took his first radio receiver apart (and put it back together) at the early age of six. He has been a ham since 1957. Since then, he's built many radios and transceivers and made his passion for designing "radios" his lifelong profession. As an Amateur Radio operator, he is known for his extensive RF technology articles in magazines such as ham radio, Communications Quarterly, RF Design, *and* QEX.

Professionally, Cornell is an accomplished RF technologist, an engineer and a scientist with over 40 years of hands-on experience in the aerospace, telecommunications and electronics industry. He has been involved in the design and development of complex RF, radar, guidance and communications systems at frequencies of up to 100 GHz. Cornell has developed several state-of-the-art RF products including ultra wide band high probability of intercept microwave receivers, complex synthesizers, multi-modulation transmitters, Doppler agile space transceivers as well as high power RF linear amplifiers. He received his formal education abroad with continuing studies and experience achieved in the United States.

Cornell has presented extensively on RF design topics at technical forums such as IEEE, RF-Expo, Sensors-Expo and has given comprehensive professional postgraduate courses in RF receiver design, synthesizer design, sensors and communications. He has published over 80 professional technical papers and articles in national and international magazines. He is the author of Radio Communications Receivers, *McGraw Hill, ISBN 0-8306-2393-0 and ISBN 0-8306-1393-5, 1982, and holds five patents. He is currently available for consulting to large and small RF enterprises. You can find out more about Cornell, his consulting and his RF course offering entitled* The Art of RF System Design *on his Web site:* **http://members.aol.com/cdrentea/myhomepage/**

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A Direct-Reading Reflection Coefficient and Power Meter

This easy to construct piece of test instrument combines power measurement and reflection coefficient measurement capabilities.

Ralph Gaze, W1RHG

W1RHG

Introduction

This easy-to-build battery-powered instrument provides a direct digital indication of the reflection coefficient (in dB) of an unknown impedance, or of the applied power in dBm, at frequencies from audio to UHF. The reflection coefficient is measured in conjunction with an external fixed or swept signal generator. The meter provides a convenient indication of the reflection coefficient of active or passive networks and antennas. Among other applications, the meter provides a very simple and useful way to measure the loss of a device or a transmission line *from one end*! Within the frequency response limitations of the instrument, it offers a means of measuring the scalar S-parameters of a 4-port network. While an instrument capable of measuring the complex values of the reflection coefficient would provide a much more comprehensive network analysis tool, the simplicity of this meter provides a useful compromise. Larry Coyle's instrumentation amplifier digital voltmeter module (DVM) driver has been incorporated to allow powering with a single battery, and an interval timer has been provided to avoid inadvertent battery discharge.[1] The return loss bridge on which this meter is based was presented by Zach Lau, KH6CP/1.[2]

Numerous articles have described RF power meters based on the Analog Devices AD8307 logarithmic detector.[3, 4] Wes Hayward, W7ZOI, and Bob Larkin, W7PUA, started the ball rolling with their June, 2001 *QST* article, using an analog voltmeter readout.[5] Bob Kopski, K3NIH, followed in the May/June 2002 *QEX* and the Sept/Oct *QEX* with a version using a digital voltmeter module.[6, 7]

Perhaps following the suggestion in Hayward and Larkin, Thomas Scherrer, OZ2CPU, described construction of a PIC-based RF Wattmeter in the October, 2002 issue of *Elektor Electronics*,[8] followed by Roger Hayward, KA7EXM, in the May/June 2005 *QEX*.[9] Bob Kopski, K3NHI, presented an RF power calibrator in the Jan/Feb 2004 *QEX*.[10]

Reflection Coefficient

Reflection Coefficient (RC) or Return Loss (RL) characterizes an impedance with respect to a reference or characteristic resistive impedance. RC comes to us from optics and electrodynamics, while RL comes from telephone usage. The definitions are essentially identical, except for a (inconsistent) difference of sign; RC is usually given as negative. RC is familiar as a radial scale on the Smith chart, and as reflection parameters S11 and S22. The Appendix gives further details of RC, RL, and SWR.

Zack Lau's Return Loss Bridge

Figure 1 shows Zack Lau's return loss bridge, as originally presented. The voltage across R4, for a signal generator open-circuit voltage of V_o and an unknown impedance $\mathbf{Z}_x$ can be derived as:

$$V_{Bridge} = \frac{V_o}{8} \bullet \left| \frac{\mathbf{Z}_x - 51}{\mathbf{Z}_x + 51} \right|$$

[1]Notes appear on page 23.

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The second term is RL expressed as a fraction, as described in the Appendix. The voltmeter reading with $\mathbf{Z}_x$ connected is noted, a short-circuit termination is substituted for $\mathbf{Z}_x$, and an external step attenuator in the signal generator line is set to give the same voltmeter reading. RL is then equal to the attenuator setting.

Circuit Description

Figure 2 shows the circuit of the revised bridge and detector. The differential input AD8307 lends itself to use as a bridge detector to measure the logarithm of the bridge voltage, with the multiplicative term V_o / 8 appearing as a dc offset in the output.

While a 50 Ω bridge is shown, changing R1 through R4 and the 3-dB pad, R5, R6, and R7 would convert it to 75 Ω or other impedances. The pi-network pad may be replaced by a minimum-loss pad (described later) to match a 50 Ω signal generator to the 75 Ω bridge.

If a signal is applied to J1, the detector will measure the applied power in dBm, again with an appropriate zero adjustment to compensate for the bridge loss.

The detector's output voltage will be scaled to 10 mV per dB to provide a direct RC or power indication, and RC or power zero adjustments will compensate for the bridge loss and the signal generator level. The power fed to the unknown impedance at J1 is 9 dB below the signal generator level applied to J2.

Figure 3 shows the unity gain DVM driver isolation amplifier, U3, and the DVM reference buffer U4A, switch-selectable zero adjustments for the RC and power modes, the analog output buffer U4B, and the low-dropout 5 V dc regulator, U2. The DVM is strapped for 9 V dc operation, 2.0 V dc full scale, and for one digit to the right of the decimal point. The analog output at J3 may be used with an oscilloscope to make swept measurements of power or RC. Amplifiers U3 and U4 were chosen for their rail-to-rail input and output, and could be replaced with various other operational and unity gain instrumentation amplifiers with the same feature.

The rail-to-rail op amp, U4, is a legacy from an earlier version of the meter. If you connect the 'U4B' pin 5 input to the U3 output, pin 6, the rail-to-rail requirement goes away. In that case an op amp such as the TL082 will work for U4.

Figure 4 shows the power timer, with an IRF7314 dual P-channel MOSFET, Q1A, protecting U5 against an inadvertent reversed battery connection, and Q1B serving as a power switch, as timed by U5, a CD4060B (not an HC4060) oscillator and binary divider. With the values shown for R22, R23, and C9, the oscillator operates at approximately 27 Hz, and the desired on-time is set by selecting the appropriate U5 output pin: pins 1, 2 and 3 correspond to 5, 2.5, and 1.25

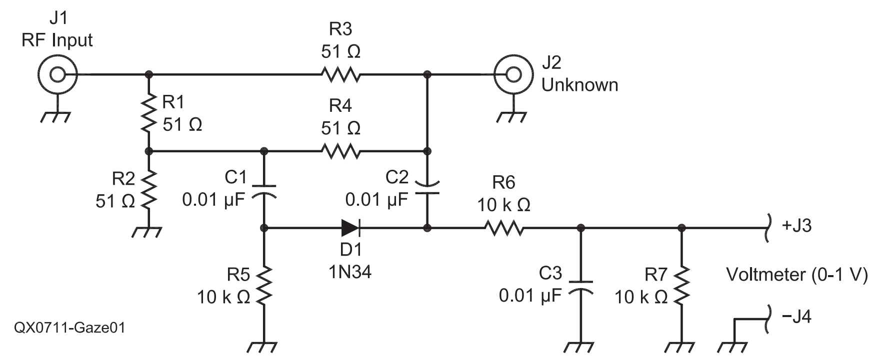

Figure 1 — This schematic diagram shows Zach Lau's return loss bridge.

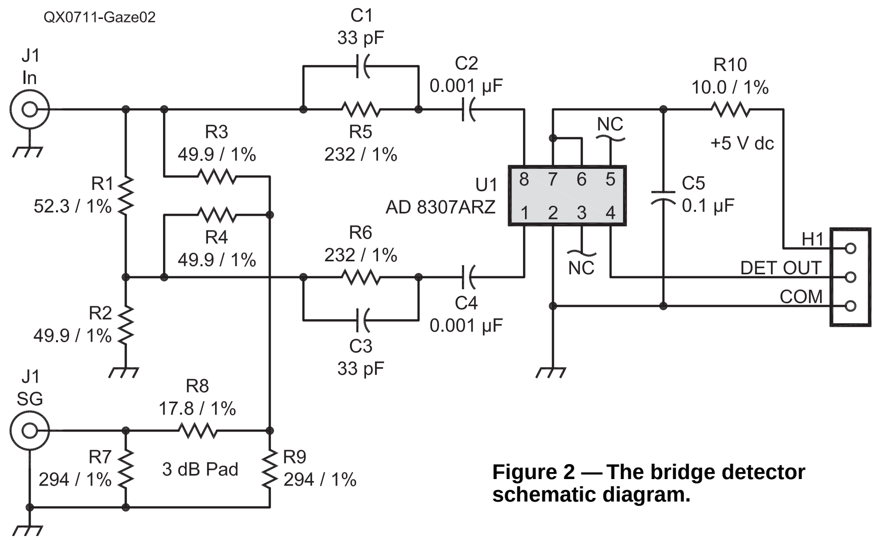

Figure 2 — The bridge detector schematic diagram.

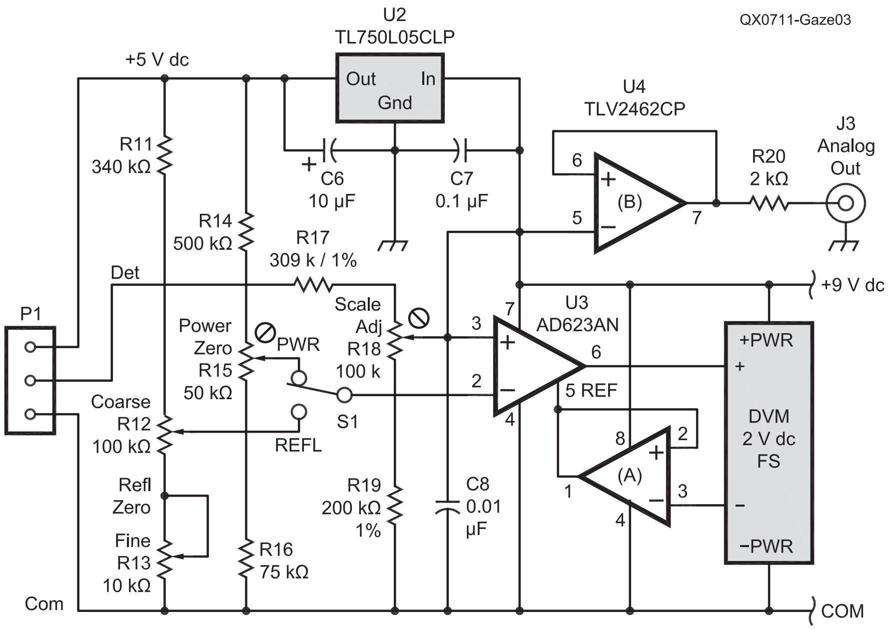

Figure 3 — The meter driver schematic.

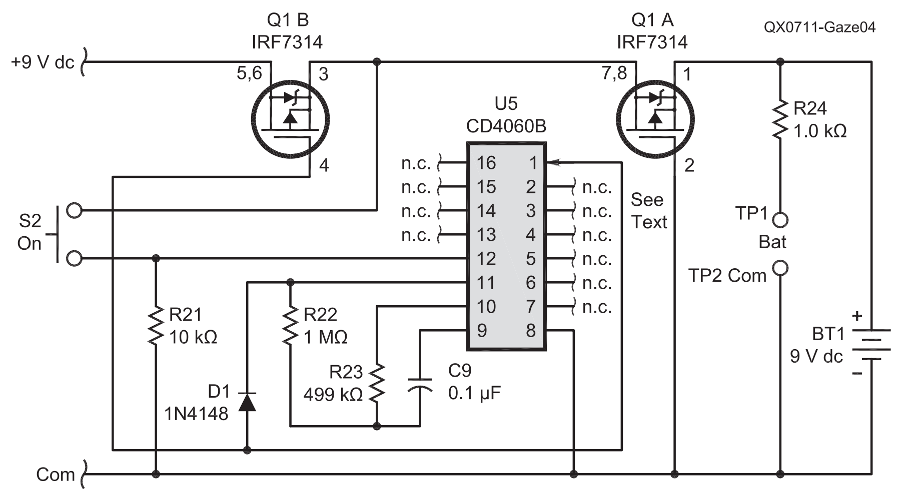

Figure 4 — This schematic diagram shows the power timer circuit.

W1RHG

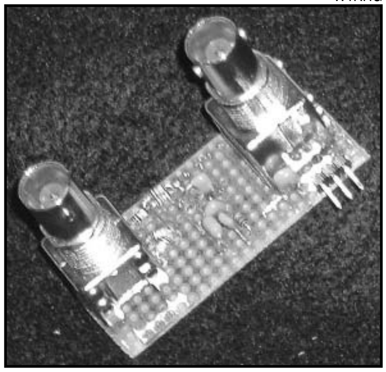

Figure 5 — This photo shows the construction of the bridge-detector assembly.

Table 1
Parts List

Qty	*Title*	*Detail*	*Reference*	*Vendor*	*Vendor P/N*	*Mfr P/N*
1	Battery	9 V alkaline	B1			
2	Capacitor, SMD	33 pF, 50V NPO, 1206	C1, 3	Digi-Key		
2	Capacitor, SMD	1nF, 50V, 1206	C2, 4	Mouser	77-VJ1206A102JXACBC	
3	Capacitor	0.1 uF, 50 V	C5, 7, 9	Mouser	581-SR215C104KAR	
1	Capacitor	10 uF/ 10 V tantalum	C6	Mouser	581-TAP106K010SRW	
1	Capacitor	0.01 uF, 100 V	C8	Mouser	581-SR151C103KAR	
2	Connector	BNC panel jack	J1, 2	Mouser	571-2276731	AMP 227673-1
1	Connector	BNC panel jack	J3	Mouser	571-2277552	AMP 227755-2
1	Digital voltmeter module	3½ digit LCD, 1.999Vdc FS	DVM	MJPA **www.mjpa.com**	16177-ME	
1	Diode	1N4148	D1	Mouser	78-1N4148	
1	Header	3 contact, rt. Angle, 0.1 in	H1	Mouser	538-22-28-8060	
1	Plug housing	3 contact, rt. Angle, 0.1 in	P1	Mouser	538-22-01-2037	
3	Plug housing contact		N/A	Mouser	538-08-55-0102	
1	Potentiometer	100k ohm, linear, ½ W	R12	Mouser	31VA501-F	
1	Potentiometer	10k ohm, linear, ½ W	R13	Mouser	31VA401-F	
1	Potentiometer, trimmer	50k ohm	R15	Mouser	72-T93YA-50K	
1	Potentiometer, trimmer	100k ohm	R18	Mouser	72-T93YA-100K	
1	Resistor, axial lead	52.3 ohm, ⅛ W, 1%	R1	Mouser	270-52.3-RC	
3	Resistor, axial lead	49.9 ohm, ⅛ W, 1%	R2, 3, 4	Mouser	270-49.9-RC	
2	Resistor,SMD	232 ohm, ⅛ W, 1%, 1206	R5, 6	Mouser	290-232-RC	
2	Resistor, axial lead	294 ohm, ⅛ W, 1%	R7, 9	Mouser	270-294-RC	
1	Resistor, axial lead	17.8 ohm, ⅛ W, 1%	R8	Mouser	270-17.8-RC	
1	Resistor, axial lead	10.0 ohm, ⅛ W, 1%	R10	Mouser	270-10.0-RC	
1	Resistor, axial lead	340k ohm, ⅛ W, 1%	R11	Mouser	270-340k-RC	
1	Resistor, axial lead	499k ohm, ⅛ W, 1%	R14, 23	Mouser	270-499k-RC	
1	Resistor, axial lead	75k ohm, ⅛ W, 1%	R16	Mouser	270-75k-RC	
1	Resistor, axial lead	309k ohm, ⅛ W, 1%	R17	Mouser	270-309k-RC	
1	Resistor, axial lead	200k ohm, ⅛ W, 1%	R19	Mouser	270-200k-RC	
1	Resistor, axial lead	2.0k ohm, ⅛ W, 1%	R20	Mouser	270-2k-RC	
1	Resistor, axial lead	10k ohm, ⅛ W, 1%	R21	Mouser	270-10k-RC	
1	Resistor, axial lead	1.0M ohm, ⅛ W, 1%	R21	Mouser	270-1.0M-RC	
1	Resistor, axial lead	1.0k ohm, ⅛ W, 1%	R21	Mouser	270-1k-RC	
1	Switch	SPDT	S1	Mouser	1055-TA1130EVX	
1	Switch	SPST, momentary	S2	Mouser	103-1012	
1	IC, log amplifier	AD8307AR	U1	Digi-Key	AD8307ARZ-ND	
1	IC, voltage regulator	TL750L05CLP	U2	Digi-Key	296-8002-5-ND	
1	IC, instrumentation amplifier	AD623AN	U3	Digi-Key	AD623AN-ND	
1	IC, operational amplifier	TLV2462CP	U4	Mouser	595-TLV2664AIN	
1	IC, counter/divider/oscillator	CD4060BE	U5	Digi-Key	296-2060-5-ND	
1	P-channel mosfet, dual	IRF7314	Q1	Digi-Key	IRF7314-ND	
2	Machine screw, nut, lockwasher		4-40 x ⅜	TP1, 2		

minute intervals respectively.

The momentary-contact pushbutton switch, S2, starts the timing cycle. The battery drain is less than 10 mA for a few minutes, so an ON/OFF switch is not necessary. The DVM serves as a power-on indicator, and stops working before the low-dropout TL750L05CLP voltage regulator, U5, runs out of dropout range, so an internal battery test function is not provided. Test points TP1 and TP2 — 4-40 screws through the plastic enclosure — allow measurement of the battery voltage with an external voltmeter.

Table 1 indicates suggested components, but their selection is not critical. This timer may find other applications as a battery protector, and operates from 3 to 16 V dc, at up to 2.5 A dc with very low internal voltage drop.

Construction

The meter was built in a 4.7 × 2.6 × 1.7 inch plastic enclosure, as shown in the lead photo. An aluminum strap mounted to the bottom of the enclosure serves as a battery holder. The meter has a tendency to tip or to wander about the bench with cables attached, so an internal or external weight on the bottom may be desirable.

The bridge and detector circuit is assembled on a ¾ × 2 inch piece of Vector 8007 prototype board, with a 0.1 inch hole pattern, a ground plane on one side, and pads on the other side. The board is held in place on the back of BNC connectors J1 and J2 by epoxy on the connector mounting pins and by the soldered ground and center conductor leads, as shown in Figure 5. Both surface-mount and axial and radial-lead components were used, with essentially point-to-point wiring between components. Various versions of power meters and RC meters assembled by the author indicate relatively little difference in performance between assembly techniques, given reasonable attention to minimal-length leads. The AD8007AR SOIC version is fortuitously arranged so that the leads can be centered over a 3 × 3 pad pattern, with pins 7 and 6 over the center pad on one side, the ground pin 2 to the center pad on the other side, and the unused pin 3 bent upward. The AD8007AN dip package may be substituted with very little change in performance.

The bridge-detector module may be replaced with one incorporating wider bandwidth detectors from Analog Devices or Hittite, although printed circuit board construction would be needed. If one wanted to try different bridge-detector modules, it might be advantageous to locate the scale and offset trimmers and a buffer amplifier on the bridge-detector board.

The remaining circuitry is assembled on a 2 × 2 inch Vector 8007 board supported on the bottom of the enclosure. The layout of the meter driver and power circuitry is not

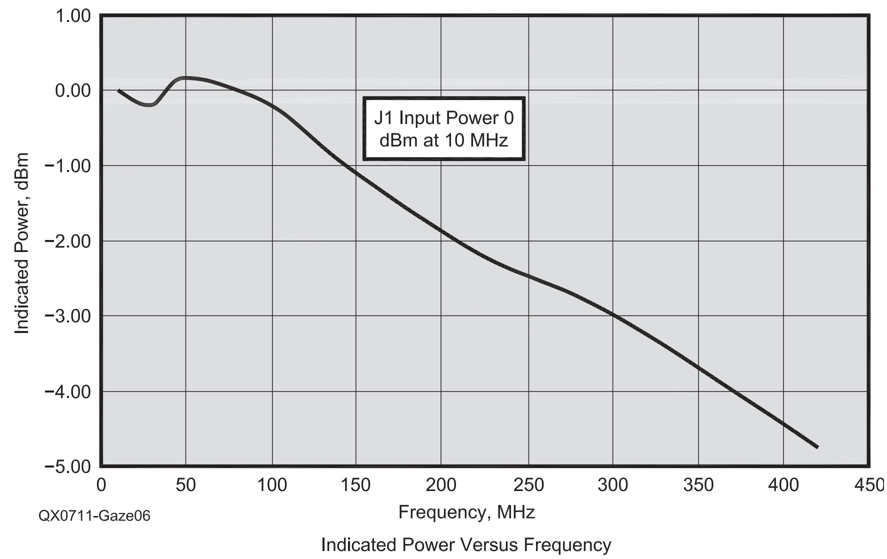

Figure 6 — Indicated power versus frequency.

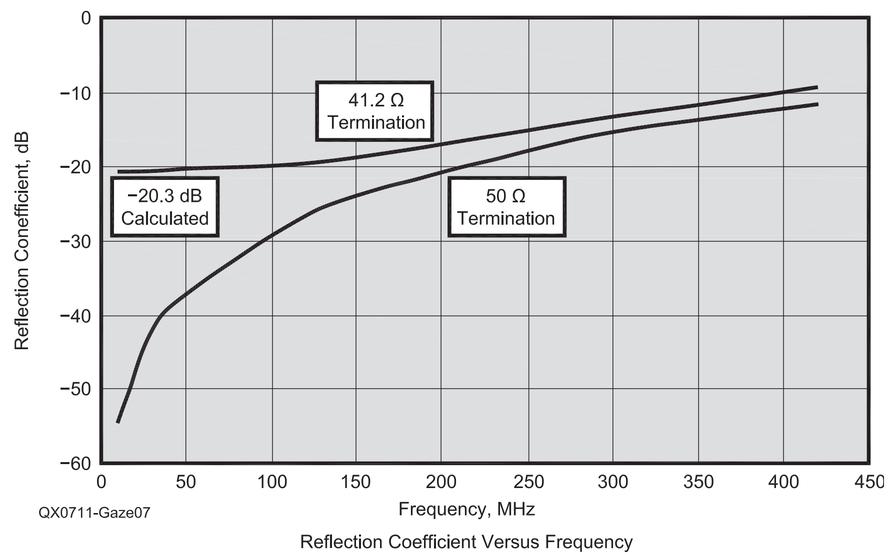

Figure 7 — Reflection coefficient versus frequency.

Table 2
Power and Reflection Coefficient Meter Performance

Frequency (MHz)	*Indicated Power (dBm)*	*50 Ω Reflection Coefficient (dB)*	*41.2 Ω Reflection Coefficient (dB)*
10	0.00	–54.5	–20.7
30	–0.20	–41.9	–20.7
50	0.17	–37.2	–20.4
100	–0.20	–29.3	–20
144	–1.01	–24.3	–19
222	–2.18	–19.4	–16.2
300	–2.98	–15.3	–13.4
420	–4.74	–11.6	–9.3

critical. The SOIC IRF7314 MOSFET, Q1, is not as nicely arranged as the AD8007AR, so the center pads of the 3 × 3 array were removed from the board, and connections were made to pins 2, 3, 6, and 7 of the SOIC package with wire-wrap wire. The interconnecting leads from the panel-mounted components were run to the associated pads on the circuit board, although had space permitted, it would have been neater to use a multiconductor cable and a connector on the circuit board.

Calibration

Calibration of the power meter mode may be accomplished with a signal generator or with a dedicated power level calibrator, such as described by Bob Kopski, K3NHI. (See Note 10.) Trimmer R18 is adjusted for 10.0 dB steps indicated on the DVM, corresponding to 10 dB steps of the signal generator or external attenuator. This should be within a few tenths of a dB from –60 to +10 dBm. The internal zero trimmer R15 is then adjusted to give a DVM indication corresponding to the signal generator or calibrator output.

In the RC mode, the meter is calibrated by feeding a signal into J2 at the desired frequency, at perhaps 0 dBm, and either terminating j1 in an open or short circuit. The coarse and fine zero adjustments are then used to set the DVM to an RC of 0 dB. Using either an open or short circuit at J1 works because either mismatch condition produces the same output voltage with a 50 Ω bridge source. The 3 dB pad improves the generator RC by 6 dB.

Operation

Indicated power should be linear within 0.3 dB up to 100 MHz and for input levels of about –60 to +10 dBm. The meter is inherently wideband, so it will respond to QRM, spurious input signals, and noise. Be sure that you are measuring the desired signal. The author's VHF / UHF log-periodic gives –17 dBm when pointed toward Boston.

The signal generator level is somewhat arbitrary when measuring RC, provided that the reading is above the QRM, spurious signals, and noise level. The wide range of the coarse zero adjustment permits reducing the level for measuring preamplifier input RC or increasing it for antenna measurements, for example. Zero adjustment should be repeated for each signal generator input power level change and for significant frequency changes.

Measurement of power or RC in 75 Ω environments may either be accomplished by use of a dedicated 75 Ω bridge as mentioned above, by incorporating a switch or relay to change the bridge impedance, or by use of an external minimum loss pad. If you build a 75 Ω bridge, remember that 50 and 75 Ω BNC connectors are compatible within this frequency range, but 50 and 75 Ω N connectors are catastrophically incompatible. Minimum-loss pads are commercially available (from Mini-Circuits or Elcom, for example), or may be built with a series 43.3 Ω resistor on the 75 Ω side, and a shunt 86.6 Ω resistor on the 50 Ω side. The loss of 5.7 dB shows up as a lower indicated power, but an 11.4 dB higher indicated RC (since the loss appears in both the forward and reverse signal direction).

This effect of loss on RC provides a useful way to measure loss. The meter is connected to the input of the passive device or transmission line under test, and zeroed with a short-circuit termination at the output port of the device or line. The termination is then moved to (or duplicated at) J1, and the new reading noted. The insertion loss is then just half of this loss in dB. A coaxial relay could connect a short-circuit termination at the antenna end of a feed line for an in-shack loss check.

The meter can be used to determine the resonant frequency of an antenna or circuit by observing the RC as the signal generator is tuned, even beyond frequencies where the RC is accurate. Antenna and filter matching may be optimized while observing the RC, either at discrete frequencies using the internal DVM, or by using a swept signal generator and an oscilloscope. Preliminary matching of a power amplifier output circuit may be performed by substituting a parallel RC network for the output tube(s) and measuring from the antenna port.

Performance

The performance of the Power and Reflection Coefficient Meter is summarized in Table 2, and in Figures 6 and 7. At lower frequencies, the power response is –1 dB at 490 kHz, –3 dB at 193 kHz, and –6 dB at 110 kHz. Increasing the coupling capacitors C3 and C4 would extend the response into the audio range, while smaller values might improve UHF response. The meter operates at battery voltage as low as 5.7 V dc, and draws less than 10 mA dc in operation, less than 16 µA dc in the power off mode.

Conclusion

A simple, easily constructed instrument has been described that combines power measurement and reflection coefficient measurement capabilities. Separation of the bridge/detector circuitry from the remaining circuits provides a means for accommodating future higher frequency detectors.

The power timer circuit may find additional applications as a means of protecting batteries in other bench instruments, digital voltmeters, calibrators, LCR metersand other circuits that can inadvertently be left turned on.

Appendix

Reflection Coefficient

Reflection Coefficient (RC), Return Loss (RL), and SWR all describe the degree of mismatch of a device or transmission line relative to a standard impedance. RL comes from telephone industry usage, and RC from the optical and electromagnetic branches of physics. The complex vectors **RL** and **RC** can be defined as the ratio of the voltages of the complex reflected and forward voltages, or as the square root of the ratio of the reverse and forward power.

$\mathbf{RL} = \mathbf{RC} = \rho = \mathbf{E}_r / \mathbf{E}_f$

$\mathrm{RL} = \mathrm{RC} = |\rho| = \sqrt{(P_r / P_f)}$

The SWR is similarly defined as:

$\mathrm{SWR} = (|\mathbf{E}_f| + |\mathbf{E}_r|) / (|\mathbf{E}_f| - |\mathbf{E}_r|)$

$\mathrm{SWR} = (1 + |\rho| / (1 - |\rho|)$

or conversely,

$|\rho| = (\mathrm{SWR} - 1) / (\mathrm{SWR} + 1)$

We can also define the complex vector **RL**, ρ as:

$\rho = (\mathbf{Z}_x - \mathrm{Z}_o) / (\mathbf{Z}_x + \mathrm{Z}_o)$

where the complex value ρ or $\mathbf{Z}_x$ can be represented in the form $\mathbf{A} + j\ \mathbf{B}$, or as a magnitude and angle. The complex ρ is also the scattering parameter S11 or S22, specified for many RF devices. The magnitude of ρ is:

$|\rho| = \sqrt{\mathbf{A}^2 + \mathbf{B}^2}$

RL is often defined as the logarithm of the magnitude of ρ:

RL (dB) = 20 log $|\rho|$

There is some confusion among various references as to the sign of RC (dB) or RL (dB). We assume here that $|\rho|$ is less than 1, so that the logarithmic scale goes from 0 dB at an open or short to – ∞ for a perfect match.

The ARRL Handbook, uses natural logarithms to define return loss.[11]

Return Loss (dB) = –8.68589 ln $|\rho|$,

Expressing this in common logarithms, we have:

Return Loss (dB) = –20 log $|\rho|$.

Similarly, the *Handbook* definition of mismatch loss translates to:

Attenuation (dB) = 10 log $(1 - |\rho|^2)$

A table in the Component Data and Reference chapter of the *Handbook* gives equivalent values of RC magnitude (in percent), attenuation, SWR and return loss. A more fine-grained table of SWR and return loss is available on the Mini-Circuits Web site (**www.minicircuits.com**).

Ralph Gaze, W1RHG, was first licensed as W1EOH in 1955. He received a BSEE from MIT in 1956, and an MS from George Washington. His work with McIntosh Laboratories, Melpar, Page Communications Engineers, Telcom, TASC and as a private consultant has gone from hi-fi to vocoder design, ionospheric scatter, tropospheric scatter, HF, VHF, UHF and microwave, telephone cable, meteor burst and satel-

lite Earth station, including those for the ECHO balloon, GOES and INTELSAT. These projects in telecommunications and related instrumentation have taken him to 15 countries, from Greenland to New Zealand and many more en-route.

Ralph is a Life Member of the IEEE, a former Member of the IEE (UK), a former Chartered Engineer in the UK and a Life Member of the Royal Scottish Country Dance Society. He and his wife Barbara have been enjoying retirement in Rhode Island and traveling for pleasure for the past five years.

Notes

[1]Larry Coyle, K1QW, "Instrumentation Amplifiers and LCDs as Measurement Tools," *QST*, May 2006, pp 55-58. Also see Feedback, *QST*, June 2006, p 31.

[2]Zack Lau, KH6CP/1 (now W1VT), "LF/MF/HF/VHF Return-Loss Bridge," Hints & Kinks, *QST*, Apr 1996, p 76.

[3]*Low Cost DC-500 MHz, 92 dB Logarithmic Amplifier*, Analog Devices, Inc, 1997.

[4]M. Pilotte, "Operation of RF Detector Products at Low Frequency," *Application Note AN-691*, Analog Devices, Inc.

[5]Wes Hayward, W7ZOI, and Bob Larkin, W7PUA, "Simple RF Power Measurement," *QST*, Jun 2001, pp 38-43. Also see Feedback, *QST*, Aug 2001, p 76.

[6]Bob Kopski, K3NHI, "An Advanced VHF Wattmeter," *QEX*, May/Jun 2002, pp 3-8.

[7]Bob Kopski, K3NHI, "A Simple Enhancement for the 'Advanced VHF Wattmeter,'" *QEX*, Sep/Oct 2003, pp 50-52.

[8]Thomas Scherrer, OZ2CPU, "Digital RF Wattmeter with LC Display," *Elektor Electronics*, Oct 2002. This article is available in Danish and English on Thomas's Web site: **www.webx.dk**. Elektor Electronics also offers the article, a list of errata, PC board layouts and free software on their Web site: **www.elektor-electronics.co.uk**

[9]Roger Hayward, KA7EXM, "A PIC-Based HF/VHF Power Meter," *QEX*, May/June 2005, pp 3-10.

[10]Bob Kopski, K3NHI, "A Simple RF Power Calibrator," *QEX*, Jan/Feb 2004, pp 51-54.

[11]Mark Wilson, K1RO, Ed, *The ARRL Handbook for Radio Communications*, 2008 Edition, p 7.48 and 21.4-21.8. *The ARRL Handbook* is available from your local ARRL dealer, or from the ARRL Bookstore, ARRL order no. 1018. Telephone toll-free in the US 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

SAN2PC: A Spectrum Analyzer to PC Interface

How to connect your old analyzer to a PC with a simple, low cost, interface and plot or print spectra.

Roland Cordesses, F2DC

Introduction

The Spectrum Analyzer (SA) is an incredibly useful tool to the homebrewer. Unfortunately, it is very expensive when purchased new, so the SAs found in the workshops of many hams are of the secondhand type, or even home built. As a result, they often lack functionalities available on more modern equipment. For instance:

- It is not possible to store the displayed spectrum in a file for later computer processing.
- Getting a hard copy of the displayed spectrum is not an easy job: dedicated cameras are scarce and expensive, and X-Y plotters are cumbersome.

In order to solve these two points, I have recently designed and built a simple interface to connect my spectrum analyzer to a PC. This unit, named SAN2PC, digitizes analog spectrum data, then processes and transmits that data to the computer via the RS232 line. SAN2PC automatically takes care of the spectrum analyzer scan time. During initial calibration, relevant information is stored in the PIC EEPROM. The heart of SAN2PC is a PIC 18F2525 processor, running embedded software written in C language.[1]

Any terminal program can be used to receive and store RS232 data in a file. This file can be processed afterward, or simply read by any spreadsheet program: The spectrum can then be displayed on the computer screen and a copy printed if needed.

SAN2PC has been tested on my HP141T analyzer, but I think it could be used with other models and brands, insofar as they have analog spectrum voltage and pen lift signal output connectors. Limited tests have been conducted on a Tektronix 2755P.

I have also written dedicated software (in the *Python* language) displaying not only the spectrum, but also the spectrum analyzer's settings (center frequency, scan width, bandwidth, and so on) on the PC screen. This picture can be stored in a file for future use. The program runs under *Windows XP* and has been tested under *Linux* (the *Python* software I developed is not limited to the *Windows* world).

Presentation

Most SAs of the kind we are talking about have the following auxiliary connectors, usually intended for connection to an X-Y plotter:

- Vertical Output (VO) is a voltage corresponding to the spectral power for each frequency step, and is similar to the vertical cathode ray tube (CRT) deflection voltage. VO can be positive or negative, according to the SA model, and generally has an amplitude on the order of a volt.

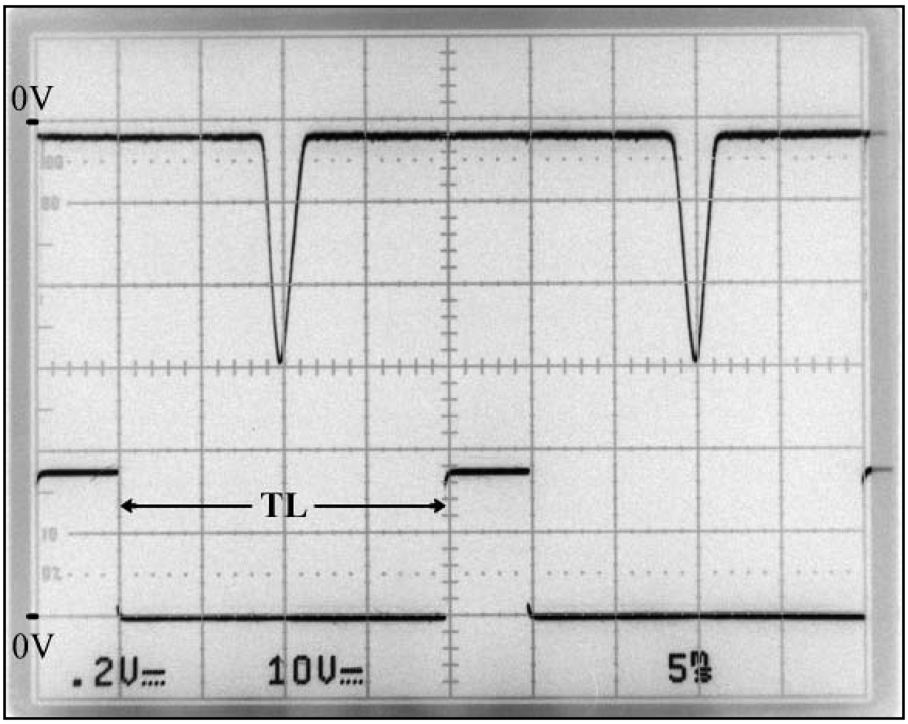

Figure 1 — VO and PL signals, as displayed on an oscilloscope screen.

[1]Notes appear on page 29.

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• Pen Lift output (PL) is a logic level (0-20 V or TTL) to lift the X-Y plotter pen up or down. Generally, a PL low level lifts the pen down, thus meaning that spectrum voltage VO is valid.

• Scan Output (SO) is a sawtooth waveform signal similar to the one used for horizontal CRT deflection. Its polarity and amplitude are different from one model to another.

At the very beginning of the project, I thought of using the three outputs, but it quickly turned out that only the first two were necessary: the idea is to digitize VO (around 1000 samples per spectrum, value not critical) as soon as PL is going down. Figure 1 shows the two signals generated by my HP141T, VO being the upper trace and PL the lower.

The PIC first measures the time, TL, during which PL is low, computes the sampling period, TS, waits for the next falling edge of PL, and then starts 1030 analog to digital (A/D) conversions. Thanks to this approach, the sampling period, TS, is automatically modified each time the spectrum analyzer scan time is changed. The drawback is that it takes twice the TL time to complete a full spectrum measurement.

Have a look at the SAN2PC block diagram presented on Figure 2:

• The PIC 18F2525 is pin-compatible with the 16F873, but features, among other improvements, a larger user RAM (nearly 4 kbytes) allowing storage of the full spectrum data without any problem. Its internal A/D converter offers a 10-bit resolution and a conversion time short enough for this application. Moreover, A/D conversions can be triggered by an internal 16 bit counter/timer. The input A/D voltage range is 0 to 5 V full scale.

• The purpose of the level converter boxes is to transform PL and VO to voltage levels PL1 and VO1, accepted by the PIC:

• If PL is a TTL signal, the level converter is omitted, and if it is a 0-18 V signal, a simple resistive bridge will do the job (that is the HP141T case). Just remember that this bridge must be compatible with the load accepted by the SA Pen Lift output: Check your equipment operation manual.

• When it comes to VO, there are several choices for polarity and amplitude.

The HP141T VO swings from 0 to –800 mV for the full 80 dB RF input range of the SA: An operational amplifier in inverting mode provides the 18F2525 A/D converter with the required positive voltage.

On some other SAs, VO can go from 0 to +800 mV or more, depending on models and brands: A non-inverting amplifier must then be used.

The amplifier gain is adjustable to allow fine calibration of the interface.

• A 20 MHz crystal is connected between the oscillator pins of the PIC. A MAX232 driver/receiver provides the level shifting for the RS232 link to the PC.

• SAN2PC is powered by any standard wall power adapter (8 V dc or more) thanks to on-board regulators and dc/dc converter. The required current is less than 100 mA.

Circuit Description

The schematic diagram presented in Figure 3 shows that the circuitry is very simple.

• The operational amplifier is a classic dual one (TL82, TS272 or similar op amp). One can note that the second half is not used: It is just a spare part for a future project!

The Printed Circuit Board (PCB) is designed in such a way that U4A may be wired as an inverting or a noninverting amplifier: delete R12 and connect VO to J1 in the first case, or ground J1 and properly choose R12 and R10, VO being connected to J14 in the second case. If the VO voltage from your spectrum analyzer is different from that of the HP141T, don't forget to modify the resistors around U4A.

• R3 and R11 are the voltage divider resistors going from PL to the RB0/INT pin of the PIC. You may adapt their values in accordance with the PL voltage of your analyzer.

• Analog voltage on pin AN0 (A/D converter input) must be kept inside a –0.3 V to +5.3 V range: Schottky diode D7 deals with the negative side and we supply U4 with ± 6.5 V to limit its positive output voltage around 5 V. In case of a positive only VO, the dc/dc converter, U2-U3 regulators and D7 could be omitted, and a single supply (5 V) op amp of the TS272 variety could be used instead of a TL82.

• Two RS232 speeds are available: 9600 and 115200 Bauds. This speed is selected by a rear panel switch.

• RC2 pin is wired to a front panel Normal / Calibrate switch.

• Two front panel LEDs keep the user informed of the SAN2PC status: waiting during TL measurement or sampling spectrum during data acquisition.

• The Start button is connected to the PIC MCLR reset pin.

The PCB file can be downloaded from my Web site at **roland.cordesses.free.fr**. Download the file **san2pcbd.pdf**.[2]

Figure 4 shows the inside of the unit and Figure 5 is a view of the interface standing on the top of the HP141T.

How It Works

Normal Mode

When the PL1 falling edge is detected on the INT pin of the 18F2525 (see Figure 1), at the beginning of the frequency scan, the software enters a procedure to measure the duration of the low level until the following PL1 rising edge appears. This time TL, related to the scan time setting of the SA, is stored as a 32 bit integer. The sampling period, TS, is then computed:

TS = TL / **sizemax** (1)

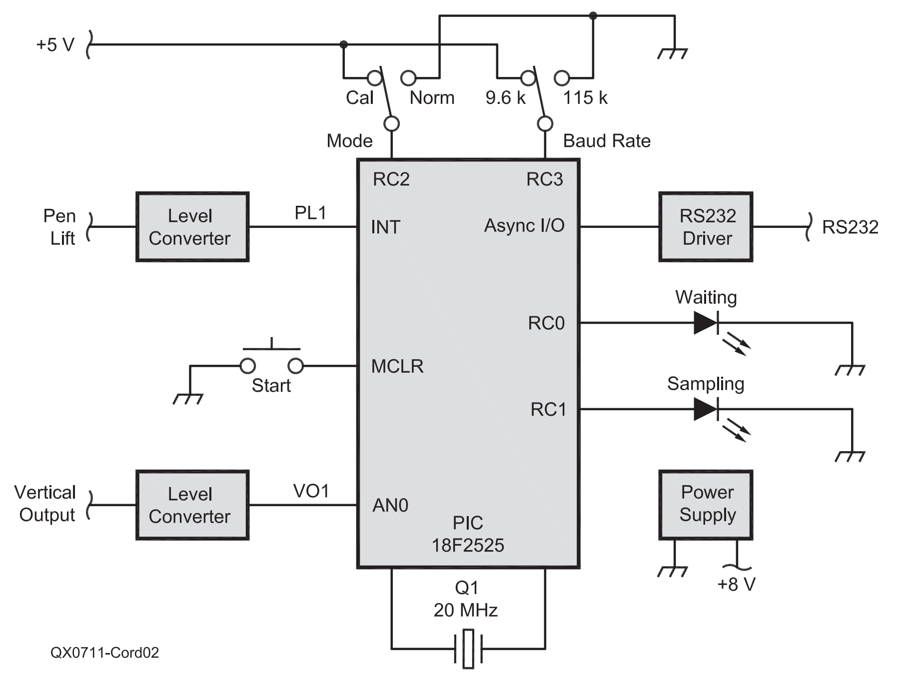

Figure 2 — This block diagram of the SAN2PC interface shows the basic circuit operation.

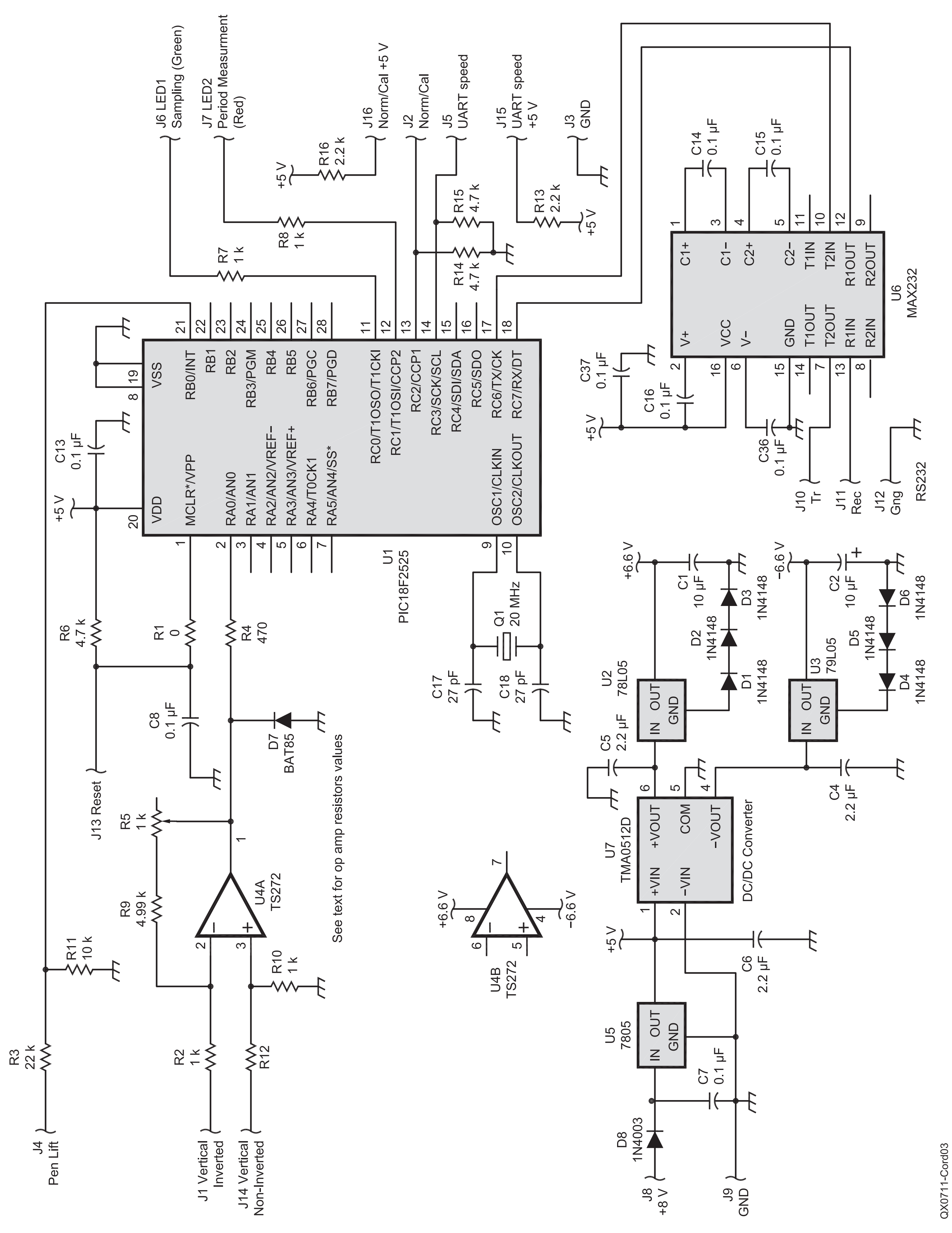

Figure 3 — The SAN2PC schematic diagram.

where **sizemax** = 1030 is the number of samples per spectrum (as said earlier, this value is not critical).

The program then waits for the next PL1 high to low transition, and enables the interrupts. It then enters a loop during which **sizemax** values of VO1 are sampled every TS by the A/D converter and stored in the PIC RAM as an array called **specarray**. The interrupts are then disabled, **specarray** samples are read from indexes **ileft** to **iright** (see calibration mode), processed, converted to dBm and sent to the PC via the UART.

We use the Compare and Reset function offered by the **Capture**/**Compare**/**PWM** (CCP) module of the 18F2525 to start A/D conversions. In **Compare** mode, the 16 bit CCP2 register is constantly compared against the **Timer1** register value. This latter register is incremented on every internal instruction cycle, TI (Fclock/4), where Fclock is the PIC crystal frequency. When a match occurs, and if the A/D converter has been enabled, a conversion is triggered and the result is stored in **specarray**. **Timer1** is then reset and the process goes on until the last sample has been stored in **specarray**.

CCP2 is loaded with a 16 bit integer, **NCCP2,** corresponding to the TS period, with **NCCP2** being computed from TL and TI. As TL ranges from a few tens of miliseconds up to 100 seconds, according to the analyzer scan time setting, **NCCP2** is automatically updated each time the scan time is changed. Moreover, TI is possibly prescaled to prevent CCP2 overflow, when long scan times are selected.

Before transmitting **specarray** data to the PC, the program sends a header describing the analyzer main settings. Because these settings can't be automatically transmitted by the SA to SAN2PC, the program initiates a dialog with the user before spectrum acquisition. This dialog, controlled by the PIC embedded software, begins just after the push of the start button. The user is asked to enter the Reference Level (dBm), Center Frequency (with units), Scan Width/division (with units), Filter Bandwidth and Video Filter Bandwidth (with units). This information, packed as an ASCII string (the header), will be transmitted as the file data first line, just before spectrum data are sent by SAN2PC to the PC.

The file format, header included, is shown below:

980M 5M 300k 10k *
955.0000 –93.3
955.0506 –93.7
________ ____
________ ____
1005.0000 –93.2
*

The asterisk following the video bandwidth (10k) performs as the "start of data"

F2DC

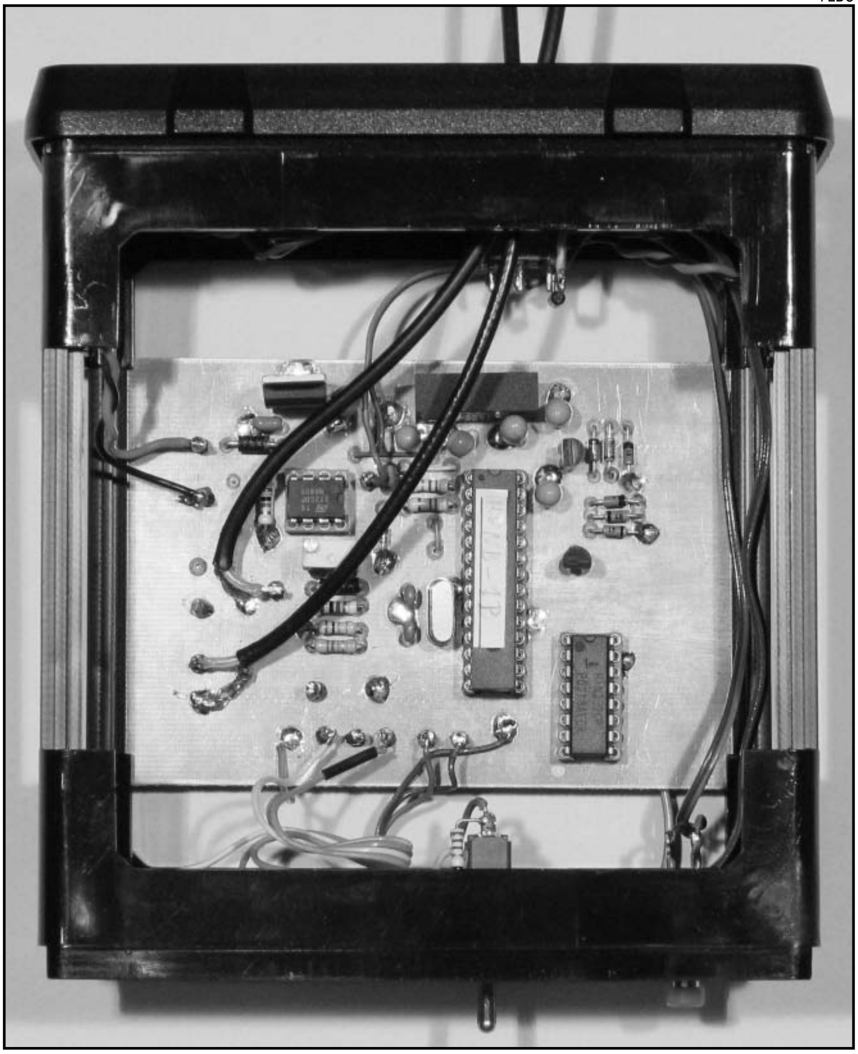

Figure 4 — This photo shows a look inside the SAN2PC interface.

F2DC

Figure 5 — The SAN2PC is a small package that easily sits on top of the HP141T spectrum analyzer.

delimiter. It is followed by the first frequency, then the first RF power level and so on, until the second asterisk (end of data) is found.

The header information will be used by the *Python* plotting program, to display the SA settings in the upper part of the spectrum picture, during the plotting phase.

Calibrate Mode

Upon first power-up, SAN2PC will read the "default" calibration information from the PIC EEPROM: You will probably want to calibrate it more accurately, as these values may not be well suited to your SA. There are two calibration procedures, the first one dealing with amplitude and the second with frequency. The PIC embedded software takes care of these two tasks when Calibrate is selected by the Mode switch.

• Amplitude: U4A gain must be trimmed until the number of A/D points corresponding to the RF reference level, (the upper graticule line of the SA screen) is 800. This step may be performed with the SA calibrator used as an RF source, the scan width control on zero (no scan) and the input attenuator adjusted to put the trace just on the upper graticule. With 800 points related to the 80 dB SA input range, the amplitude resolution is 0.1 dB.

• Frequency: We said earlier that **sizemax** samples are written in **specarray** during *Normal* mode operation. Figure 6 (not to scale) shows that the PL falling edge appears earlier than the time the analyzer CRT spot crosses over the left graticule: it is thus necessary to know this delay, **t1,** in order to get a good frequency accuracy. Index **ileft** is a measure of this delay, and **iright** is representative of the delay between the time the spot crosses over the last right graticule and the rising edge of PL. During transmission to the PC, the only **specarray** values included in **ileft** to **iright** are sent.

The frequency calibration is very simple: you connect an RF source to the SA input (the calibrator for instance) and are asked to put the displayed "peak" (using the SA frequency control) on the first left graticule then to press Enter. Then you put the "peak" on the last right analyzer graticule and press Enter again. The software calculates the RAM indexes **ileft** and **iright** of the two maxima and stores them in EEPROM: **ileft** and **iright** will be the values used by the **specarray** reading procedure during Normal mode operation, when sending data to the PC.

Software

The PIC Embedded Software

The PIC program, named **san2pc.c**, is written in C language for easier development and debugging. For some years I have developed projects using the Custom Computer Services (CCS) C-Compiler.[3] This compiler is easy to use and a lot of useful tools as well as sample programs are available as part of the package.[4] I would also like to thank the CCS French representative, Hi Tech Tools, for their kindness.[5]

Programming the PIC was done with a home made programmer inspired by the ProPic2 and using the WinPic800 software.[6, 7] In addition, the CCS bootloader was of great help during program development, speeding up debugging.

The HEX file (**san2pc.hex**) can be downloaded from the Spectrum Analyzer Interface directory of my Web site or from the *QEX* Web site as part of the **11x07_Cordesses.zip** file. See Note 2.

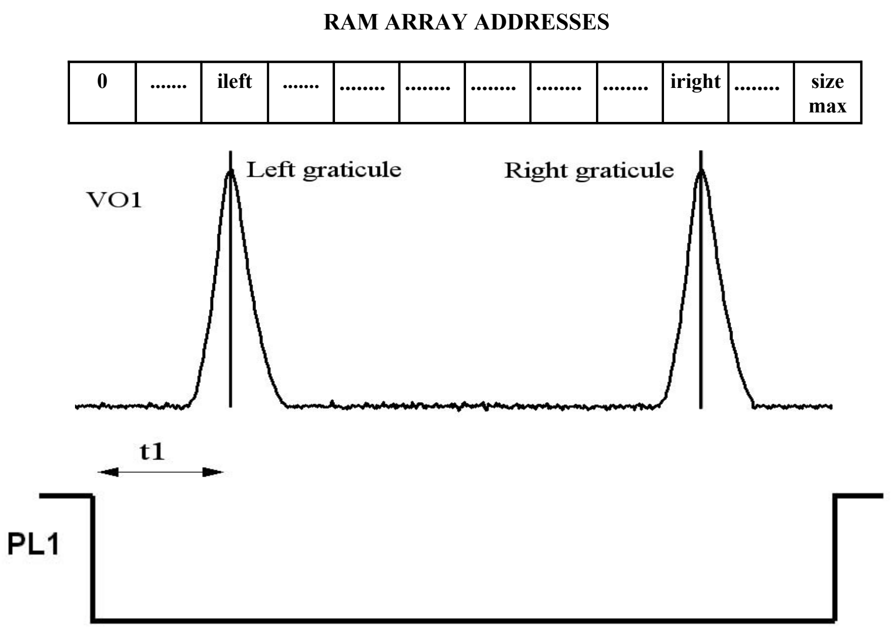

Figure 6 — This diagram illustrates the concept of the SAN2PC frequency calibration technique.

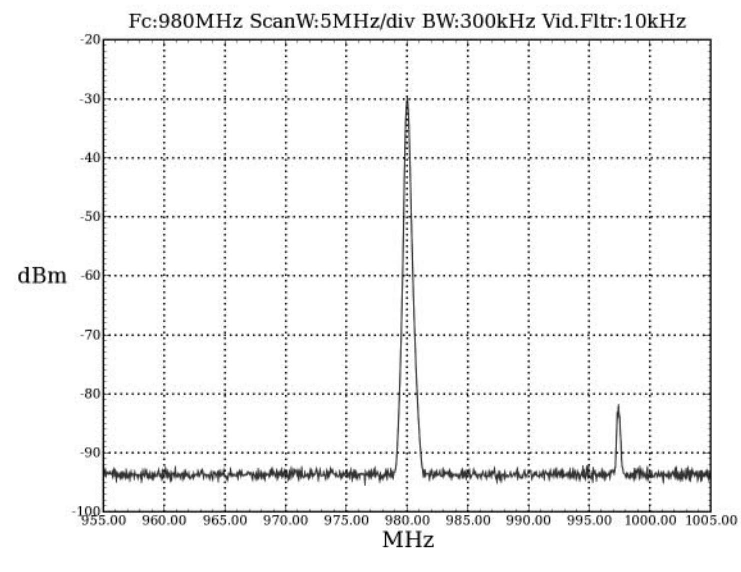

Figure 7 — This spectrum sample display is the result of the author's *Python* plotting program, drawsan.py.

The PC Plotting Software

I have also developed a dedicated program running on my PC under *Windows XP* (and successfully tested under *Gentoo Linux*) in order to plot, on the computer screen, not only the spectrum, but also the analyzer settings (the same way they are displayed on a modern SA screen). This program, named **drawsan.py**, relies on the header of the stored file.[8] I chose to write it in *Python* language, mainly because of the availability of a first class graphics library, Matplotlib.[9, 10] This library emulates Matlab™ commands (but does not require Matlab) and creates very nice plots. The **drawsan** user has the opportunity to save the picture as a png, bmp or vector file. A sample picture of a spectrum plotted with **drawsan.py** is shown in Figure 7.

Please, be lenient with **drawsan.py**: It is my first *Python* program and must be seen as a beginner's exercise!

Operating SAN2PC

It is very easy to use this interface: connect PL and VO lines to the corresponding SA connectors, link SAN2PC to your computer through the RS232 line and run a terminal program. Then push the Start button and answer the questions described earlier.

Please note that SAN2PC amplitude and frequency resolutions are better than the accuracy of most spectrum analyzers of this kind. Keep that in mind when interpreting data files or pictures!

Conclusion

This simple interface may help to solve the hard-copy problem associated with many secondhand spectrum analyzers. It also provides the user with a file for subsequent computer processing. It could probably also be used with other equipment, such as network analyzers, that don't have any printing function.

Acknowledgments

I would like to thank my two sons, Lionel and Joël, who strongly suggested that I use *Python* in order to get nice looking spectral displays. Although the learning curve was sometimes a rough one, I don't regret this contact with a new programming language.

Roland Cordesses, F2DC, was first licensed in 1962, and he has been a member of ARRL since 1964. He graduated as an Electronics Engineer and worked as a Research Engineer in an Observatory devoted to atmospheric and earth sciences.

For nearly 30 years, he worked in the design and development of radar systems for remote sensing of the atmosphere and for monitoring of volcanic eruptions. During his career, he has presented or published many papers related to electronics and geophysics.

Recently retired, he continues to be an avid homebrewer: over the years he has designed and built many receivers, transceivers and measuring-equipment projects.

In addition to Amateur Radio, he enjoys building and flying model aircraft (homemade receivers and transmitters, of course) and hiking in the mountains.

Notes

[1] See **www.microchip.com**.

[2] The circuit board pattern is also available for download from the *QEX* Web site. Go to **www.arrl.org/qexfiles** and look for the file **11x07_Cordesses.zip**.

[3] PIC MCU C Compiler Reference Manual, CCS, Inc, October 2005.

[4] See **www.ccsinfo.com**.

[5] See **www.hitechtools.com**.

[6] See **www.propic2.com**.

[7] See **www.winpic800.com**.

[8] The program files are available for download from the author's Web site at **roland.cordesses.free.fr** as well as from the *QEX* Web site. (See Note 2.)

[9] See the official *Python* Web site: **www.python.org**.

Program Your Own Voice Keyer/Recorder

Use the free Turbo Delphi Explorer *software to program this useful station accessory.*

Steve Gradijan, WB5KIA

Make your own customizable voice keyer using free software from CodeGear (Borland). The free *Windows* compiler called *Turbo Delphi Explorer* was described in the Sep/Oct 2007 issue of *QEX*.[1] With *Explorer* and a simple-to-build audio interface, you can use your PC to record audio off the air or key your transmitter with short audio messages in your own voice.

Explorer software lets you build a useful, customizable tool to very easily record SSB and digital signals off the air. Recorded information provides data for demonstrations and equipment testing, and lets you archive your DXing accomplishments. You may also discover that modern *Windows* object oriented compilers are not that difficult to master. Programming can be fun and is an especially useful skill. If you are not interested in programming, a compiled version of the software *Voice Keyer/Recorder* as shown in Figure 1 is provided for download.[2]

Overview

It is very simple to program a basic audio recorder and keyer with *Explorer*. The compiler's design screen is shown in Figure 2. If you are satisfied with VOX control of your transmissions, the code is easy to implement. Additional coding effort is required to use the PTT circuit in your radio, and requires an additional interface connected to a PC serial port or USB/Serial adaptor.

Source code is provided for the recorder / voice keyer. The code described is for *Turbo Delphi Explorer*, although the techniques apply to other *Windows* compatible compilers such as *Visual Basic*, *C*, and *Visual C++*.

[1]Notes appear on page 36.

The *Windows* API "PlaySound" function simplifies "building" the basic keyer. The PlaySound function can play *.wav files synchronously or asynchronously. This project uses both types of program call. The behavior of the sound playback is different for each, without modifying the played wave file. Synchronous playing of a wave file starts

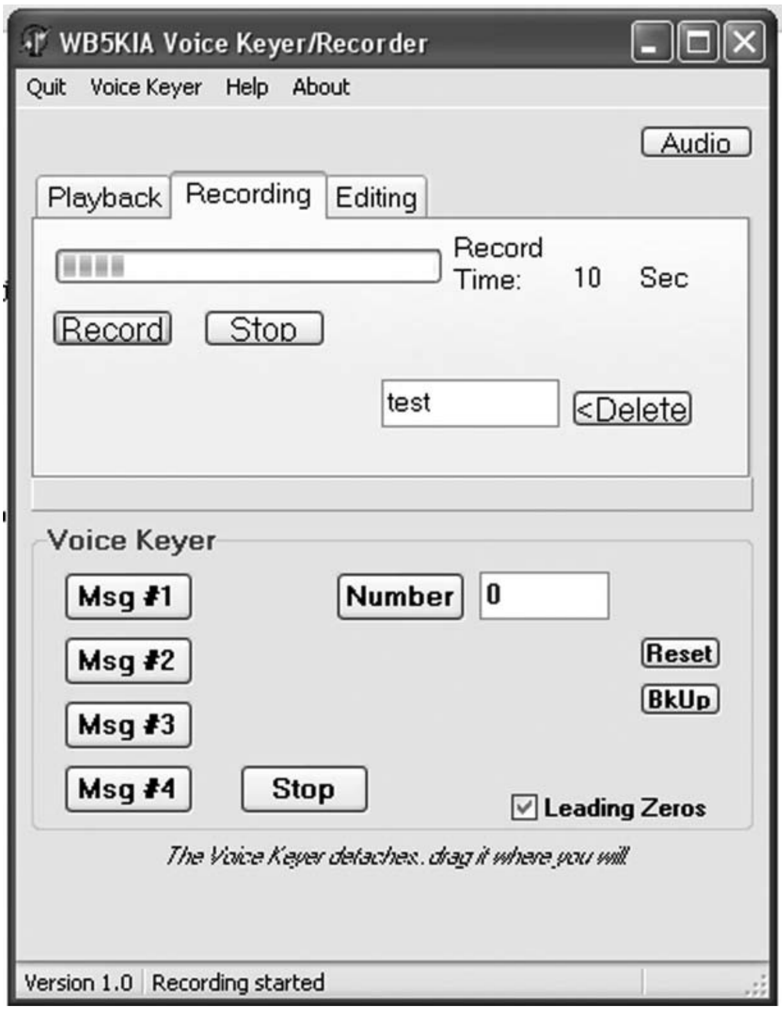

Figure 1 — *Voice Keyer/Recorder* allows recording of on-the-air audio, can playback wave file recordings, and functions as a voice keyer for calling CQ, contest exchanges and other repetitive tasks.

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and then continues without interruption until finished. It is possible to terminate the playing of an asynchronous wave file at any point. It is not possible to interrupt synchronous playback.

If you already have *Pascal* programming skills, a complex program with automatic audio levels and other tools using the *Windows* wave mixer can be coded. This will require additional code or a free, third-party control. Unfortunately, third-party controls are not usable with the free version of *Explorer*, so you have to provide a significant amount of manual coding or buy one of CodeGear's advanced compilers to use them.

The Example Software

Voice Keyer/Recorder has a wave file recorder, a wave playback, four message buttons and a "counting" message button. It requires a simple audio interface to connect between your transceiver and PC sound card, as shown in Figures 3 and 4. The interface protects both your radio and PC, and serves to eliminate ground loops and hum. Use the compiled version of the software to get started immediately or load the source code into *Explorer* and modify the program to design your own voice tool box.

Use the Software to Modify the Program

Download *Explorer* from CodeGear (**http://turboexplorer.com/downloads**) along with the prerequisite files. (See the *QEX* article about *Turbo Delphi Explorer*, listed at Note 1.) You want the *Explorer* version for Win32. Register the software with CodeGear to obtain an access code (this requires filling out a simple survey). The *Voice Keyer/Recorder* example code is provided on the *QEX* Web page (see Note 2); download and unzip the files. The easiest way to load *Explorer* is to double-click on the example code file, which is called AudioKeyer.bdsproj. Wait a moment and *Explorer* will automatically load the source code. The source code will appear in the *Explorer* design screen (Figure 2). Click the recorder tab, then click the Run toolbar menu (at the top of the compiler window). The source code compiles and runs the code, creating a stand-alone AudioKeyer.exe file.

The Source Code

There are about 600 lines of code in the *Voice Keyer/Recorder* project; most of these lines are generated automatically by the compiler. The source code is commented throughout. Several of the algorithms are critical, and worth reviewing. Table 1 (on the next three pages) shows five critical procedures.

The "TWaveRec.Button1Click(Sender: TObject);" procedure engages the procedure that does the sound recording. Wave files are produced in 8 bit mono, 8 kHz at 8,000 bytes/s. This format produces a wave file that is about 850 Kbytes in size for a 60 second recording. Recording length is

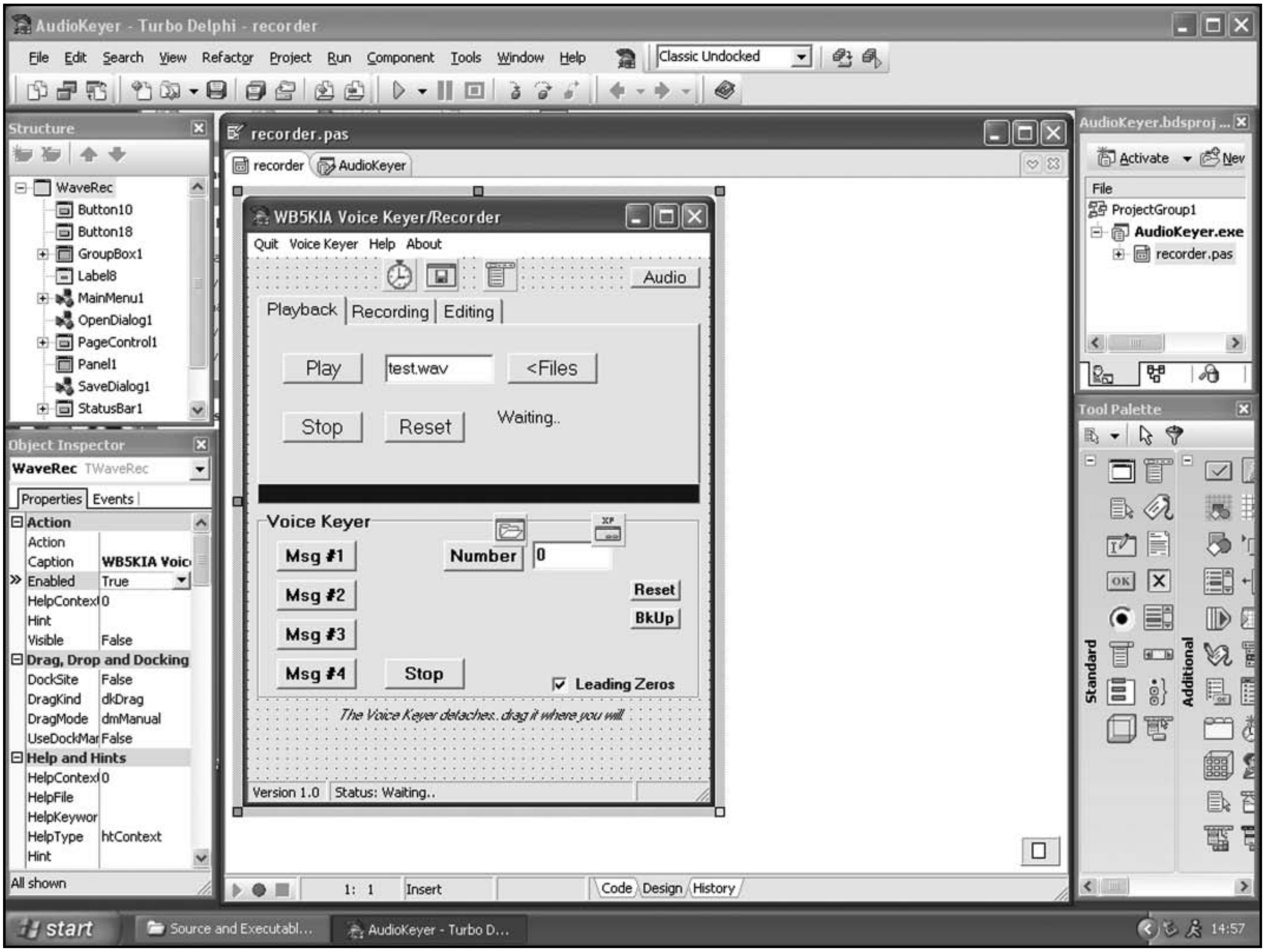

Figure 2 — This *Turbo Delphi Explorer* design screen shows the voice keyer form under construction.

Table 1
***Voice Keyer/Recorder* Source Code Snippets**

The Recording Routine

```
procedure TWaveRec.Button1Click(Sender: TObject);
begin
   button3.visible:=false; //show Save buttons
   button4.visible:=false;
   Label2.Caption:='0';  //Sets label2 to 0
   Timer1.Enabled:=true; //Turns timer1 on for graphically showing recording length
   StatusBar1.Panels[1].text:='Recording started';
   Progressbar1.position:=0;  //reset

   mciSendString('OPEN NEW TYPE WAVEAUDIO ALIAS MicSound', nil, 0, Handle);

   mciSendString('SET MicSound TIME FORMAT MS ' +      // set the time format

        'BITSPERSAMPLE 8 ' +                            // 8 Bit
        'CHANNELS 1 ' +                                 // MONO
        'SAMPLESPERSEC 8000 ' +                         // 8 KHz
        'BYTESPERSEC 8000',                             // 8000 Bytes/s
        nil, 0, Handle);

  mciSendString('RECORD MicSound', nil, 0, Handle);
end;
```

Stop Recording

```
procedure TWaveRec.Button2Click(Sender: TObject);
begin
  Timer1.Enabled:=false;   //Stops the timer
  mciSendString('STOP MicSound', nil, 0, Handle);
  button3.visible:=true; //show Save
  button4.visible:=true; //Show Save AS
  StatusBar1.Panels[1].text:='Recording concluded';
end;
```

Save Wave File of what Was Just Recorded

```
procedure TWaveRec.Button3Click(Sender: TObject);
begin
  mciSendString(PChar('SAVE MicSound ' +'recWaves\' +trim(Edit1.text)+ '.wav'), nil, 0, Handle);
  mciSendString('CLOSE MicSound', nil, 0, Handle);
  StatusBar1.Panels[1].text:= 'Wave file saved to : '+edit1.text+ '.wav'; //Lets you know it is saved
  button3.visible:=false; //hide the Save buttons
  button4.visible:=false;
end;
```

Save a Wave File

```
procedure TWaveRec.Button4Click(Sender: TObject);
begin    // Save As button
 if SaveDialog1.Execute then
    mciSendString(PChar('SAVE MicSound ' + SaveDialog1.filename), nil, 0, Handle);
 mciSendString('CLOSE MicSound', nil, 0, Handle);
 StatusBar1.Panels[1].text:= 'Wave file saved to ' + SaveDialog1.filename; //Lets you know it is saved
 button3.visible:=false; //hide Save
 button4.visible:=false;
end;
```

Play a Wave File

```
procedure TWaveRec.Button5Click(Sender: TObject);
var
wfile:string;
begin     // Play button
  Label1.Caption:='Playing';
  StatusBar1.Panels[1].text:='Playing';
  wfile:='recWaves\'+trim(edit2.text);
```

```
  if copy(wfile,length(wfile)-4,1) = '.' then
    begin
      wfile:=wfile+'.wav';
      PlaySound((PChar(wfile)), 0, SND_aSYNC); //play the sound asynchronously
      StatusBar1.Panels[1].text:='Finished Playing.';
    end
 else
   begin
     PlaySound((PChar(wfile)), 0, SND_aSYNC);
     StatusBar1.Panels[1].text:='Finished Playing.';
   end;
end;
```

Stop Playing a File Played Asynchronously

```
procedure TWaveRec.Button6Click(Sender: TObject);
begin
  PlaySound(nil, 0, 0); // stop the current sound
  //  PlaySound(nil, 0, NULL); // stop the current sound needs 0 instead of Null
  Label1.Caption:='Finished';
  StatusBar1.Panels[1].text:='Finished';
end;
```

Play the Voice Keyer Messages

```
procedure TWaveRec.Button11Click(Sender: TObject);
var    //play the msg1.wave file you recorded
wfile:string;
begin
  wfile:='recWaves\' +'msg1.wav'; //trim(edit2.text);
  if copy(wfile,length(wfile)-4,1) = '.' then
    begin
      wfile:=wfile+'.wav';
      PlaySound((PChar(wfile)), 0, SND_aSYNC);
      StatusBar1.Panels[1].text:='Finished Playing.';
    end
  else
    begin
      PlaySound((PChar(wfile)), 0, SND_aSYNC);
      StatusBar1.Panels[1].text:='Finished Playing.';
    end;
end;
```

Play the Contest Numbers

```
procedure TWaveRec.Button14Click(Sender: TObject);
var   //play the  ordinal number wave files you recorded (these are in the
      //defaultWaves folder
i:integer;
wfile,wfile2,wfile3,wfile4:string;
begin

  edit3.text:= inttostr(strtoint(edit3.text)+1);  //increment the counter
  if checkbox1.checked=true then
    begin   //add leading zeros if necessary
      if length(edit3.text)<3 then edit3.text:='0'+edit3.text;
      if length(edit3.text)<3 then edit3.text:='0'+edit3.text;
    end;
  i := length(trim(edit3.text)) ;
  case i of //if there are no leading zeros, handle it this way for single digits
    1:  begin
          case  strtoint(copy(trim(edit3.text),1,1)) of
            0..9:wfile:= 'defaultWaves\' +copy(trim(edit3.text),1,1)+'.wav';
          end; //case
          PlaySound((PChar(wfile)), 0, SND_SYNC);
          StatusBar1.Panels[1].text:='Finished Playing.';
        end;
```

Table 1
***Voice Keyer/Recorder* Source Code Snippets (Continued)**

```
    2:  begin //if there are no leading zeros, handle it this way for double digits
          case  strtoint(copy(trim(edit3.text),1,1)) of
            0..9:wfile:='defaultWaves\' + copy(trim(edit3.text),1,1)+'.wav';
          end; //case

          PlaySound((PChar(wfile)), 0, SND_SYNC);

          case  strtoint(copy(trim(edit3.text),2,1)) of
            0..9:wfile2:= 'defaultWaves\' +copy(trim(edit3.text),2,1)+'.wav';
          end; //case
          PlaySound((PChar(wfile2)), 0, SND_SYNC);
          StatusBar1.Panels[1].text:='Finished Playing.';
        end;

    3:  begin //if there are no leading zeros, handle it this way for triple digits
          case  strtoint(copy(trim(edit3.text),1,1)) of
            0..9:wfile:= 'defaultWaves\' +copy(trim(edit3.text),1,1)+'.wav';
            end; //case
          PlaySound((PChar(wfile)), 0, SND_SYNC);

          case  strtoint(copy(trim(edit3.text),2,1)) of
            0..9:wfile2:= 'defaultWaves\' +copy(trim(edit3.text),2,1)+'.wav';
            end; //case
          PlaySound((PChar(wfile2)), 0, SND_SYNC);

          case  strtoint(copy(trim(edit3.text),3,1)) of
            0..9:wfile2:= 'defaultWaves\' +copy(trim(edit3.text),3,1)+'.wav';
            end; //case
          PlaySound((PChar(wfile2)), 0, SND_SYNC);
          StatusBar1.Panels[1].text:='Finished Playing.';
         end;

     4:  begin
          case  strtoint(copy(trim(edit3.text),1,1)) of
            0..9:wfile:= 'defaultWaves\' +copy(trim(edit3.text),1,1)+'.wav';
          end; //case
          PlaySound((PChar(wfile)), 0, SND_SYNC);

          case  strtoint(copy(trim(edit3.text),2,1)) of
            0..9:wfile2:= 'defaultWaves\' +copy(trim(edit3.text),2,1)+'.wav';
          end; //case
          PlaySound((PChar(wfile2)), 0, SND_SYNC);

          case  strtoint(copy(trim(edit3.text),3,1)) of
            0..9:wfile2:= 'defaultWaves\' +copy(trim(edit3.text),3,1)+'.wav';
          end; //case
          PlaySound((PChar(wfile2)), 0, SND_SYNC);

          case  strtoint(copy(trim(edit3.text),4,1)) of
            0..9:wfile2:= 'defaultWaves\' +copy(trim(edit3.text),4,1)+'.wav';
          end; //case
          PlaySound((PChar(wfile2)), 0, SND_SYNC);
          StatusBar1.Panels[1].text:='Finished Playing.';
        end;
    end;
end;
```

limited, but the main factor is the size of these files, almost one Mbyte per minute of recording. The keyer program provides a maximum 60 seconds of recording time, but you can modify the code to record larger sound bites.

The "TWaveRec.Button2Click(Sender: TObject);" procedure contains the routines to stop recording from the radio. The "TWaveRec.Button3Click(Sender: TObject);" procedure saves the newly recorded wave file to disk. The "TWaveRec.Button5Click(Sender: TObject);" procedure plays back any sound file.

The "PlaySound((PChar(wfile)), 0, SND_aSYNC);" code allows the wave file to be played asynchronously. Change the aSYNC to SYNC and the file will play synchronously until the sound bite is finished, but then the command "PlaySound(nil, 0, 0);" cannot be used to terminate file playing early. These sections of code are the guts of the recorder/player. The remaining project code makes everything look "pretty" and is described with comments within the source code.

Do you need more keyer messages? Four buttons to call wave messages are provided in the example software. Replicate the appropriate code (in the Msg# buttons), create additional "messages" and you've got them.

Don't like the *Voice Keyer/Recorder* control layout? Move the tool controls around almost anywhere within the *Explorer* design environment. Resize them, change the font style or size or hide them. This is the convenience in using an object oriented compiler. Move things around, engage the compiler "Run" and a new face to the program unfolds!

The methodology used to play and record wave files is simple and efficient. The "tMediaPlayer" control provided with *Explorer* could have been used to provide similar sound playback functions for the project using the *Windows* Media Player instead of *Windows* API calls.

Making Appropriate Voice Keyer Wave Files

The wave files included in the example "defaultWave" folder are recorded using my voice. You will want to record and edit similar files for your version of the recorder/keyer.

The program includes code that allows you to access S*ound Recorder,* the sound "editor" provided with various versions of the *Windows* operating system. Use *Sound Recorder* to edit wave files, except in *Windows Vista*. An editing tool could be programmed into the native code of the software but Microsoft includes what you need with the operating system. It is easy to take advantage of the simple utility. Access *Sound Recorder* with the button on the "Editing" panel or directly from *Windows* (START>Programs>Accessories>Entertainment>Sound Recorder).

Modify the characteristics of a wave file, edit it or examine the wave form with this tool. Plug a PC microphone (these microphones are relatively inexpensive at flea markets) into the microphone jack of your PC sound card and run *Sound Recorder* to record the basic sound bites, to use the software with wave files in your own voice. Record the ordinal numbers, 0 through 9 and save them as 0.wav, 1.wav, and so on in the program's "defaultWave" folder. Also record your call sign and any messages you may want to access using the voice keyer part of the project. Name them msg1.wav through msg4.wav and place these sound files in the "recWave" folder, so your program can find them.

Recording from a Radio

Connect your PC sound card at the microphone connector to the audio output of your

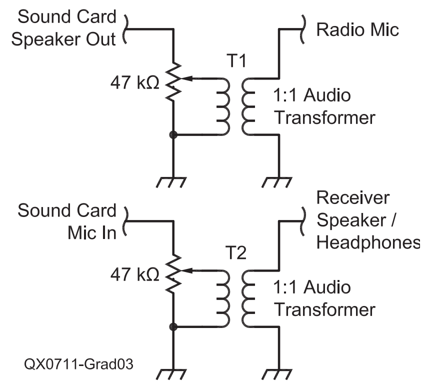

Figure 3 — A simple audio interface circuit will connect a radio audio output to a sound card microphone or line input. A similar circuit connects the sound card output to a radio microphone input circuit. T1 and T2 are RadioShack 273-1374 1:1 audio transformers.

WB5KIA

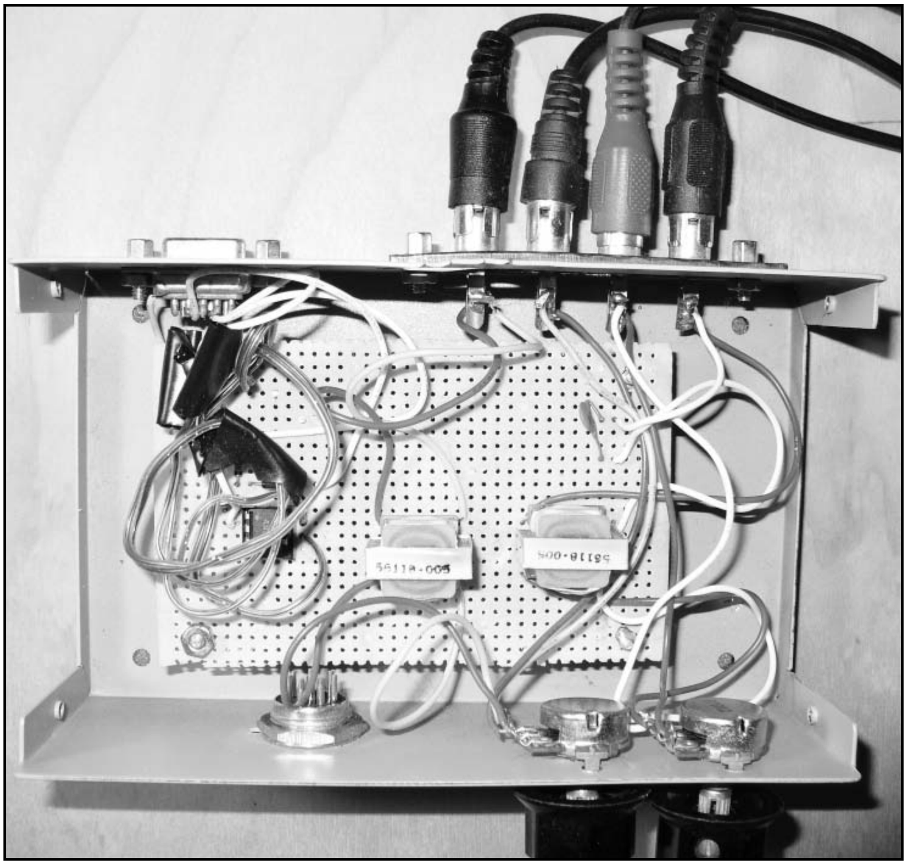

Figure 4 — The homebrew audio interface uses point-to-point wiring and basic construction techniques. Commercial units also work well. The interface need not be elaborate, just electronically stable.

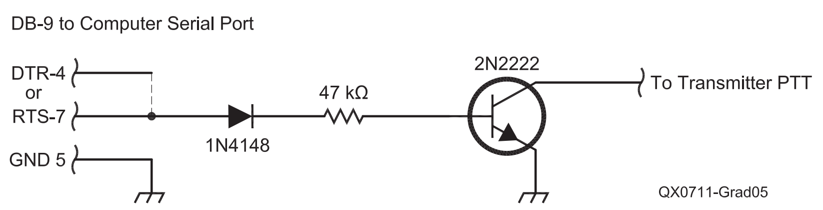

Figure 5 — A PTT interface circuit is shown in this diagram. An opto-isolator circuit could also be used. The circuit can be eliminated if VOX is used in conjunction with the voice keyer.

radio using the headphone jack, secondary speaker jack or other connection through a suitable interface. This interface can be a simple homebrew device (Figure 4) or a commercial interface designed for PSK or other digital operations. Do not connect the radio speaker output directly to your sound card. In addition to generating hum and ground loops, you are exposing yourself to the potential of destroying your sound card with excessive audio from the radio. The interface provides isolation and signal limiting. Use the program's "Record" panel and press the "Record" button. Use the "Stop" button to stop. You're not done yet! Use the "Save" button to save the sound bite, but be sure to name it in the box to the right first (callsign.wav as an example). Go to the Playback panel and use the "<Files" button to locate the wave file you just recorded and use "Play" to monitor the result. Delete a wave file previously recorded by typing its name in the box and pressing "Delete."

There is no audio VU meter to monitor recording level control. You have to experiment with sound card microphone levels accessible using the "Audio" button in the section devoted to recording (Sound recording).

Play Sound Back

No connections are required. Locate the appropriate sound file using the "<Files" button. Double-click the appropriate file on the resulting screen to cue the file for playback. Press the "Play" button.

Voice Keyer

You will need a connection from the sound card line output or speaker output to the microphone input jack on your radio. Many radios also have auxiliary microphone inputs. Again, you need a suitable interface, mainly to provide impedance matching and to avoid ground loops and hum. Turn on your transceiver's VOX circuit. Press a Msg# button (make sure you have prerecorded a "msg1.wave" type message using *Sound Recorder,* have named it appropriately (msg1, msg2, msg3, or msg4.wav) and placed the file in the program's recWaves file folder). Adjust the speaker or line output volume while the message is playing to trip your VOX and to ensure the ALC is working. The "Audio" button brings up the appropriate mixer screen.

The Voice Keyer panel can be detached from the form and moved anywhere on your screen using a left mouse click and dragging the panel.

PTT

Push-to-talk operation requires the use of a serial port and an interface to connect to your radio PTT circuitry. The interface can be a simple transistor switch (Figure 5) or a commercial interface. Additional code is needed to allow the software to change the state of the serial port RTS or DTR line from low to high. The PTT code is provided at my Web site, **qsl.net/wb5kia**.

WB5KIA

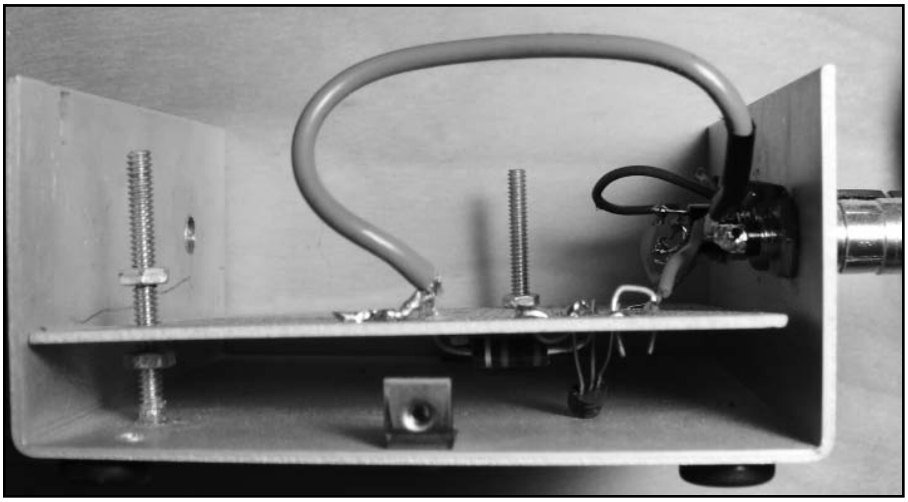

Figure 6 — This photo shows the inside of my PTT circuit built into a box. This could also be built into a serial connector hood.

Audio Interfaces

I built my audio interface from plans in an older version of *The ARRL Handbook*. Similar circuits are at the following Web sites:

www.w5bbr.com/soundbd.html, **www.mtn.org/handiham/psk-31_interface.htm**, and **www.waypoint.com/users/~discobay/Sound_Card_Presentation.htm**.

I recommend using a design for an audio cable that uses an audio isolation transformer (Figure 3). The transformer will reduce hum and ground loops. The variable resistors are used to limit the audio signal from the radio, which can potentially destroy a PC soundcard if the signal level is excessive. These interfaces can be built for about $10 to $20. They are not difficult to build. I also built the PTT interface in the same box — this is not audio but an electronic switch using a transistor or optoisolator to use for PTT with a PC COM port. The 2008 edition of *The ARRL Handbook* includes a multi-featured interface that you can build.[3]

Other people use commercial interfaces. They are more expensive, pretty and do the same job as a homebrew interface. People use MFJ, SignalLink and the various RigBlaster commercial interfaces. Most of these come with cables that can connect your radio to them. All of these solutions work. Find the one best for you.

Help with Delphi Programming

The Help system that is part of *Turbo Delphi Explorer* is its worst feature. Try **www.delphibasics.co.uk/** or visit **www.qsl.net/wb5kia** or the sites recommended in the Sep/Oct 2007 *QEX* article.

For You to Do

Voice Keyer/Recorder is a basic program that allows you to record from "the air." You can create a "play list" using *Explorer*'s file listboxes, add radio control (see the Sep/Oct 2007 *QEX* article about *Explorer*) and other enhancements like a recording VU meter. This is *your* program.

You downloaded *Explorer*. Consider copying the CD iso file to a CD and sharing the compiler with members of your radio club. CodeGear allows this. Individuals still have to register the software to get a software key to run it on their PC, but the key is free.

Steve Gradijan, WB5KIA, is an Extra Class licensee, and holds BS and MS degrees in geology. He has been developing Amateur Radio software since the early 1980s. He lives near Dallas, Texas.

Notes

[1]Steve Gradijan, WB5KIA, "*Turbo Delphi Explorer*: Develop Amateur Radio Projects for *Windows* with a Free Compiler," *QEX*, Sep/Oct 2007, pp 44-47.

[2]The source code as well as a compiled version of the "Voice Keyer/Recorder" program are available for downloading from the *QEX* Web site. Go to **www.arrl.org/qexfiles** and look for the file **11x07_Gradijan.zip**.

[3]Mark Wilson, K1RO, ed, "An Improved Digital Communications Interface," *The ARRL Handbook*, 2008 edition, pp 19.47-19.50. *The ARRL Handbook* is available from your local ARRL dealer, or from the ARRL Bookstore, ARRL order no. 1080. Telephone toll-free in the US 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

Signal Resilience to Ionospheric Distortion of HF Digital Chat Modes

This study may help you select a digital mode for your next keyboard-to-keyboard chat.

Daniel Crausaz, HB9TPL

The use of digital modes for HF QSOs is very widespread, with RTTY and PSK being the most popular. Comparison between them doesn't seem to have gone much further than dogmatic opinions and casual "on the air" tests with unqualified propagation conditions. Only one publication has been found to be dedicated to an extensive comparison of digital mode performance, but it doesn't take into account more recent developments like Olivia.[1]

HF digital signals are tiny high tech vessels, sent most of the time at low power and relying on the ionosphere's refracting properties to travel to the QSO partner. This article investigates how they are affected by ionospheric fading, and how they compare in transmitting information in several simulated propagation conditions that can be found on the HF bands.

Mode and Transmission Quality Criteria

The examined modes (or mode "flavors") are those allowing "chatting" or at least exchanging words at reasonable speed. Only the most common modes at the moment have been chosen. They are all available in *Multipsk*, free software that has been written by Patrick Lindecker, F6CTE. The selected modes are: BPSK31, RTTY (45 bauds), Olivia (1000 Hz, 32 tones), MFSK 16, DominoEX 11 and Feld-Hell. Technical descriptions of these modes are given in various texts.[2, 3] For a technical description of Olivia, see the Wikipedia entry.[4]

[1]Notes appear on page 45.

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The criteria to measure a transmission's quality is the character error rate (CHER), which is the number of characters that are different or missing in the received text due to the signal distortion resulting from the waves travelling in the ionosphere between the transmitter and the receiver.

A CHER of 2% seems to be adequate for chatting and contests while 5% may be tolerated for simple RST-name-QTH contacts, with the possibility of requesting a repeat for missing text. Here are some examples prepared by randomly introduced errors on a typical sentence to illustrate this choice.

Original

oh/dk4zc de hb9tpl/p how do you read this fred ? noise, bursts, qrm as well as multipath delays and dop pler spread affect the quality of the transmission btu fred oh/dk4zc de hb9tpl/p k

2% CHER

oh/dk4zc de hb9tnl/p how do you read this fredF? noise, bursts, qrm as well as multipath delays a(d dop pler spread affect the quality of the transwission btu fred oh/dk4zc de hb9tpl/p k

5% CHER

oh/dk4zc dN hb9tnl/p how do you read this =redF? no(se, bursts, qrm as well as multipath delays a(d dop pler Ppread affect the quality o(the transwission btuyfred oh/dk4zc de hb9tpl/p k

10% CHER

oh/dk4zc dN hb9tnl/p how do you read this =re{F? no(se, bursts, q%m as well as multipath 6elays a(d doE pler spread affec2 the quality o(tDe transwission btuyfred oh/dk4zcIde hb9tpl/p k

Effects Investigated

White noise (Additive White Gaussian Noise)

The ability to extract information from weak signals is a very interesting property of digital modes. It is measured by adding a desired level of white Gaussian noise to the information-carrying signal. Of course, white noise is the mildest manifestation of noise, but it's practical to use it as a basis for comparison. It is always present, whatever the other distortions or kinds of noise (QRM, lightning and so on) that are present.

Multipath Delays and Doppler Spread

HF waves traveling between the Earth's surface and the ionosphere are not reflected like the light rays of our elementary physics textbooks. The ionosphere is not a mirror, it's not homogenous at all, and the refracting media is moving. The transmitted wave will split into several signals traveling on different paths in the ionosphere. These signals may recombine at the receiver's antenna in a way that can deeply affect the received signal. The split signals will show different travel times. Complex refraction processes in nonhomogenous and moving parts of the ionosphere create Doppler spread: this leads to the broadening of the spectra (this effect can easily be observed on a pure carrier like WWV with the help of a spectral analysis software) and the apparition of phase modulation, which can be identified on PSK signals on a waterfall display: strong but fuzzy tracks and no possibility to decode them. See Figure 1.

Multipath delays have two effects: bit confusion (or intersymbol interference) and frequency selective fading. Bit confusion appears when the delayed signals superimpose on the "main" signal with such timing that bits cannot be separated anymore by the detector. Figure 2 illustrates this. Bits of 45 baud RTTY last 22 ms. If the replica arrives at the receiver with a delay similar to this figure, the detector will become completely confused. The conditions in which this effect can be observed are discussed below.

Multipath delays can easily be observed now and then on Feld-Hell print: the replica will appear as a "ghost" of the stronger signal. See Figure 3 for an example of this "ghosting."

Frequency selective fading is a bit more complicated. When two signals or signal components add with one having a 180° delay with respect to the other, they will cancel each other. Similarly, when two components of a nonmonochromatic signal are added, with one component delayed respect to the other, some frequencies will be enhanced and others will be cancelled. Cancellation appears at frequencies equal to n × 1000 / [2 × delay (in ms)] where n is an even integer. Table 1 lists some frequencies while Figures 4 and 5 illustrate this effect with a broadband signal. MT63 is perfect for this purpose. "Holes" or "canyons" in the case of 3D spectrograms, appear in the spectrum of the signal after splitting it in two components, delaying one of them and then summing them again, exactly as predicted in Table 1.

If Doppler spread is added, the canyons will start moving in the spectrum, as illustrated by Figure 6. If the value of the Doppler spread increases, the canyons will appear in a much more irregular way.

The theory underlying these phenomena is well known and accessible background can be found in texts such as *Communication Systems Engineering*, and *Fundamentals of Wireless Communication*, for instance.[5, 6]

Long haul contacts involving several hops are particularly subjected to important multipath delays. To the extreme, signals running the short and the long path can combine at your partner's antenna (located near the antipodes) with delays up to 50 ms.

Near vertical incidence signal (NVIS) propagation is used on short distance HF communication in difficult topographical conditions. The signal coming back from the ionosphere can interact with the ground wave with delays extending from 5 up to 20 ms.[7]

Doppler spread strongly affects transmissions on polar routes or those crossing the aurora oval due to the importance of ionized layers and spot movements. The Doppler spread value can be as high as 50 Hz in such circumstances.

Multipath delays and Doppler spread are not restricted to extreme paths. Extensive measurements have been conducted by E. M. Warrington and others, on a Leicester (UK) to Uppsala (Sweden) path, a modest 1440 km distance.[8] Delays up to 8 ms and Doppler spreads up to 10 Hz have been measured (with seasonal and day/night patterns). Such values can affect some digital

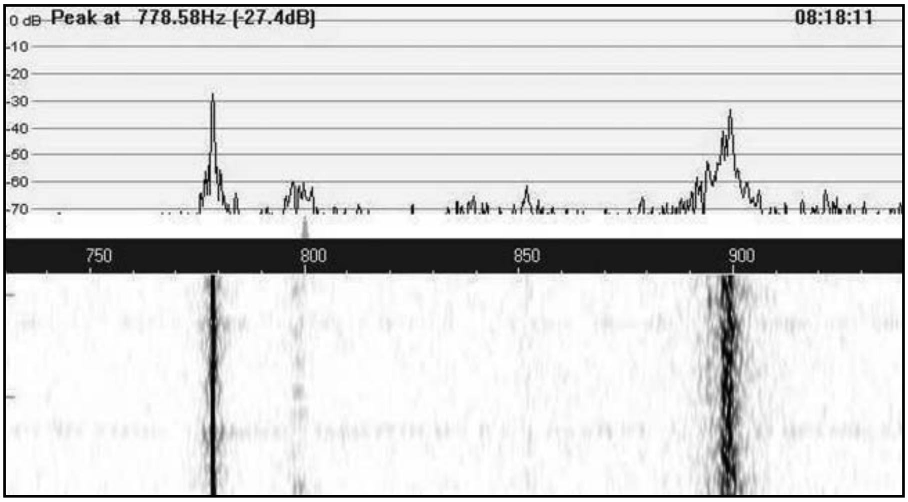

Figure 1 — Two HF signals side by side on this waterfall display. The signal on the left is a "sharp" one and the one on the right is affected by Doppler spread. (Horizontal scale units are in Hertz.)

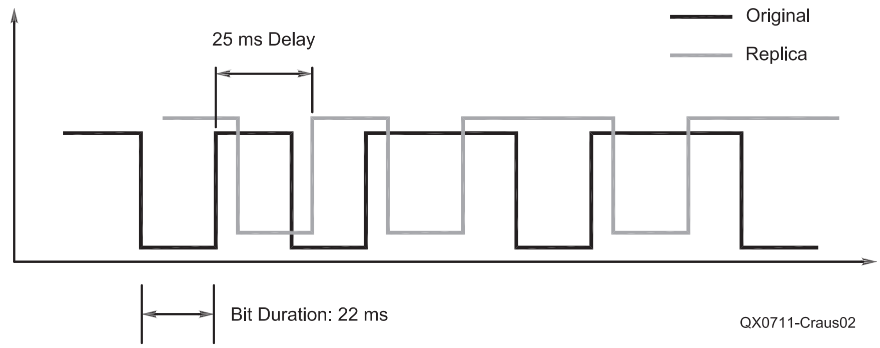

Figure 2 — This diagram illustrates the loss of identity of an RTTY character due to the delay between the two HF path components.

Figure 3 — Ghosts appearing on a Feld-Hell transmission due to a 12 ms delay between the two signals. (The delay was simulated to create this display.)

Table 1

Null Frequencies (in Hz) of the Sum of the Main and the Delayed Signal

	Delay (ms)								
n	0.1	0.5	1	2	3	5	6	7	10
1	5000	1000	500	250	167	100	83	71	50
3	15000	3000	1500	750	500	300	250	214	150
5	25000	5000	2500	1250	833	500	417	357	250
7	35000	7000	3500	1750	1167	700	583	500	350
9	45000	9000	4500	2250	1500	900	750	643	450
11	55000	11000	5500	2750	1833	1100	917	786	550

modes to the point that they cannot be used.

Mid latitude paths are also affected, though less severely. The delay is usually less than of a few miliseconds and the Doppler shift reaching a maximum of 5 Hz.

Experimental Setup

The commonly accepted way to simulate HF fading channels is the Watterson channel model.[9, 10] A signal is split in two identical parts. Delay is introduced between the two signals, and frequency dispersion is added as well as Gaussian noise. The two signals are then summed before being submitted to the decoder.

This task is performed by a "path simulator," which is a computer program processing an audio signal, splitting it in components that can be delayed and affected by Doppler spread and noise before being added again. The PathSim program has been developed by Moe Wheatley, AE4JY, and can be downloaded freely from his Web site: **www.qsl.net/ae4jy/files**

The digital mode signal is produced using *Multipsk*, a free program for digital mode operation by F6CTE. (See Note 2.) It is then treated by PathSim, the software that does the work of simulating the ionosphere. The resulting signal is stored as a .WAV file and than played back and decoded by by *Multipsk*. The transmitted "QSOs" consist of strings of the character "8", which can easily be counted by the dedicated function available in *Multipsk*.

The 2% CHER threshold in Average White Gaussian Noise has been determined by using a 525 character string, which ensures a 95% confidence interval when 5 or less error numbers are found. There is a Maxim Application Note that explains the underlying mathematics.[11] Otherwise, a more quickly transmitted 200-character-long transmission has been used.

Characters transmitted by Hellschreiber modes cannot be counted by the computer, and some subjective interpretation of what appears on the screen cannot be avoided.

Propagation Conditions Explored

Mapping results over a wide range of signal to noise ratio (SNR), multipath delays and Doppler spread values requires computing skills that are beyond mine. I found it useful and illustrative to select a set of known quantified conditions defined by the ITU or other sources.[12, 13] The following conditions were used.

A) Pure Additive White Gaussian Noise conditions (AWGN)

B) Mid latitudes good conditions

C) High Latitude Moderate

D) Flutter

E) Fading channel, as described by the ITU Radiocommunications Group.[14]

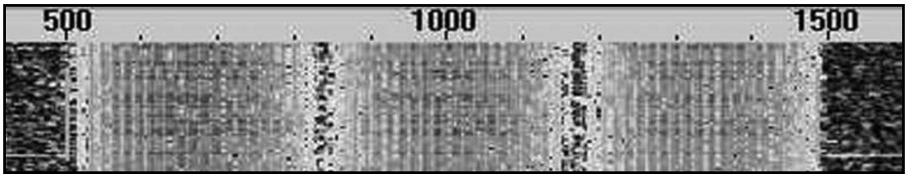

Figure 4 — A waterfall picture of MT63 signals, with 3 ms delay, showing nulls at 833 Hz and 1167 Hz.

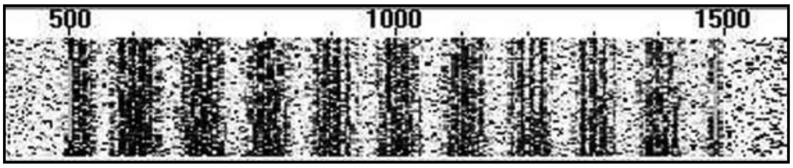

Figure 5 — A waterfall picture of MT63 signals, with 10 ms delay, showing nulls at 550 Hz, 650 Hz and so on.

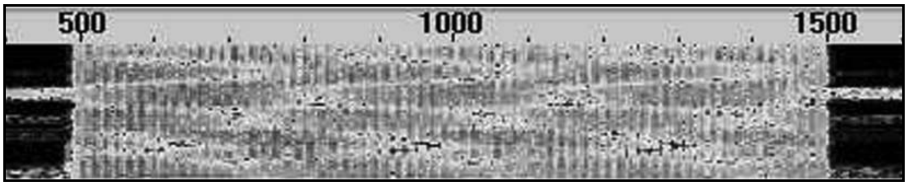

Figure 6 — This MT63 signal shows selective fading (3 ms delay, 0.5 Hz Doppler spread).

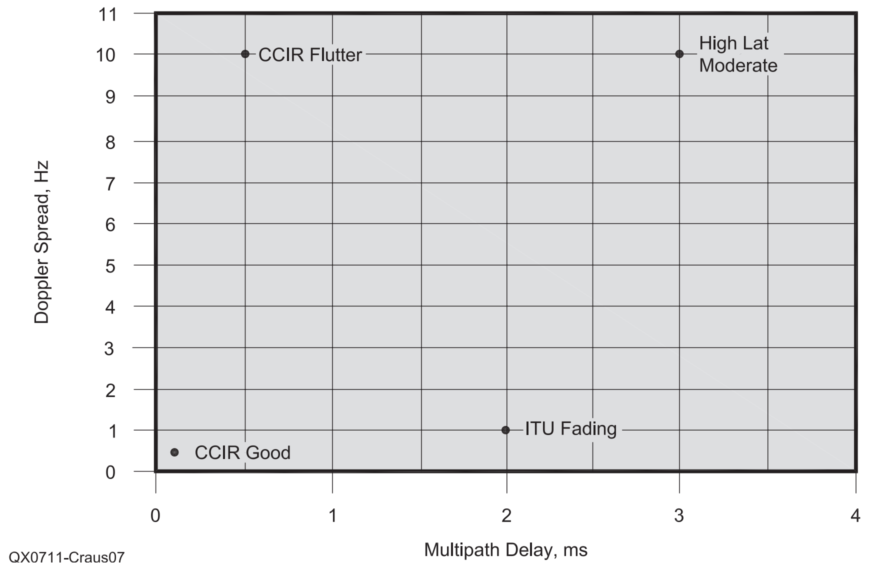

Figure 7 — Standardized Ionospheric distortion parameters.

Table 2
Standardized Ionospheric Distortion Parameters

	Mutipath Delay	*Doppler Spread*
CCIR Good Conditions	0.1 ms	0.5 Hz
ITU Fading Channel Conditions	2.0 ms	1.0 Hz
CCIR Flutter Conditions	0.5 ms	10.0 Hz
High Latitude Moderate Conditions	3.0 ms	10.0 Hz

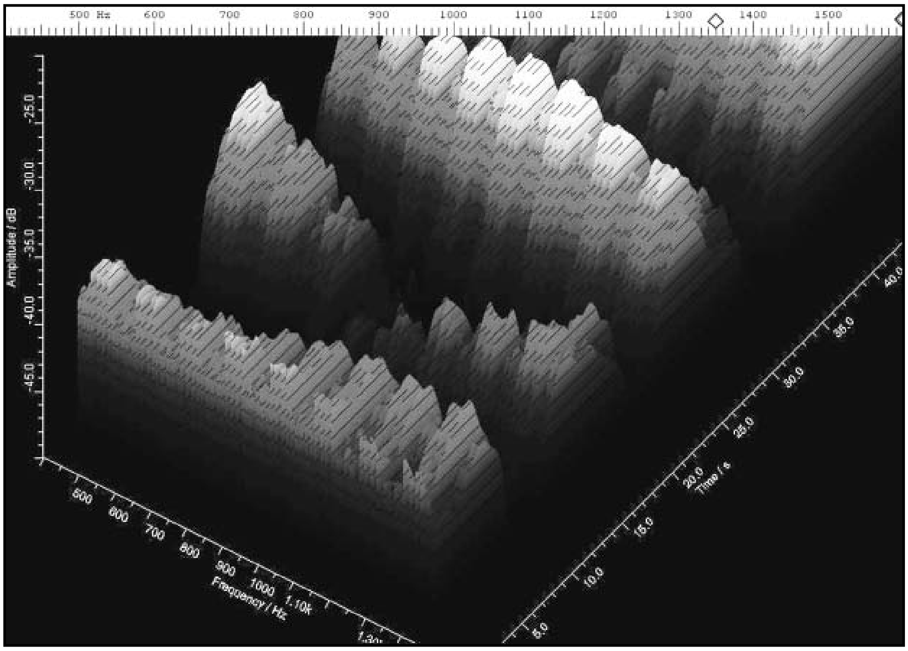

Figure 8 — The effect of CCIR good conditions on an MT63 signal (9 dB SNR). Note the 10 s interval between the crests due to Doppler spread, as well as the selective fading notch.

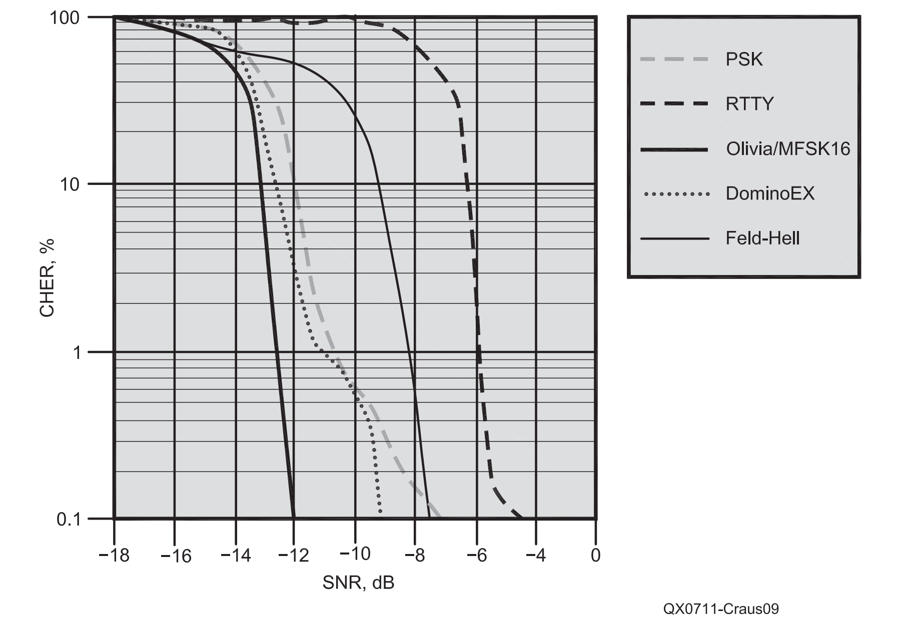

Figure 9 — Character Error Rate (CHER) of various digital modes as a function of signal to noise level in Average White Gaussian Noise conditions.

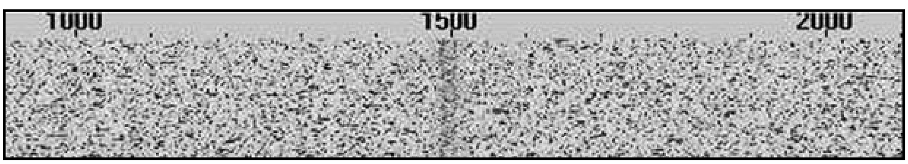

Figure 10 — Waterfall screenshot of a BPSK31 signal at –12 dB signal to noise ratio.

Conditions B) to E) include both multipath delays as well as Doppler spread, as stipulated in Table 2 and illustrated in Figure 7.

Multipath delays and Doppler spread contribute in a quite different way to the fading shape. In the time domain, Doppler spread creates a wave-like amplitude pattern that has the periodicity of the inverse of the value of the spread. If Doppler spread is 0.1, the amplitude will enhance every 10 seconds for instance. Short multipath delays cause "holes," due to selective fading in the spectrum.

Figure 8 shows the effect of CCIR good conditions on an MT63 signal with a 9 dB SNR. Note the deep fades and the 10 second interval between the crests. This is the result of Doppler spread as well as selective fading.

Results

All signal-to-noise ratios refer to a 3kHz bandwidth. To change to another bandwidth subtract 34 dB and add 10 log of the desired bandwidth in Hz.

Additive White Gaussian Noise (AWGN)

Figure 9 illustrates the typical threshold behavior of digital modes, going from about 100% of errors to less than 2% with a difference of only 2 dB in SNR. This effect can be noticed on the air with one partner copying perfectly while the other receives only random characters, due probably to a somewhat weaker power or a less efficient antenna at one extremity of the path.

Those used to digital modes will agree that the signal of Figure 10 at the 2% CHER threshold is a quite faint one, and that very quiet conditions are required to decode it.

Olivia and MFSK16 are the most sensitive modes, RTTY being the worst. The difference between them is about 6 dB. In other words, RTTY needs four times more power than MFSK or Olivia for the same result. No wonder the use of a 1 kW linear amplifier is widespread among RTTY operators. Table 3 gives a list of the 2%

Table 3
SNR level for a 2% Character Error Rate in Additive White Gaussian Noise (95% confidence level)

Mode	*Signal to Noise Ratio (3 kHz bandwidth)*
PSK 31	–11
RTTY	–6
Olivia 1000/32	–13
Olivia 500/16	–13
Olivia 500/8	–11
DominoEX	–11
DominoEXFEC	–11
MFSK 16	–13
Feld-Hell	–9
FM Hell	–10

CHER threshold for a wider list of modes.

Standardized Ionospheric Conditions

Introducing ionospheric distortion leads to a very different picture, as Figure 11 shows for PSK and RTTY.

The curves are shifted to the right indicating the need for more signal strength to keep the same rate of errors. The 2% CHER threshold changes from –12 dB to + 3 dB for PSK and from –7 to + 9 dB for RTTY as compared to the values measured in additive white Gaussian noise (AWGN).

The differences are quite important. It must be pointed out that the term "good conditions" is misleading, as Figure 8 shows. There is a rather deep, slow periodic fading, and slowly oscillating selective fading that affects broadband signals.

Table 4 indicates the required SNR for the 2% CHER threshold of the modes under study. Table 5 indicates the difference in SNR of the 2% threshold between AWGN and the several standardized ionospheric conditions.

The increase of SNR needed to maintain the same transmission quality is quite different according to the modes and the conditions. PSK, Domino Ex and Feld-Hell could not even reach the 2% CHER threshold at a 12 dB SNR. Except for Olivia — and to a lower extent MFSK — the necessary increase in signal levels are impressive when conditions are not perfect. Let's see how the different modes react to changing ionospheric parameters.

PSK

The performance of PSK shows an important degradation when conditions are not perfect. This is illustrated by Figure 12. PSK cannot be copied at all in high latitude moderate, as well as in flutter conditions, no matter the strength of the signal. These are typical polar path conditions characterized by an important amount of Doppler spread (10 Hz). PSK doesn't meet the 2%CHER threshold even with signals 24 dB stronger than in the AWGN conditions.

In less extreme conditions (ITU fading conditions) it will keep showing errors even if signal strength is increased. (This could be easily handled by using a Forward Error Correction process as proposed by *Multipsk*, but unfortunately seldom used on the air).

RTTY

Although less sensitive in quiet conditions, RTTY will manage to come through very tough conditions provided the signal is strong enough. It's not surprising that true RTTY aficionados use high power. Figure 13 shows that RTTY has similar (limited) performance in all the examined conditions.

Olivia and MFSK

Olivia and MFSK are particularly resilient to ionospheric distortion as shown by Figures 14 and 15.

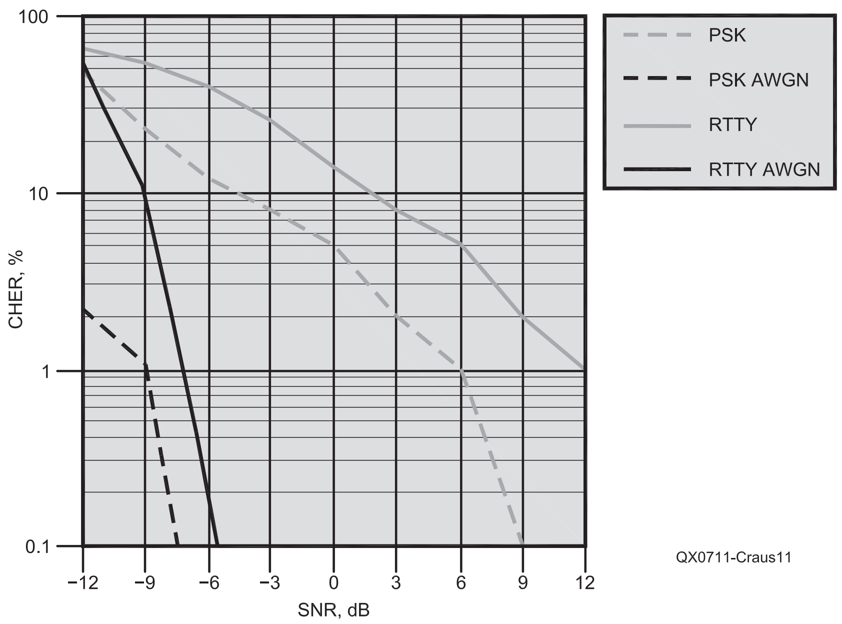

Figure 11 — Effects of ionospheric distortion (CCIR good conditions) on character error rate (CHER) for PSK and RTTY.

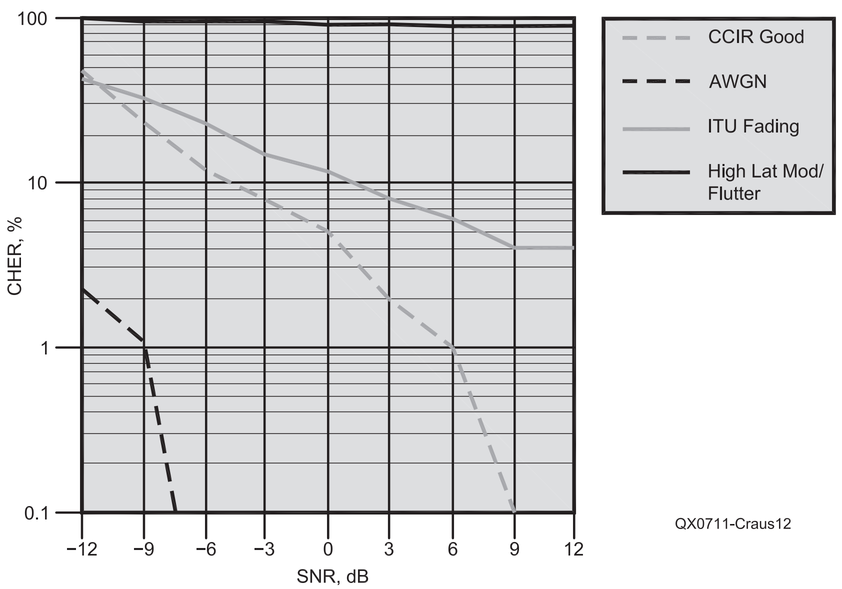

Figure 12 — PSK performance under various conditions.

Table 4
Required SNR in dB for a 2% Character Error Rate

	Olivia	*MFSK*	*DominoEX*	*PSK*	*RTTY*	*Feld-Hell*
AWGN	–13	–13	–11	–11	–6	–9
CCIR good conditions	–7	–5	2	3	9	5
ITU fading channel	–10	–7	6	*	10	*
Flutter	–10	–5	*	*	6	*
High latitude moderate	–10	–5	*	*	12	*

* 2% CHER threshold not reached at 12 dB SNR

It should not be surprising that Olivia's performance happens to be worse in CCIR good conditions than in more extreme ones. This is due to the slow and broad selective fading effect that is associated with "good conditions." Large portions of the signal completely disappear in the notch. (See Figure 8.)

DominoEX

DominoEX is quite similar to PSK: it doesn't transmit any information in Flutter and High Latitude Moderate conditions. Figure 16 shows the performance of DominoEX.

Feld-Hell

Like RTTY, Feld-Hell is robust provided the signal is strong enough. See Figure 17.

Side by Side Comparisons

Now that we have a feeling of how the different modes react in a different way to ionospheric parameters let's see how they compare when placed in similar conditions.

CCIR Good Conditions

To say in words what Figure 18 shows, lets see what will happen with a signal at –6 dB SNR: copy is perfect in Olivia, still good in MFSK, rough in PSK and impossible in Feld-Hell, RTTY and DominoEX

Fading Channel Conditions

The fading conditions are being defined by the ITU as a standard for commercial digital modes comparisons. (See Note 14) They are characterized by a rather important amount of multipath delay and significant Doppler spread. These conditions are likely to be found on DX routes over the pole in rather good conditions, like Europe to the US or Canadian west coast.

Fading contains a quick "slicing" of the signals as well as changing selective fading as illustrated by Figure 19. Observe the "streets" separated by short time intervals and the more irregular dark canyons due to selective fading.

As Figure20 illustrates, MFSK and Olivia give excellent copy at SNRs down to –8 or –9 dB. At the other end, PSK, RTTY and Feld-Hell need a 20 dB increase in signal strength to be at best readable. In other words, if my signals can be read while transmitting with 20 W in MFSK or Olivia, I need more than 2.5 kW to get comparable results with the other modes. PSK and Feld-Hell do not converge towards an error free transmission in these conditions.

Flutter

Flutter is a quick fading that can affect polar or transequatorial paths. Flutter and high latitude moderate conditions are characterized by the same amount of Doppler spread (10 Hz) but differ by the multipath delay (0.5 ms for flutter, 3 ms for high latitude moderate conditions). Figures 21 and 22 are quite similar, suggesting that Doppler spread is the

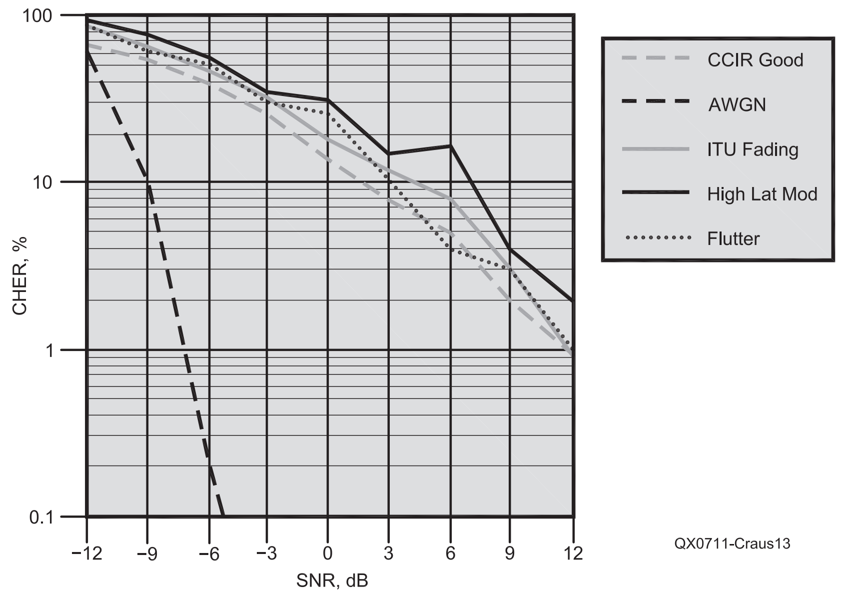

Figure 13 — RTTY performance under various conditions.

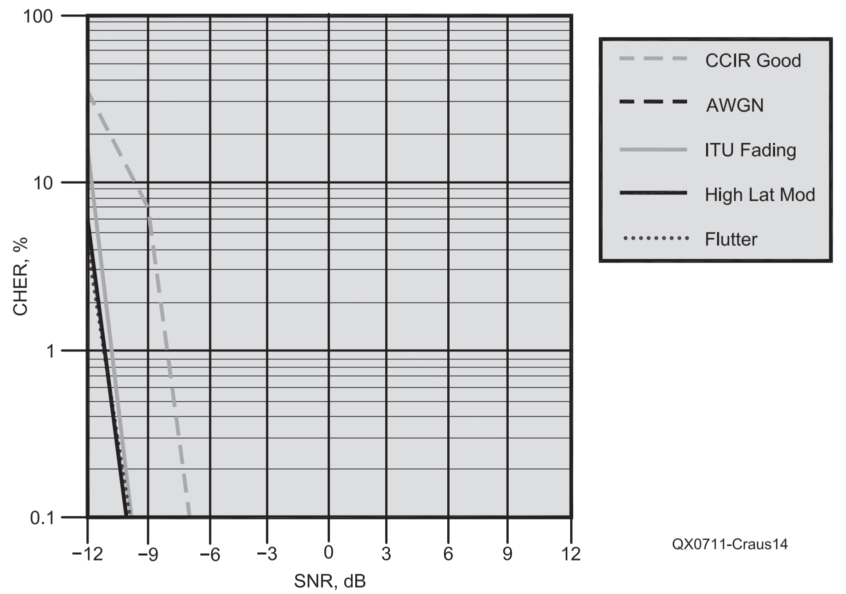

Figure 14 — Olivia performance under various conditions.

Table 5

Difference between the thresholds levels in the indicated conditions and the AWGN case (SNR in dB).

	Olivia	*MFSK*	*DominoEX*	*PSK*	*RTTY*	*Feld-Hell*
CCIR good conditions	4	5	6	9	6	12
ITU fading channel	3	6	17	*	16	*
Flutter	3	8	*	*	12	*
High latitude moderate	3	8	*	*	18	*

* 2% CHER threshold not reached at 12 dB SNR

main factor determining the performance in such conditions.

High Latitude Moderate

These conditions are characterized by a substantial amount of Doppler spread and some multipath delay (10 Hz and 3 ms).

As shown by Figure 22, PSK and DominoEX don't come through no matter how strong the signals are. RTTY and Feld-Hell show reasonable performance provided the signal is strong enough. Olivia's performances are unaffected while MFSK is a little bit worse but still gives perfect copy with a signal to noise ratio more then 15 dB weaker than needed in RTTY.

Analysis of Individual Contributions

Multipath Delays

Although multipath delays and Doppler spread cannot be entirely dissociated, it has been found useful to investigate, somewhat artificially, their individual contribution to ionospheric distortion.

To measure the effect of delays only, various time differences have been introduced between two signals (0 dB SNR) before adding and analyzing them. The results are given in Table 6. Remember that the standardized conditions explored here above do not include delays in excess of 3 ms. Delays longer than 10 ms can only be observed in special case like NVIS or simultaneous short and long path propagation. Taking this into consideration leads to the conclusion that multipath delays alone do not contribute to reading errors for the modes observed in usual conditions, provided the signal doesn't fall in a notch due to selective fading.

RTTY's poor performance from 15 ms delay are related to the 22 ms bit duration in 45 baud mode: bits lose their identity in such circumstances as discussed earlier. The absolute nonperformance at 3 ms is an illustration of frequency selective fading: RTTY signals consist of two main peaks separated by 170 Hz.

As the cursor was placed on the *Multipsk* waterfall at 1000 Hz during the test, the upper component was at 1170 Hz, exactly in the null due to selective fading with a 3 ms delay, as illustrated in Figure 4. The resulting signal had only one frequency component, and the loss of information was total. In real conditions such effects are less severe because the notch is moving across the spectrum.

Doppler Spread

Doppler spread causes two effects: a periodic fluctuation of the received signal level (time domain) and a broadening of the spectra (frequency domain).

As illustrated by Figure23, Olivia and MFSK are not affected at all by Doppler spread. Although PSK and DominoEX don't

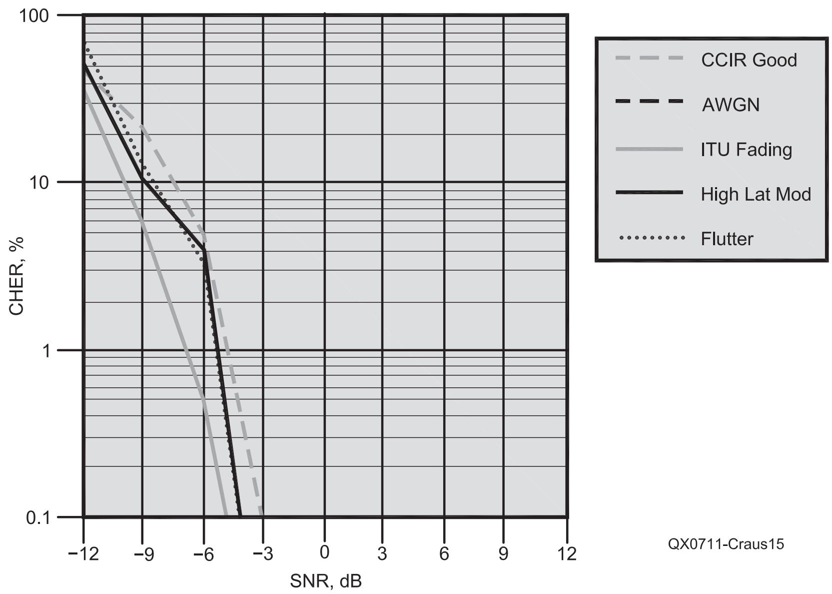

Figure 15 — MFSK performance under various conditions.

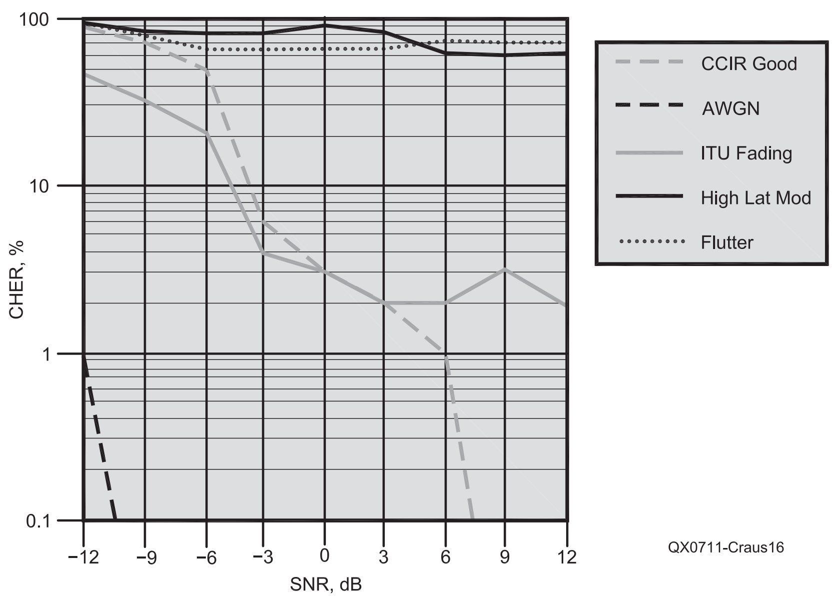

Figure 16 — DominoEX performance under various conditions.

Table 6
Character Error Rate as a Function of Multipath Delay (at 0 dB SNR)

Muliple Path Delay (ms)

Mode	1	3	5	7	10	15	20	25	30	40	50
PSK	0	0	0	0	0	0	0	20	100	100	100
RTTY	0	100	0	0	0	10	41	37	98	100	100
Olivia	0	0	0	0	0	0	0	0	0	0	7
MFSK	0	0	0	0	0	0	0	0	0	43	96
DominoEX	0	0	4	34	25	12	40	25	73	63	76
Feld-Hell	0	0	0	5	100	100	0	100	100	0	0

suffer too much from Doppler spread under 1 Hz, they become impossible to use with higher Doppler spread values. Clearly, PSK is affected by the random phase modulation associated with Doppler spread. The reasons for this DominoEX behavior are still unclear to the author. Feld-Hell and RTTY are affected in a rather constant way. This confirms what have been observed before for flutter and high latitude moderate conditions.

Discussion

Digital modes show considerable differences in the way they are affected by ionospheric distortion, and the required signal strengths to have a QSO are quite different from that measured in the standard additive white gaussian noise conditions. In such conditions, the modes examined show good reading thresholds ranging from –13 dB SNR (Olivia) to –6 dB SNR (RTTY) with most modes allowing contacts below –10 dB SNR. These conditions are likely to be found in 1-hop paths in moderate latitudes, so those prevailing on a majority of contacts held in the ham community, and help understanding why working QRP can be so effective in digital modes.

Among the factors characterizing ionospheric distortion, it has been found that Doppler spread is the main contributor to introducing errors in reception. Olivia, and to a somewhat lesser extent MFSK, shows an extraordinary ability to come through in Doppler spread affected paths like polar routes with quite weak signals. Feld-Hell and RTTY behave in a quite similar way, and are not too bad on polar routes. They need SNRs stronger than 15 to 20 dB, however, to give the same results as Olivia and MFSK in such circumstances.

PSK and DominoEX show similar performances: good in quiet conditions even with weak signals, but unable to come through under tough conditions.

Can these conclusions help in predicting the ability of one mode to perform well under certain circumstances? Not much, as we have no direct means to measure the parameters (multipath delay and Doppler spread) that characterize the ionosphere. A broadband signal should be monitored to get a feeling of what the ionosphere is doing along the particular path.

The RSGB MT63 daily bulletins and military "stanag" signals are potentially useful.[15] DRM broadcast transmissions are also useful, as suggest by John Stanley in the Jan/Feb 2007 issue of *QEX*.[16] Interpreting pictures like Figure 24 in terms of Doppler spread and multipath delays is the scope of a future project. Observing the broadening of a pure carrier is very interesting too. WWV signals are perfect for this, although the path they help characterize is only the one between Fort Collins Colorado and your receiver (Figure25).

The path that the signal will travel is a factor of choice. As an illustration, KH6CW could be heard in Europe late in the afternoon at the end of February 2007 with signals ranging from S7 to S9 levels on a dipole but could hardly be decoded. The PSK signal appeared fuzzy on the waterfall. Traveling across the auroral doughnut destroyed the phase pattern of the signal. On a similar path, 3D2BA can be heard nearly every morning in Europe for long chats using Olivia.

Acknowlegments

This project has been made possible by the extraordinary qualities of the following free software:

Multipsk by F6CTE; a digital mode coder/decoder: **f6cte.free.fr/**

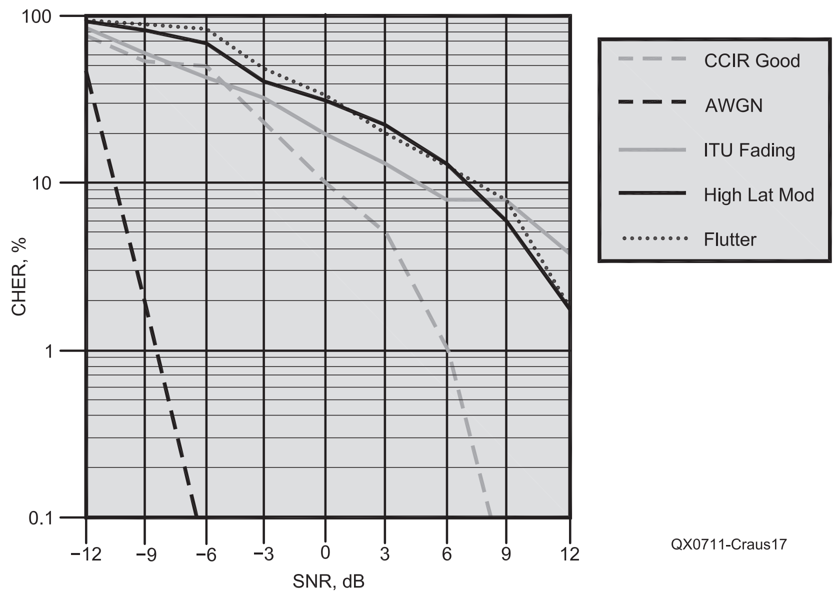

Figure 17 — Feld-Hell performance under various conditions.

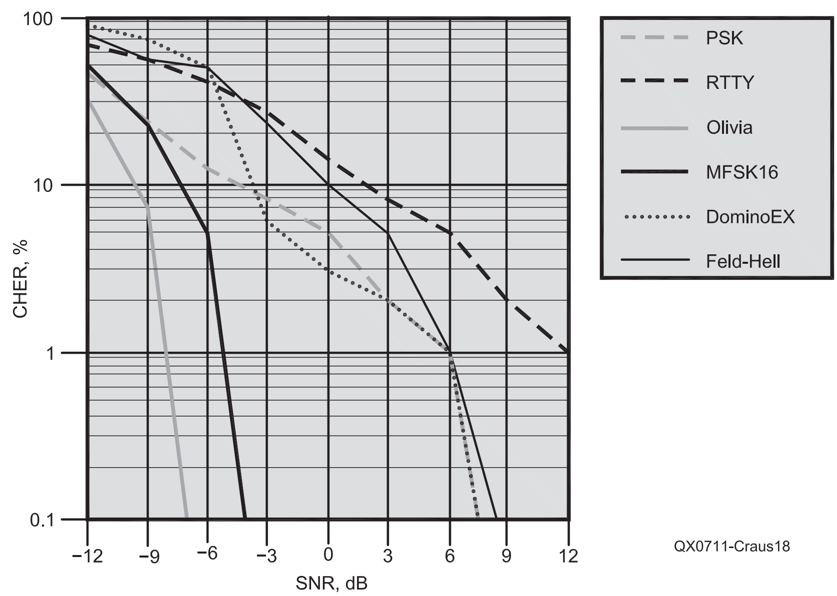

Figure 18 — Character Error Rate in CCIR good conditions (0.5 ms multipath delay and 0.1 Hz Doppler spread).

Spectran by KA7OEI, for spectrum analysis; **www.weaksignals.com/**

Cadet, a convolution and deconvolution imaging software aimed at astrophotography: **www.terra.es/personal2/oscarcj/introeng.htm**

Irfan, image processing software well know by those practicing digital SSTV: **www.irfanview.com/**

PathSim by AE4JY, a Watterson model channel simulator: **www.qsl.net/ae4jy/pathsim.htm**

Special thanks to Fred, N9GUE, and his wife Carol for proofreading the text as well as the numerous tests on 80 m. Thanks also to Fred, OH/DK4ZC, for his always positive and professional attitude.

Notes

[1]Steve Richards, G4HPE, "A practical evaluation and comparison of some modern data modes," **www.rsgb.org/emergency/operating/datamodes/datmodes2.pdf**

[2]See **f6cte.free.fr**

[3]See ZL1BPU's homepage: **www.qsl.net/zl1bpu/**

[4]For a technical description of Olivia see: **en.wikipedia.org/wiki/Olivia_MFSK**

[5]John G. Proakis, *Communication Systems Engineering*, "Chapter 10 — Digital Transmission on Fading Multipath Channel," Prentice Hall, 2002. This chapter is available at: **zone.ni.com/devzone/cda/ph/p/id/61**

[6]David Tse, Pramod Viswanath, *Fundamentals of Wireless Communication*, Cambridge University Press, 2005.

[7]Murray Greenman, ZL1BPU, *Digital Modes*, RSGB, 2004.

[8]E.M. Warrington and others, "Propagation of HF Radio Waves Over Northerly Paths," **www.cost271.rl.ac.uk/AugMeeting/final_meeting_presentations/S5_P5_Warrington.pdf**

[9]Clark C. Watterson and others, "Experimental Confirmation of an HF Channel Model," *IEE Transactions on Communication Technology*, Vol Com-18, no. 6 Dec 1970, pp.792-803.

[10]See the articles by Johann Forrer on his Web site: **www.johanforrer.net/SIMULR/index.html**

[11]Dallas Semiconductor/Maxim, Application Note 1095, "Statistical Confidence Levels for Estimating BER Probability, Oct 2000," **pdfserv.maxim-ic.com/en/an/AN1095.pdf**

[12]Moe Wheatley, AE4JY, PathSim User and Technical Guide, Ver 1.0 Dec 1, 2000, **www.qsl.net/ae4jy/files/pathsimtech100.pdf**

[13]CCIR Recommendation 520-1; Use of High Frequency Ionospheric Channel Simulation.

[14]ITU Radiocommunications Group, "Bandwidths, Signal-to-Noise and Fading Allowances in HF Fixed Systems," Draft revision of recommendation ITU-RF339-6, January 2006.

[15]Such signals can be heard day and night, in Europe on 3820 kHz, 7555 kHz and 12725 kHz. Unfortunately, the transmitter locations are not known to me.

[16]John Stanley, K4ERO, "Observing Selective Fading in Real Time with Dream Software," *QEX*, Jan/Feb 2007, pp 18-22.

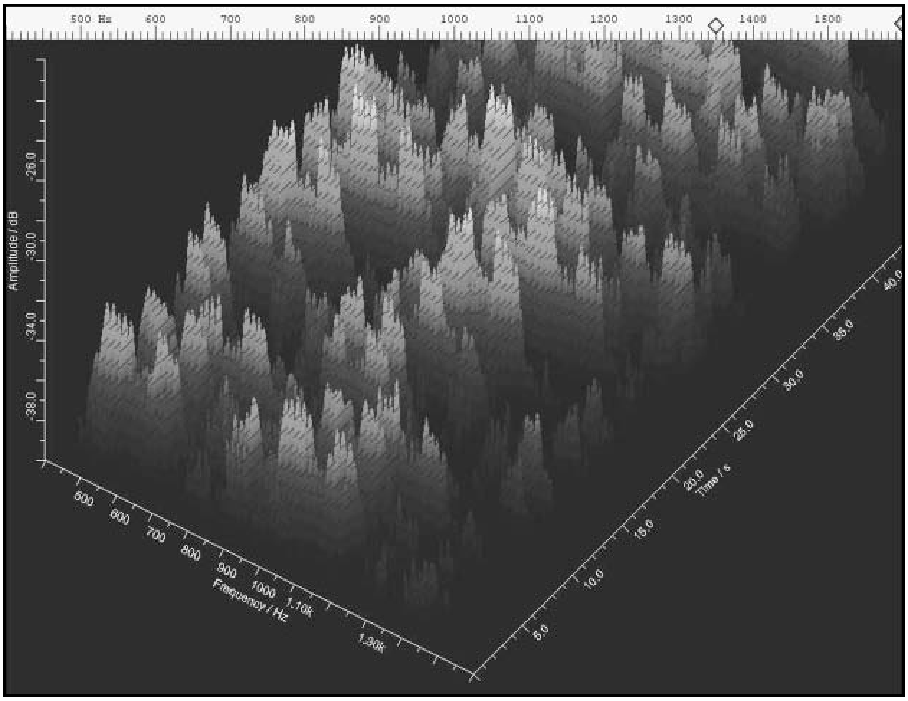

Figure 19 — MT63 signals at 9 dB SNR in ITU fading channel conditions.

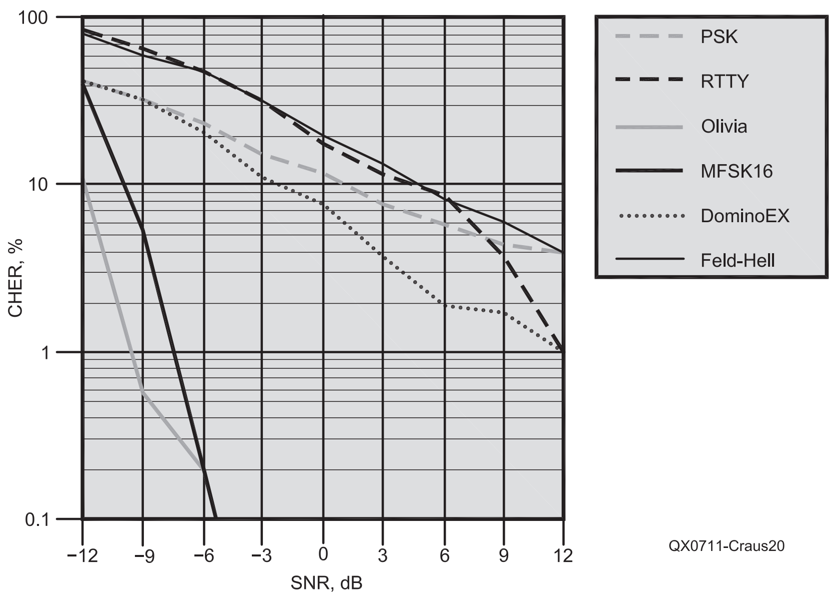

Figure 20 — Character error rate (CHER) in fading channel conditions (2 ms delay, 1Hz Doppler spread).

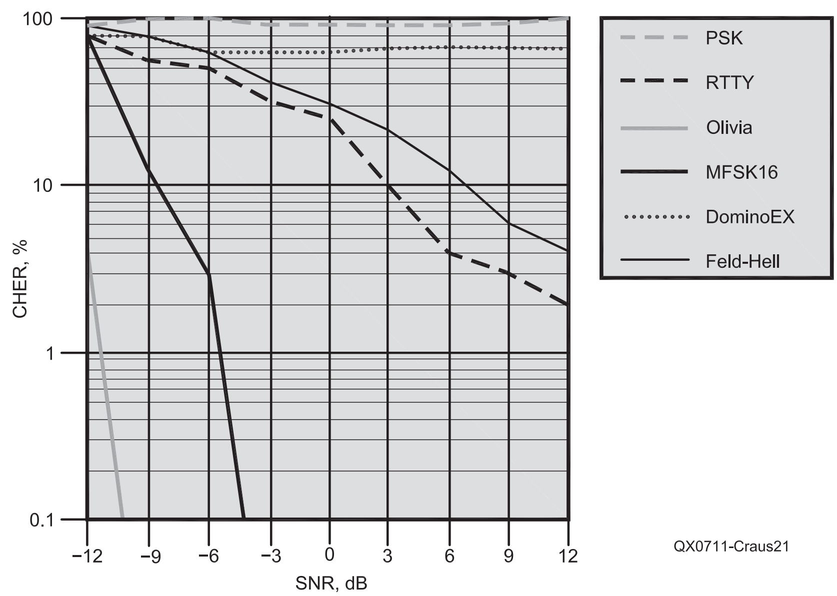

Figure 21 — Character error rate in fluttering conditions (0.5 ms multipath delay and 0.1 Hz Doppler spread).

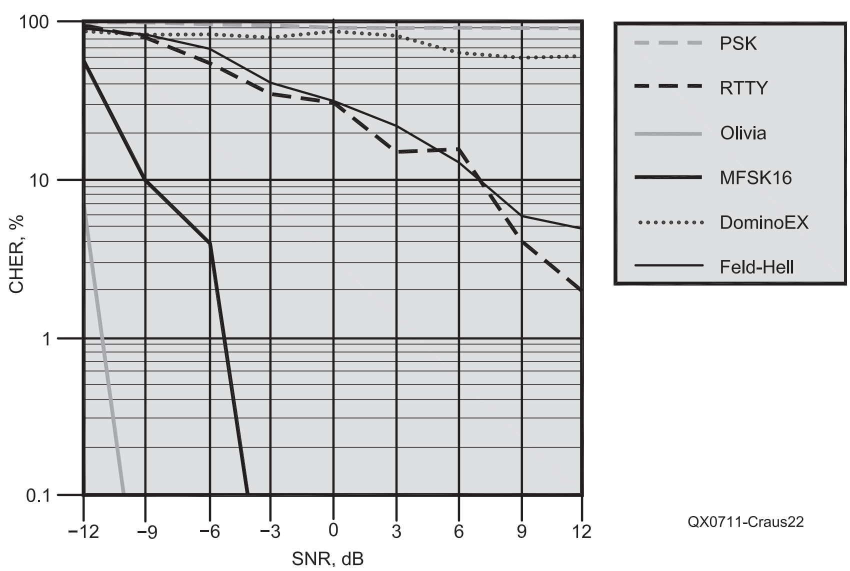

Figure 22 — Character error rate in high latitude moderate conditions.

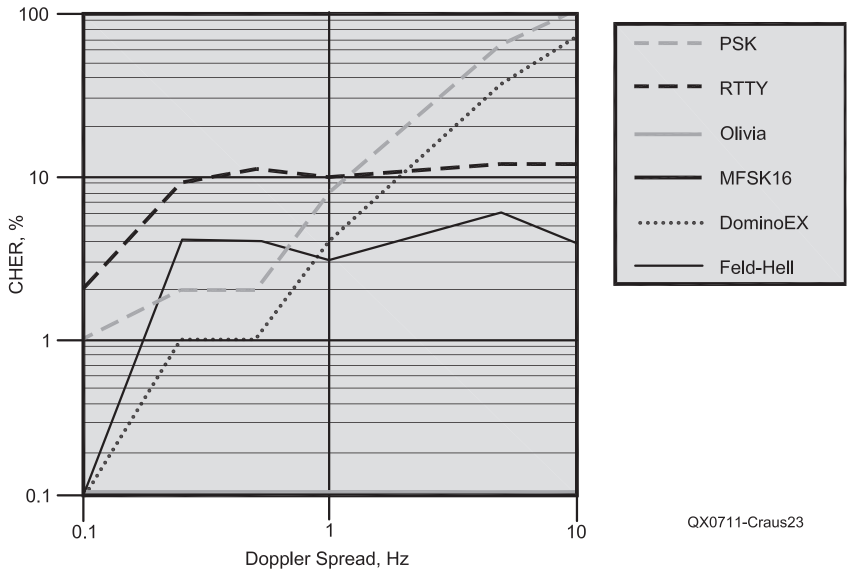

Figure 23 — Sensitivity of the different modes to Doppler spread at 3 dB SNR.

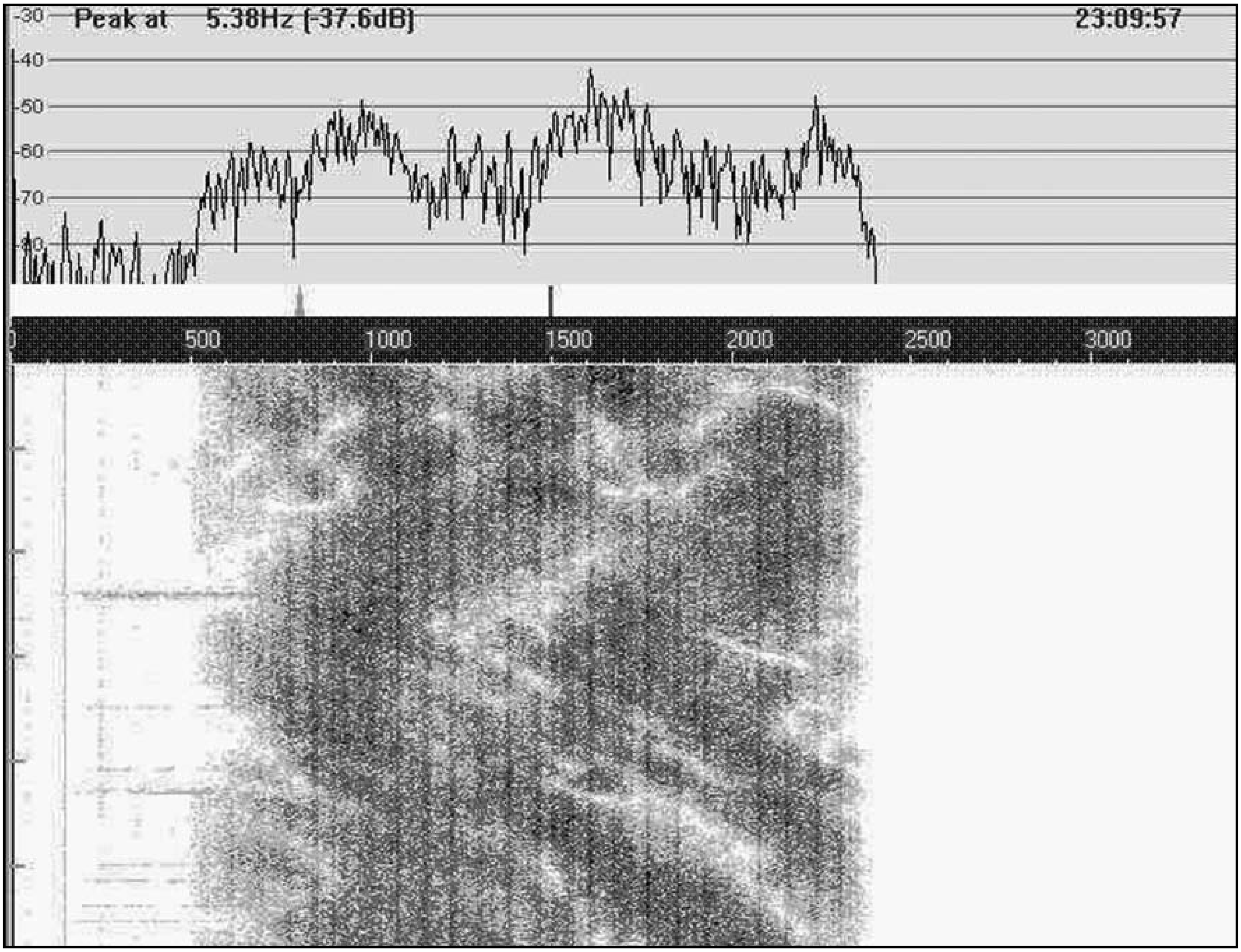

Figure 24 — Selective fading affecting a "stanag" signal on 3820 kHz.

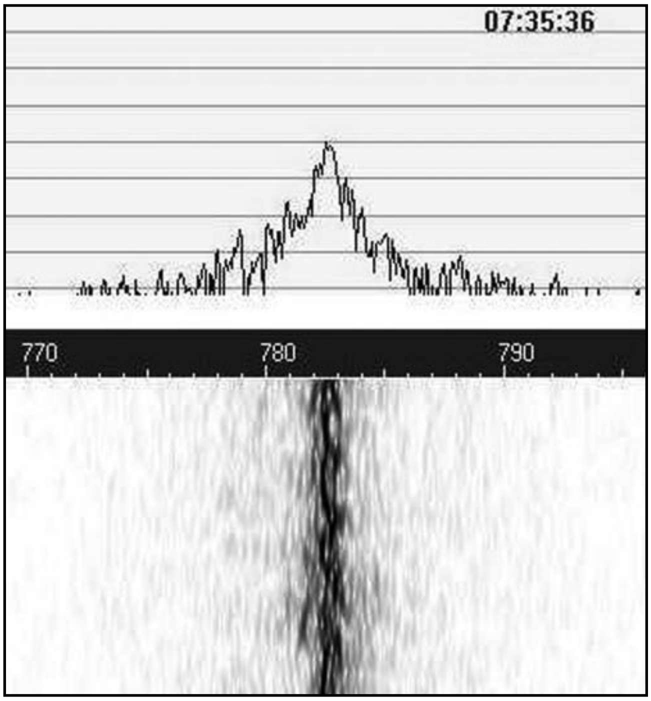

Figure 25 — Broadening of WWV signals due to Doppler spread.

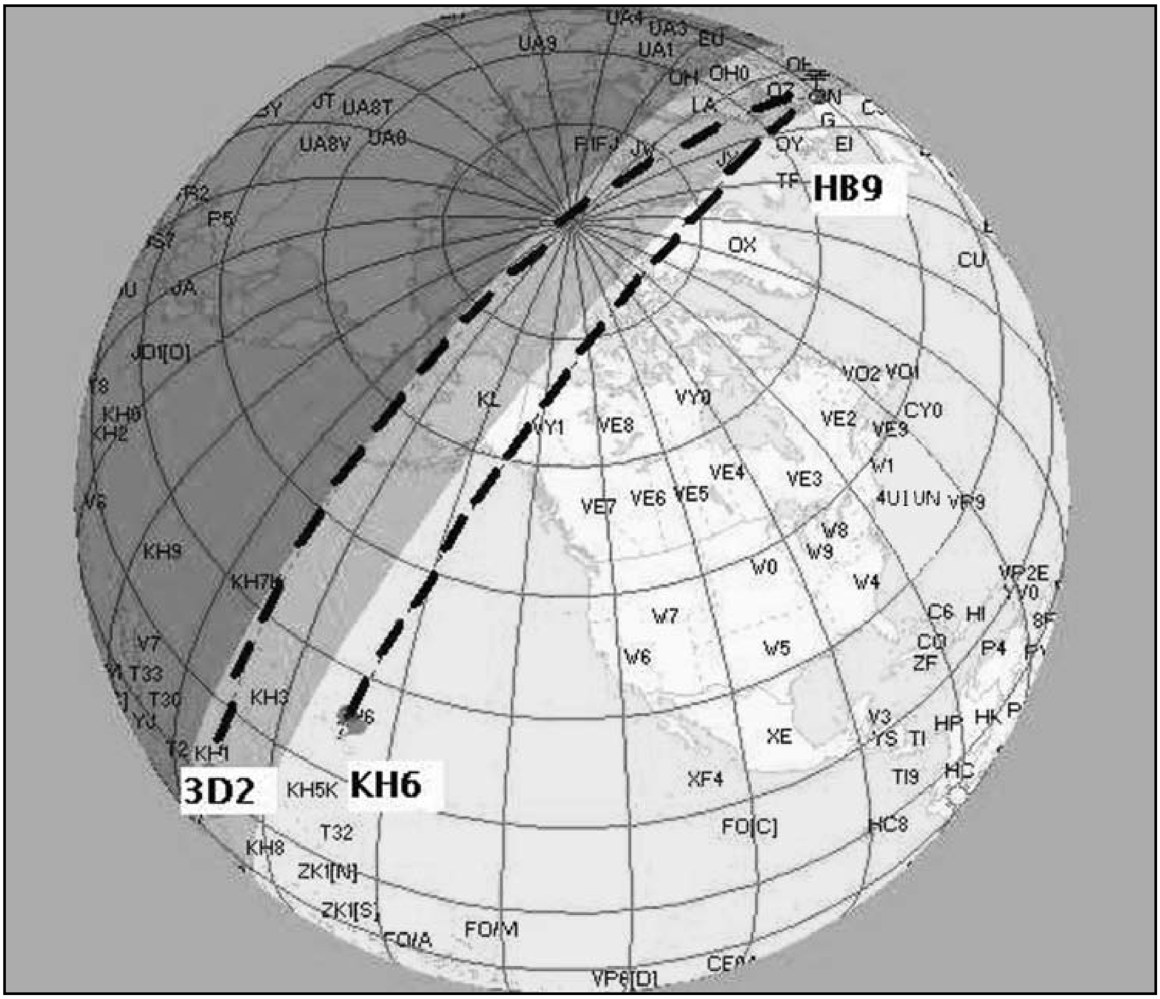

Figure 26 — Polar paths likely to be affected by Doppler spread (picture drawn with DXAtlas).

A Low-Cost Atomic Frequency Standard

A rubidium oscillator and a handful of parts result in a very accurate way to calibrate test equipment.

John S. Raydo, KØIZ

I recently added a used frequency counter and a signal generator to my complement of test equipment. Now, how do I check their calibration? The traditional way, of course, is zero-beating to WWV. This is adequate for ordinary crystal calibrators, but not nearly good enough for more precise equipment. I needed an accurate frequency standard. Doing an online search revealed that commercial frequency standards are based upon atomic technology. Cesium was best and is even used to control GPS satellites, but is *very* expensive. The next best choice is rubidium.

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Here's what I learned about rubidium oscillators: A microwave signal is derived from a voltage controlled crystal oscillator (VCXO). The amount of light from the internal rubidium discharge lamp drops by about 0.1% when the rubidium vapor in the resonance cell is exposed to microwave power near the transition frequency of 6.834,682,612 GHz (!). This light dip while sweeping an RF frequency synthesizer, locks the voltage-controlled oscillator. It sounded complicated to me but the result is accuracy some 1000 times that of zero-beating WWV. Shall we say adequate for my needs?

Commercial rubidium standards are still in the $2500 and up category, however. Too much for me; eBay to the rescue? Doing a "rubidium" search turned up several used rubidium oscillators at prices in the $200 to $300 range. These were self-contained 10 MHz oscillators in small metal boxes. I also saw some at the Dayton Hamvention flea market for $150. To make my own frequency standard I would only need to add a power supply and some incidental parts. I bid on and won an Efratom (Datum) model LPRO-101 oscillator. Cost was $207 including calibration and shipping (mine came from China and only took eight days).

The LPRO-101 is the most widely used rubidium oscillator and is frequently available on the used market. Other suitable rubidium

KØIZ

Figure 1 — The rubidium frequency standard output is displayed on a frequency counter. The display reads (1)0,000,000.00 Hz (the leading digit, 1, is not shown).

oscillators include the Efratom FRS series, Frequency Electronics FE-5650A/60A/80A series, Stanford Research PRS10, and others. Some may require supply voltages other than the 24 V dc that mine uses.

Building the Frequency Standard

These days a metal cabinet can be a fairly expensive part of a project. I further searched on eBay and found a nice box that contained a defunct power supply. I had no need for most of those parts but the box was an ideal 8 × 9 × 2½ inches. After some spray paint and rub-on lettering, I was in business, cheap. Figure 1 shows the completed project.

Almost any layout will do, depending upon the size of the cabinet or box you plan to use. The LPRO-101 oscillator requires a heat sink. I used an 8.5 × 7.25 inch piece of 0.25 inch thick aluminum plate, which cost $5 from a metal scrap yard. I mounted the plate on the inside of the box bottom. Next, I applied a very thin layer of thermal grease (using my finger) to both the bottom of the oscillator and top of the plate. Do not attempt to test or operate the oscillator without a heat sink, because this will produce excessive internal temperatures.

The LPRO-101 oscillator requires 24 V dc with startup current of 1.7 A, dropping to 0.5 A after a few minutes of warm-up.[1] A small switching power supply easily meets this requirement. The oscillator provides a "BITE" output (Built In Test Equipment) to indicate when unlocked from the rubidium standard. As shown in Figure 2, a small circuit board contains an inverting Schmitt Trigger to take this output and drive an LED to show the locked condition. A 5 V regulator and some additional power supply filtering are also on the circuit board. Figure 3 shows the layout inside the chassis.

The 10 MHz oscillator output is 0.5 V rms at 50 Ω. I wanted to use the unit as the reference oscillator for both my frequency counter and signal generator. A two-way 50 Ω splitter provides the two isolated outputs. You can skip the splitter if you only need one output.

The LPRO-101 and most other rubidium oscillators can be GPS disciplined for even higher accuracy. Compared with a crystal oscillator, the use of a rubidium oscillator allows extended sampling cycle duration. It also improves holdover performance should the GPS signal be lost. If you want this capability connect pin 7 of the oscillator to an external jack. The oscillator adjustment

[1]Notes appear on page 51.

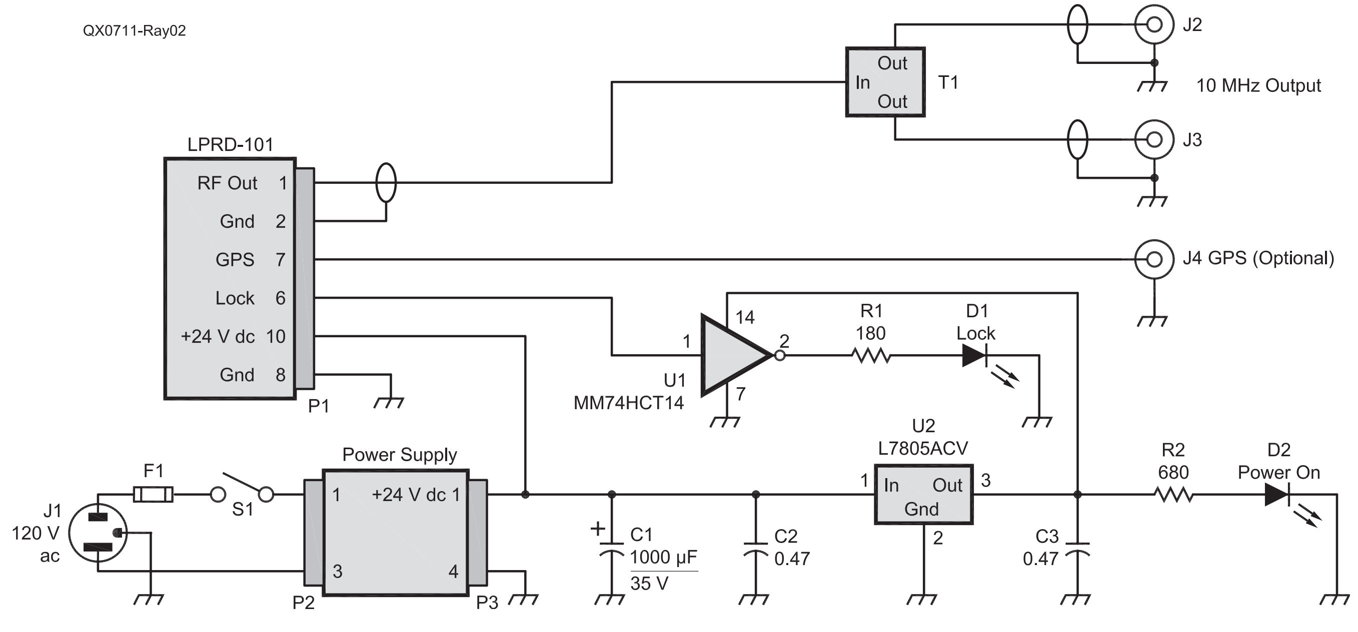

Figure 2 — This schematic diagram shows the circuit of the rubidium frequency standard.

C1	**1000 µF/35 V capacitor (Mouser 647-TVC1V1012MCD)**
C2, C3	**0.47 µF/63 V film capacitor (Mouser 581-BQ074D0474K)**
D1	**Green LED (Mouser 696-SSI-LXR3612GD)**
D2	**Red LED (Mouser 696-SSI-LXR3612ID)**
F1	**Fuse holder and 1 A fuse**
J1	**Chassis jack, ac (Mouser 161-0707-1-E)**
J2, J3	**BNC chassis connector (optional J3 if splitter is used)**
J4	**Optional connector for GPS circuit**
P1	**LPRO connector housing (Mouser 571-871332) and pins (6) (Mouser 571-5-87165-2)**
P2	**Power supply ac plug (Mouser 538-09-50-3031)**
P3	**Power supply dc plug (Mouser 538-09-50-3061)**

Connector pins (8) for above ac/dc power supply plugs (Mouser 538-08-50-0108)

PS	**PSA60-124-R 24 V switching power supply (Mouser 552-PSA-60-124-R)**
R1	**180 Ω ¼ W carbon film resistor**
R2	**680 Ω ¼ W carbon film resistor**
S1	**SPST toggle switch**
T1	**Mini-Circuits ZFSC-2-1 splitter or equivalent and BNC connectors (optional)**
U1	**MM74HCT14N Schmitt Trigger (Mouser 512-MM74HCT14N)**
U2	**L7805ACV 5 V regulator (Mouser 511-L7805ACV)**

Circuit Board (RadioShack 276-150)

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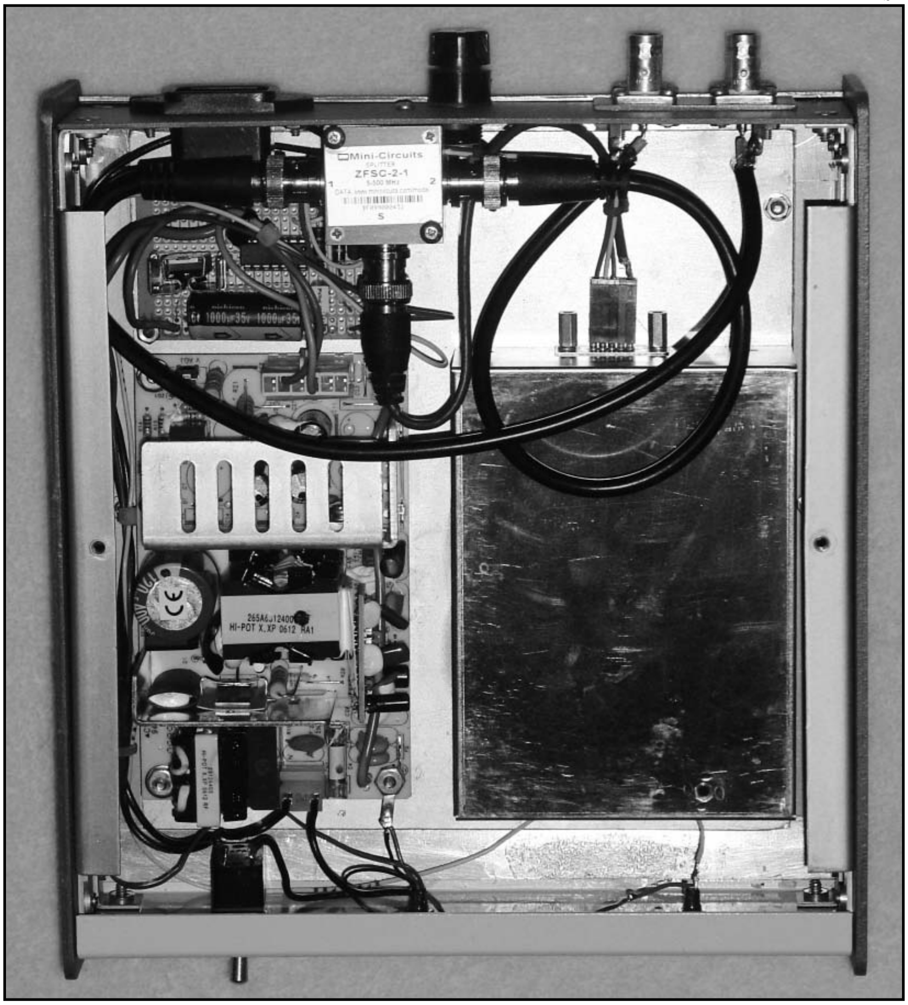

Figure 3 — Interior view of the rubidium standard. The LPRO-101 oscillator is the box to the right and the switching power supply is to the left. In the upper left is the circuit board for the Schmitt trigger. Fastened to the back panel is an optional Mini-Circuits two-way splitter.

range is 0 to +5 V, and is compatible with VE2ZAZ's GPS circuit.[2]

Good Results

Four minutes after power turn-on, the oscillator locks to the atomic frequency. The "Lock" LED then lights. Within ten minutes, accuracy is an impressive plus or minus 0.01 Hz at 10 MHz. The LPRO-101 spec is better than 5×10^{-11} per month (0.0005 Hz at 10 MHz) after a somewhat longer warm-up.

My equipment and that of others in our radio club have been calibrated and are now on frequency. This was a fairly quick project and my total cost was about $300. At this price an accurate frequency standard might also be a good addition to your station.

John Raydo, KØIZ, received his Novice license in 1957. From early on he has enjoyed designing and building ham radio equipment and antennas, and has written a number of articles for QST. *He is an active member of the Johnson County (Kansas) Radio Amateur Club. John is a graduate electrical engineer who also has a liberal arts degree in math and science, plus an MBA. He started his career working for TWA in the engineering department and later headed up their Information Services and Purchasing departments. He is now retired from his second career as an investment principal.*

Notes

[1]The LPRO-101 user manual is available at **www.symmetricom.com/media/pdf/manuals/man-lpro.pdf**

[2]Bertrand Zauhar, VE2ZAZ, "A Simplified GPS-Derived Frequency Standard," *QEX*, Sep/Oct 2006, pp 14-21.

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An ASCII Keyer

This keyer uses a paddle to send ASCII code instead of Morse.

Paddles and keyers have been long used to send Morse code. I have started experimenting with a non-iambic paddle sending ASCII code to a computer, as an alternative to using the keyboard. The paddle is a lot slower for me than the keyboard, but I enjoy using the paddle occasionally. Here, I describe an ASCII keyer with serial RS232 output, implemented in software on an old 8088 PC (personal computer).

As a demonstration, I log in to a Linux computer serial port using the keyer. I find the keyer romantic, but slow and error prone. It may not be the most *practical* way to communicate with a computer, but it *is* different.

In the usual operation with a Morse keyer, the paddle operator listens to the sidetone and synchronizes paddle motions with the keyer's timing to send the desired dots and dashes.

In this ASCII keyer, each byte is converted to standard asynchronous format, and output as RS232 serial data. All characters are eight-bit bytes, and there is an explicit word-space character. This means that the operator's timing no longer matters, and the operator does not have to synchronize paddle motions to the keyer:

- Each debounced closure of the dot contact is interpreted as a 0 bit.
- Each debounced closure of the dash contact is interpreted as a 1 bit.
- Every eight bits are interpreted as one byte, most significant bit first.

Send fast, send slow, even pause in the middle of a byte; the keyer correctly reads what is sent.

The 8088 PC runs MSDOS, which is excellent for simple programs that directly access the hardware, since DOS gives the running program complete control over the whole machine, including all the hardware ports. Any small microcontroller can do the same job. I happened to have the PC gathering dust. The PC has a keyboard, so I included the option to send characters using either paddle or keyboard at any time, so I can switch back and forth.

As shown in Figure 1, the 8088 PC parallel printer port is used to input the paddle signals:

- The common contact is connected to ground on pin 18.
- The dot contact is connected to /INIT on pin 16.
- The dash contact is connected to /AUTOFD on pin 14.

The intended functions of pins 14 and 16 are /AUTOFD and /INIT signals to a printer. The keyer program repurposes these pins as simple input bits with built-in pull-up resistors (so no power supply is needed for the contacts). This repurposing is trivial in an MSDOS machine because the running program has direct access to the hardware.

In the old PC, parallel-port pins 16 and 14 are each electrically an open collector output bit with passive pull-up resistors, plus an input bit to read back the state of the connector pin. The keyer program turns off the open-collector outputs, leaving the pull-up resistors to provide high input logic levels when the paddle contacts are open. Closed paddle contacts pull the inputs low.

Before starting the keyer program, the following MSDOS command initializes the serial port to 9600 baud, no parity bit, eight data bits, and one stop bit:

MODE COM1: 9600,N,8,1

The keyer program initializes the printer port, and then enters an endless loop that continually polls the contact status, filters

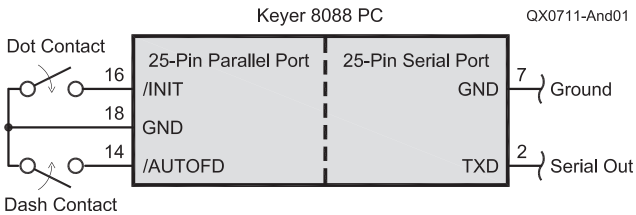

Figure 1 — This diagram shows the wiring of the paddle to the serial input and output of the 8088 PC keyer.

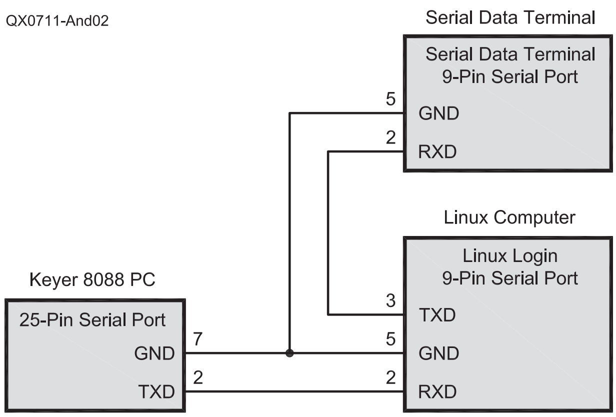

Figure 2 — A three-ended RS232 cable connects the 8088 PC keyer and serial data terminal to the Linux computer login serial port.

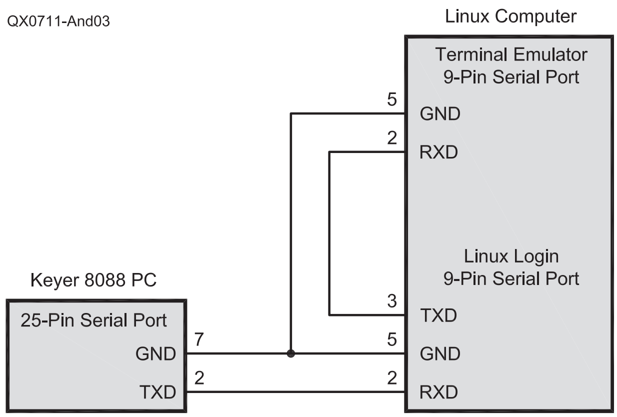

Figure 3 — A three-ended RS232 cable connects the 8088 PC keyer and terminal emulator to the Linux computer login serial port.

the contact data, and accumulates the input bits into bytes.[1] Whenever a byte is complete (or a byte is typed on the PC keyboard), the program writes the byte to the serial port.

The contact data is put through a low-pass filter to debounce the contacts. The filter is a single-pole low-pass filter implemented as follows:

new_filter_output = old_filter_output + f*(new_data - old_filter_output)

Where fraction f is between 0 and 1, and sets the time constant of the filter:

f = time_per_pass_through_loop / time_constant

I used $f = 1/12$ to give a time constant of about 20 ms. Note that the time constant depends upon the speed of the computer, so f will have to be changed for a computer with a speed that differs substantially from the 8 MHz of my 8088. The numbers used in the filter are scaled in the keyer program, so that integer math is used for the filter.

This ASCII keyer outputs RS232 serial data. To test the keyer, I used some ancient features of a Linux desktop computer.

Modern Linux desktop computers have more-or-less-familiar graphical user interfaces (GUI's), and most users are familiar with these interfaces. Linux desktop computers also have ancient interfaces showing their multitasking, multiuser heritage. In particular, more than one user can be logged in at one time through various interfaces. One user can be logged in through the GUI while other users are logged in through the network, and another user is logged in through a serial port.

If one logs in to the serial port using an ASCII serial data terminal, one can run any text-mode program. This includes the command-line shell (command-line interpreter), text editors, compilers and even text-mode Web browsers.

The serial login feature is normally disabled, but is easily enabled. Search your documentation and the Web for "Linux Serial Console" for details.

It's this option that I used to test the keyer, as shown in Figure 2. I used a three-ended RS232 cable. The RS232 data from the keyer goes to the serial port on the Linux computer. The RS232 data from the Linux computer goes to an RS232 serial data terminal, to be displayed.

I place the paddle next to the terminal's display, and I have a paddle-input text-mode computer to experiment with.

[1]The source code for this simple program is available for downloading from the *QEX* Web site. Go to **www.arrl.org/qexfiles** and look for **11x07_Anderson.zip**.

An added twist to the above is this: I also use the same Linux computer as the RS232 serial terminal!

Recall that Linux computers are true multitasking, multiuser computers, and have always been able to do multiple things for multiple users at once. Linux computers also come with lots of ancient software hidden behind the modern GUI. Among that ancient software, are some serial terminal emulators, left over from when people phoned into bulletin-board systems (BBS) using modems.

My Linux computer has two serial ports. I connect the three-ended serial cable as shown in Figure 3, with the keyer data going to the lower serial port, and data from the lower serial port going to the upper serial port.

I log in to the GUI on the Linux computer, and start a terminal emulator program running, using the upper serial port. This has no effect on what might be happening with the lower serial port.

I can now log in to the Linux computer again through the lower serial port using the keyer for input, and the output shows up on the terminal emulator's display window in the GUI.

I am now logged into the Linux computer twice, and every character I send using the paddle goes through the Linux computer twice to reach the screen. It's like old reflex radio receivers where the signal went through one tube twice, first as RF and later as audio.

So, how is it to use the paddle versus the keyboard? I found the paddle very romantic, though not as romantic as Morse on a straight key.

I enjoyed not having to worry about timing of keying, though I did occasionally send too few or too many bits. The paddle was much slower than the keyboard, but was usable for Web browsing, where a little sending caused the reception of a lot of text to read.

I found myself to be sufficiently error-prone that I had to send passwords (which do not print on the terminal's display) using the 8088 PC keyboard.

In the future, I'd like to implement a USB version of the keyer, as USB is more compatible with modern computers and operating systems, and also would permit GUI operation with one hand on the trackball and the other hand on the paddle!

In summary, the described paddle-driven ASCII keyer with RS232 serial output is simple, feasible, and slow. Old features of a modern Linux computer provided a simple method of demonstrating and testing the ASCII keyer.

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Antenna Options

Horizontally Polarized Omnidirectional Antennas: Some Compact Choices

Obtaining omnidirectional coverage with vertical polarization is simple: use a version of the many vertical antennas including the λ/4 ground-plane monopole, the vertical dipole with or without a J-pole matching section, or any number of collinear variations on these antennas. However, if we wish to have omnidirectional coverage with horizontal polarization, solutions are less automatic. In fact, the search for a perfect horizontally polarized omnidirectional (HPOD) antenna goes back into the dim recesses of antenna history. We shall examine a number of options, their limitations, and, in some cases, ways to overcome those limitations. We shall divide the work into two parts, looking at some of the more compact choices in this episode. Next time, we shall examine a few larger omnidirectional horizontal arrays and take a longer look at stacking them.

The search has two dimensions. The first is obtaining a perfect circle for an azimuth radiation pattern. How far from circular you may be willing to accept a pattern may determine how much work you will put into the antenna design and construction — or vice versa. The second dimension is field strength, meaning the antenna gain at low angles. As we shall see, some designs with good patterns unfortunately send a goodly part of their energy in useless directions, such as straight up or down. While helpful for satellite reception, these antennas are less than ideal for some VHF point-to-point applications.

All of the antennas in these notes use 144.5 MHz as the test frequency. Patterns and performance values assume a height of 20 feet above average ground.

Turnstile or 90° Phase Fed Systems

Perhaps the oldest HPOD antennas employ one or another form of phase feeding, using at least two elements. Figure 1 shows the most common form of feeding one element with the same current magnitude, but phase-shifted 90° from the other. Both elements are identical, but are at right angles to each other. The phase line characteristic impedance is the feed-point impedance of the directly fed element. However, with both elements connected, the net feed-point impedance is one-half of the

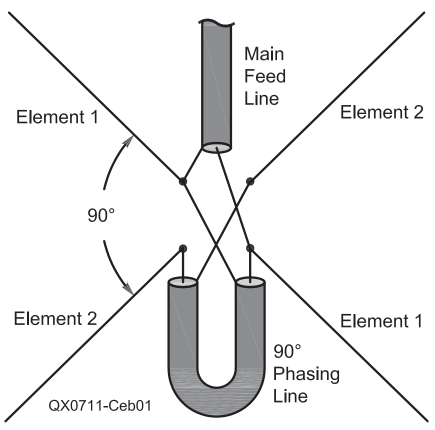

Figure 1 — The basic turnstile feed system.

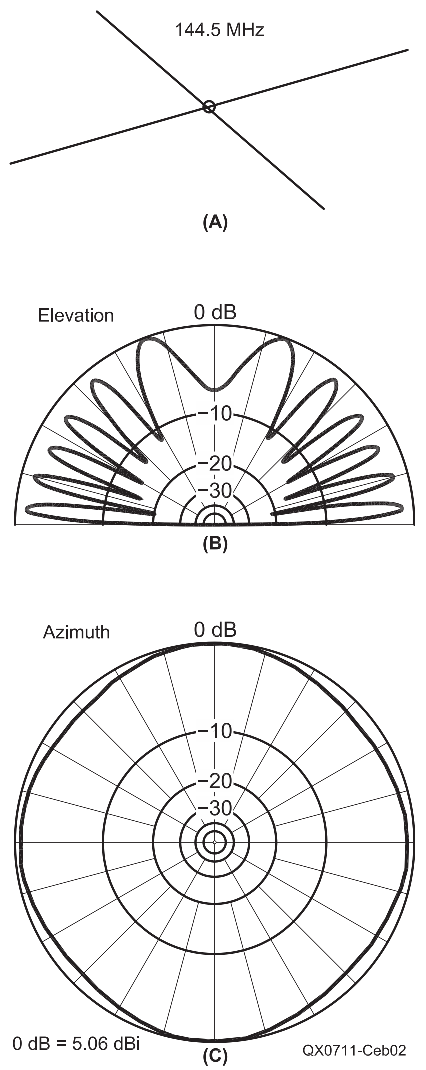

Figure 2 — Turnstiled dipole elements: General outline and patterns at 20 feet above average ground.

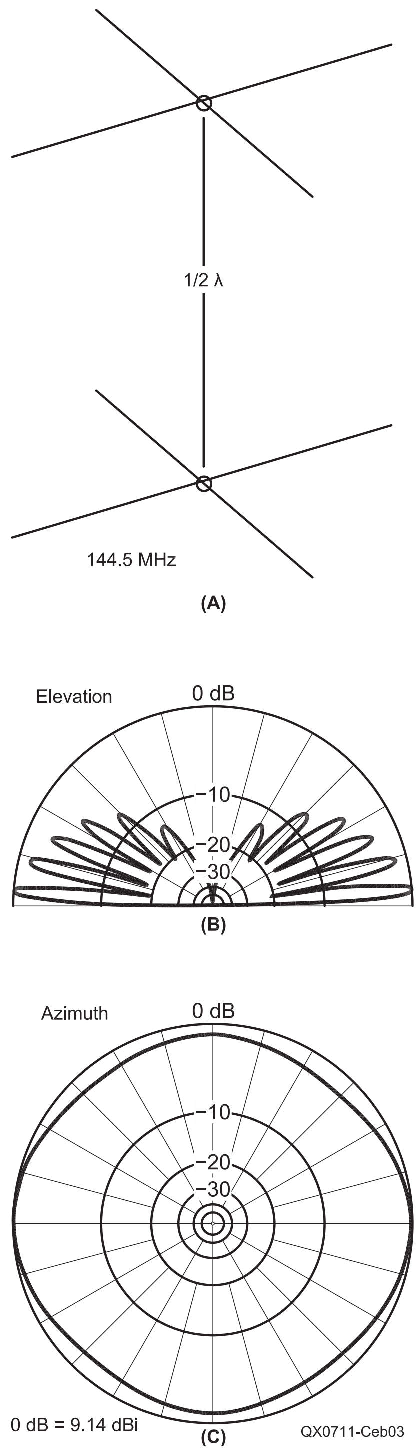

Figure 3 — A dipole turnstile 2-stack: General outline and patterns with the lower antenna 20 feet above average ground.

impedance of an isolated element. There are alternative feed systems to arrive at the same goal, but a few of them concern themselves with impedance matching rather than obtaining the correct current magnitude and phase angle at the center of each element.

The simplest version of this antenna also gave birth to the generic names for the feed system. A pair of resonant dipoles at right angles to each other presents the appearance of a turnstile. Figure 2 shows the general outline of the antenna. The version from which we drew the patterns uses 0.125-inch aluminum for the 38.96-inch elements. The model places the elements 0.25 inches apart, center-to-center. Each dipole presents a 70-Ω impedance. With the 70-Ω λ/4 phase line in place, *NEC-4* reports a net feed-point impedance of 35 Ω. Since the impedance does not change over a very broad frequency range, we may accept the 1.43:1 50-Ω SWR, or we might use a series matching system to match more exactly a 50-Ω feed line.

The azimuth patterns show a slight squaring, but the gain range is only about 0.5 dB, less than we could detect in operation. The chief limitation of the turnstiled dipoles is revealed in the elevation pattern. We find more energy broadside to the dipole pair than off its edges. Hence, the maximum gain at 20 feet and a 4.8° TO (take-off) angle is only 5.06 dBi.

Ott Fiebel, W4WSR, pointed out to me an interesting variation on the standard turnstile by placing two λ/2 elements at a 90° angle, but using one end of each element to form the apex. A simple λ/4 parallel line matching section connected in series with the ends at the apex allows a 50-Ω match. Although the pattern is not quite as clean as the standard turnstile pattern, the construction is simple and reliable.

Figure 3 shows one way to overcome the broadside radiation of turnstiled dipoles: create a stack with λ/2 separation. In the model, the lower antenna is at 20 feet, with the upper antenna about 80 inches above it. The elevation pattern shows a radical reduction in high-angle radiation. The maximum gain of the antenna pair is 9.14 dBi, with a

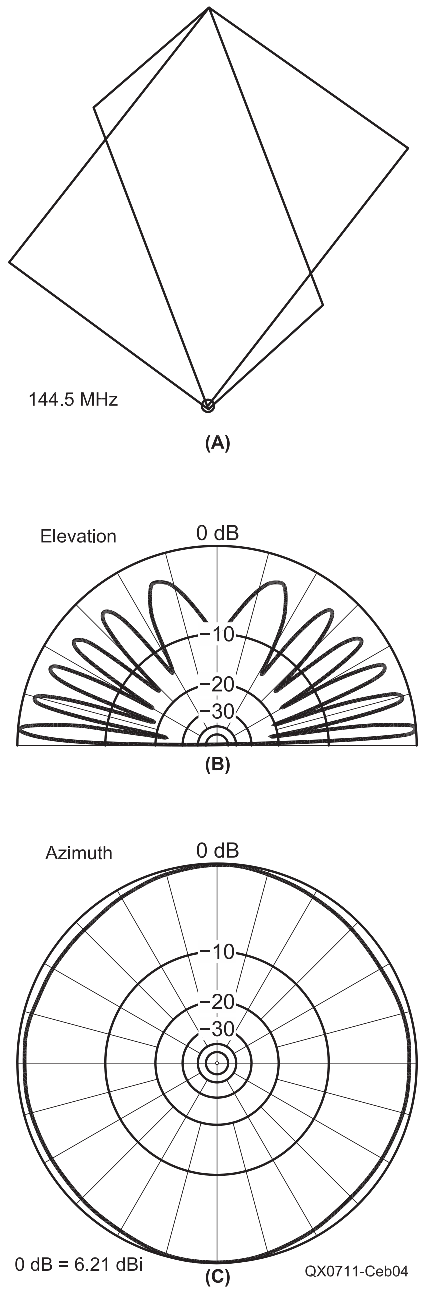

Figure 4 — Turnstiled quad loop elements: General outline and patterns at 20 feet above average ground.

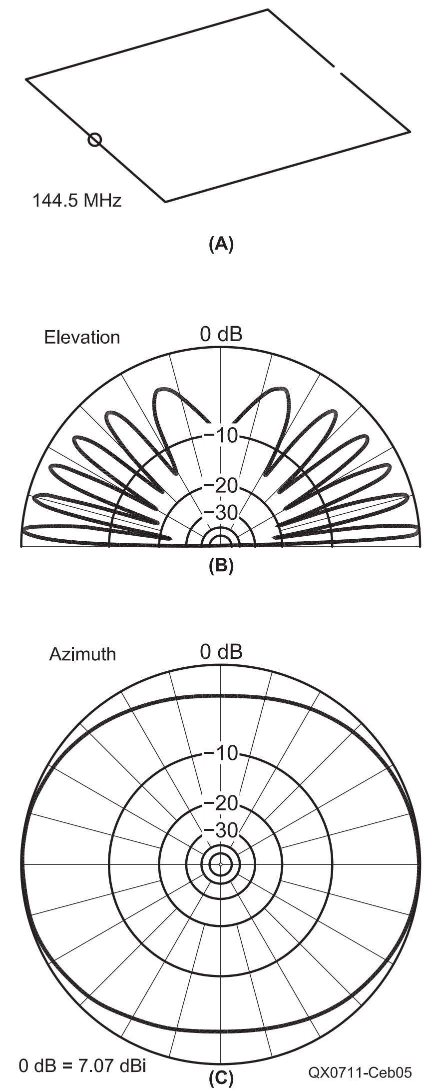

Figure 5 — Single-element halo (interrupted half-wavelength loop): General outline and patterns at 20 feet above average ground.

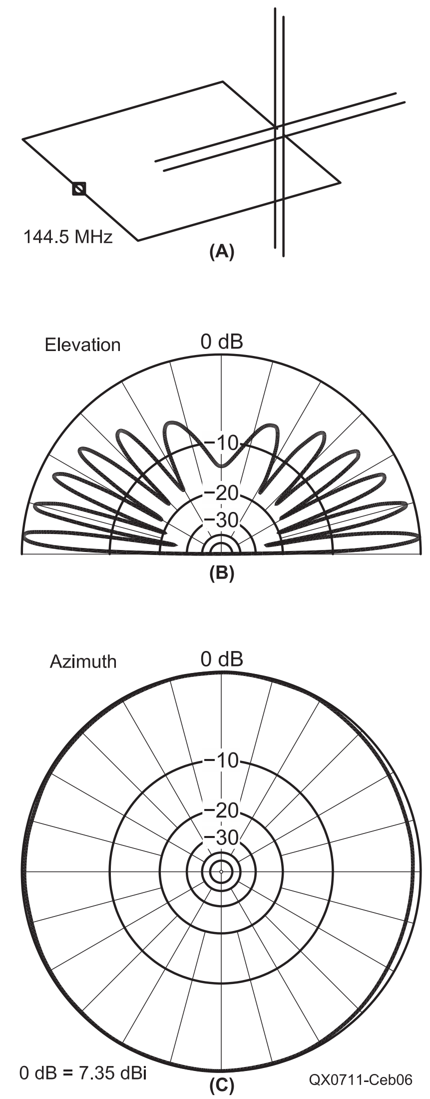

Figure 6 — Single element halo with capacitive compensation: General outline and patterns at 20 feet above average ground.

0.9-dB range of gain around the perimeter of the azimuth pattern. However, we cannot use the same dipoles that we used in the single turnstile. Mutual coupling between the bays requires that we lengthen the dipoles to 40.2 inches. Without this adjustment, the pattern becomes very distorted. We would not notice the distortion from the SWR alone.

For all turnstiles, the SWR bandwidth is very much wider than the operating bandwidth measured in terms of an acceptable pattern. For further information on the performance behavior of turnstiled antennas, see "Some Notes on Turnstile Antenna Properties," *QEX*, Mar/Apr, 2002, pp 35-36.

We may also turnstile 1-λ quad loops at right angles to each other. A single quad loop has an impedance of about 125 Ω, for a net feed-point impedance of 62.5 Ω for the turnstiled pair. RG-63 (125-Ω) coax is suitable as a phase line. Figure 4 shows the outline and patterns for one version of the antenna using a diamond configuration for simplified construction. (See "A 6-Meter Quad Turnstile," *QST*, May 2002, pp 42-46, for one version of this antenna.) The elements are no. 12 copper wire, with each loop having a circumference of 87.7 inches. Alternatively, we could use quad loops in a square configuration, the so-called *eggbeater*. In either configuration, we could leave a gap between the top wires at the crossing point or connect them together. Performance does not change.

The elevation pattern shows a significant improvement in the direction of radiation from the antenna, with the lowest lobe as the strongest. The maximum gain of the modeled turnstiled quad is 6.21 dBi, with a gain range of about 0.5 dB, as shown by the not-quite-perfect circle of the azimuth pattern. The TO angle is 4.8°. All of the patterns in these notes will use the maximum gain at the TO angle as a measure of performance. It serves as a stand-in for our real concern with HPOD antennas: the signal strength for point-to-point communications over some fixed distance and a fixed observation or reception height.

Once we overcome the basic turnstile's broadside radiation that robs energy from the desired edgewise signal path, matching becomes the most obvious construction hurdle. However, the turnstile antenna has a hidden limitation. The pattern shape is highly dependent upon the magnitude and phase relationship between the two feed points. Most common feed systems provide a correct relationship at only one frequency, and the values change as we move away from that frequency. Small inequalities in the current magnitude and departures from the required 90° phase-angle result in considerable distortion to the nearly circular pattern at the design frequency.

We might easily turnstile or phase-feed a number of other antennas. However, most of them would serve better for satellite communications than for horizontally polarized direct communications. Therefore, we may let the dipole and quad loop turnstile pairs serve as examples of our initial technique for obtaining omnidirectional patterns.

Halos

The second common category of antenna for obtaining at least a semblance of a circular azimuth pattern is one or another version of the *halo*. More correctly, this class of antenna rests on a λ/2 dipole bent so that the ends almost meet. The most tempting form for the halo is either a circle or a square. However, as Figure 5 shows, a symmetrical halo with a single element yields a highly non-circular pattern. The model uses 0.5-inch aluminum for the element that is 10 inches on a side with a 0.52-inch gap. The azimuth pattern gain varies by nearly 3 dB. To boot, the impedance is only about 8.3 Ω.

One way to circularize the pattern is to add a surface on each element end to increase the capacitance between ends. The model outlined in Figure 6 simulates plates by using four facing 6-inch wires on each side of the gap. A normal halo would employ a disk. We see an obvious improvement in the pattern shapes, with a maximum gain of 7.35 dBi and a gain variation of less than 0.4 dB. As well, the feed-point resistance becomes about 49 Ω. However, the plates have added an inductive reactance of *j* 1000 Ω. Hence, we need a series capacitance at the feed point of 1.1 pF. This requirement creates a considerable matching difficulty, since very small changes in the capacitance will create large changes in the feed-point impedance.

To overcome this difficulty — at least to some extent — we may use a double loop, as shown in Figure 7. The double loop is essentially a halo version of a folded dipole,

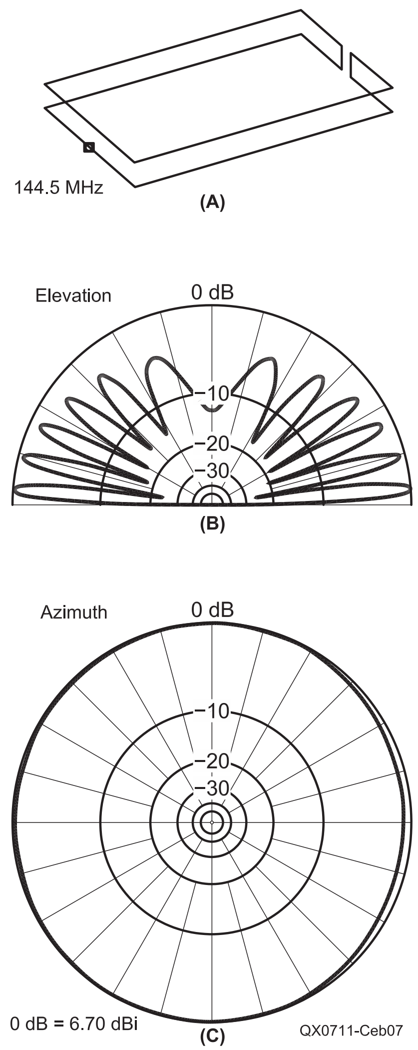

Figure 7 — Double rectangle: General outline and patterns at 20 feet above average ground.

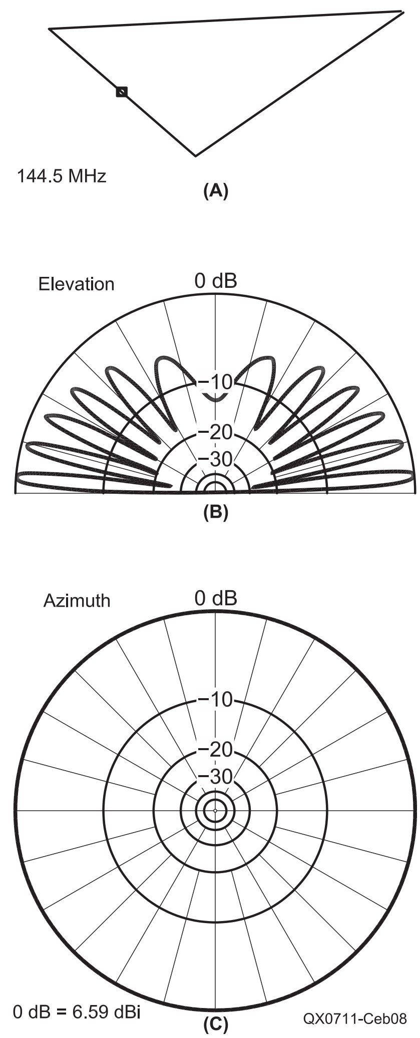

Figure 8 — Small triangle: General outline and patterns at 20 feet above average ground.

which raises the feed-point impedance. At the same time, the design uses a rectangular form to circularize the azimuth pattern without the need for a large capacitive structure. The sample model uses no. 12 copper wire. Across the feed point, the side is 8.8 inches, while the long sides are 14.3 inches. The gap between ends is 0.35 inches and the spacing between wires is 1.34 inches.

The double rectangle shows a maximum gain of 6.7 dBi (at 20 feet above average ground), with a variation of only about 0.5 dB around the azimuth pattern. The feed-point resistance is 64 Ω, but there remains a considerable inductive reactance. To compensate for the 721-Ω reactance, we require a series capacitor at the feed point (1.53 pF), which complicates effective and efficient matching. Alternatively, we may adjust the capacitance across the gap by reshaping the wires that face each other.

The halo need not be either circular or rectangular to form an interrupted half-wavelength loop. One interesting alternative shape is a triangle with a gap at the apex, across from the feed point. With the correct ratio of leg-length to feed-point side leg and the correct gap, we can obtain a very circular azimuth pattern. In general, we find two versions of the triangle: A smaller version with a circumference that is less than 0.6 λ and a larger version with a circumference greater than 0.75 λ.

Figure 8 shows the outline of a small triangle that uses a 0.125-inch diameter aluminum element. The feed-point side is about 14 inches long, while the angled legs are each 16.8 inches. The circumference is 47.5 inches. The apex gap is quite small in the model: 0.12 inches. At 144.5 MHz, the maximum gain is 6.59 dBi, with less than 0.1-dB variation in the gain. However, the resonant feed-point impedance is only 8.4 Ω. Hence, we require an impedance transformer to use the antenna effectively.

The large triangle is an alternative to the smaller one. The modeled sample in Figure 9 has a circumference of 64.3 inches, with a 23.8 inch feed-point side and 20.3-inch angled legs. The large triangle achieves about 6.37-dBi gain at the test height. Like the small triangle, the gain variation around the azimuth pattern is under 0.1 dB. The ostensible advantage of the large triangle is the feed-point resistance: 58 Ω. However, the feed-point impedance also shows a remnant inductive reactance of 525 Ω. Hence, we require once more a series capacitance (1.85 pF) or other treatment to arrive at resonance.

In general, accurately constructed halos exhibit very circular azimuth patterns. Unlike the turnstiles, the pattern shape is quite stable over frequency spreads within the SWR bandwidth of the antenna. Nevertheless, halos have their own limitations for the home builder. Prototypes that I have built of various halos — both rectangles and triangles — in my modest shop suggest that the antenna type presents us with two significant hurdles. The elimination of remnant reactance is the more obvious of the challenges. The second difficulty lies in the susceptibility of these antennas to changing resonant frequency with only minor flexing of the elements. The interrupted-loop construction adds the size and alignment of the gap to the list of dimensional concerns. An effective halo must freeze both the size of the gap and the alignment of the element at the gap. As well, the remaining lengths of element material should not be susceptible to flexing that would change the design shape.

The Uniform or Constant Current Loop

An overlooked design emerged in 1944 in Donald Foster's "Loop Antennas with Uniform Current," *IRE*, Oct 1944. Recently, Robert Zimmerman, NP4B, resurrected the idea in "Uniform Current Dipoles and Loops," in *antenneX* for Apr and May 2006. The principle is to divide the circumference of a loop into sections such that the inductance of each wire length is offset by a periodic capacitor and so that the loop exhibits a 50-Ω impedance — without need for any form of matching. Let's divide a square of wire into seven sections. Each section will be 0.12 λ long, for a total circumference of 0.84 λ. At each wire junction, we shall insert a capacitor. The capacitor size will vary with the wire diameter. Number 12 wire calls for 4.11-pF capacitors.

In physical terms for 144.5 MHz, each no. 12 wire section is 9.8 inches long. The square is 17.15 inches on a side, for a circumference of 68.6 inches. The number of

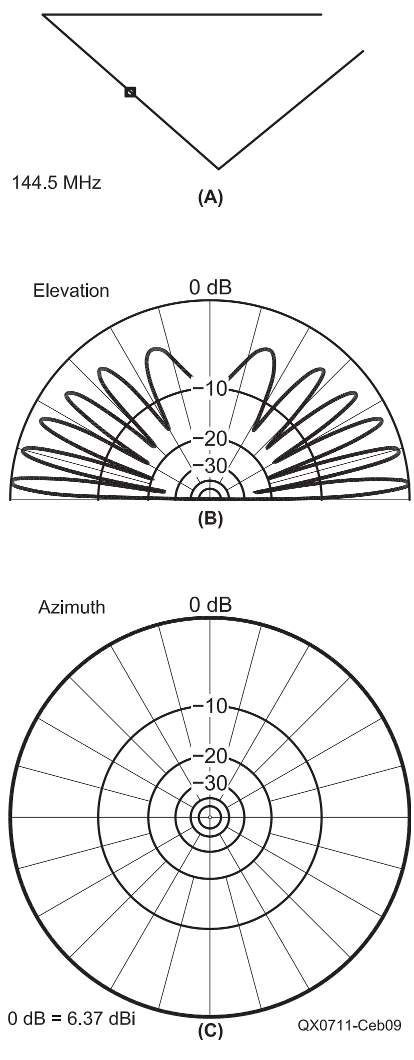

Figure 9 — Large triangle: General outline and patterns at 20 feet above average ground.

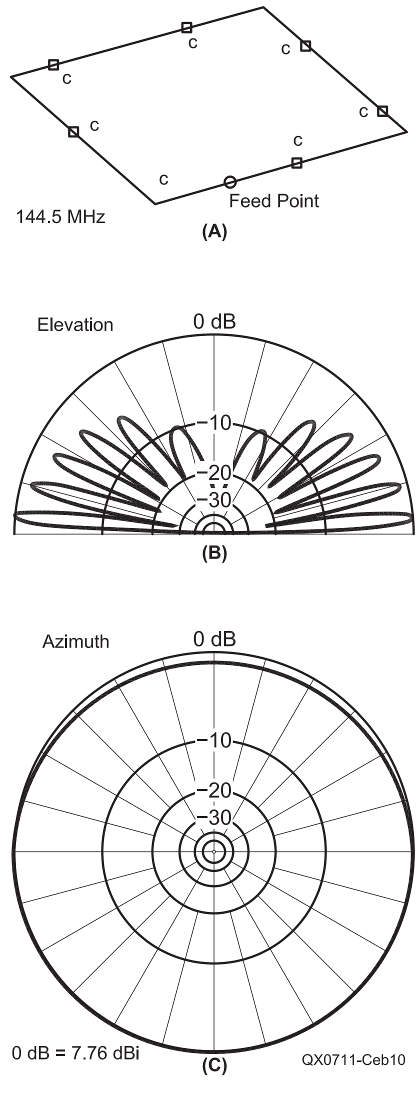

Figure 10 — Constant-current loop: General outline and patterns at 20 feet above average ground.

sections (seven) does not correspond to the number of sides (four), which is no hindrance to effective antenna operation. Although the component arrangement yields omnidirectional patterns, the appearance of the antenna, as shown in Figure 10, may seem initially strange.

It does not matter if the feed point is placed mid-side or near a corner, so long as the feed point is in the middle of a wire section. The relative current magnitude along the circumference of the loop changes by less than 4% all along the perimeter. (Initially, this phenomenon appears to have been the goal of the open-ended CCD long doublet, but the open ends preclude obtaining that result.)

The uniform-current square loop provides horizontally polarized radiation. Although only a little larger than the triangles, the results are equal in omnidirectional pattern and superior in gain. At 20 feet above average ground (close to 3 λ), the maximum antenna gain is 7.76 dBi, with a total variation in gain of about 0.8 dB. The gain is about a dB better than the best triangle. The elevation pattern reveals one significant reason for the improved gain from the loop. If you compare the elevation pattern with the ones shown for the triangles, you will see that the loop produces virtually no radiation straight upward, leaving more energy for the lower lobes.

Since the antenna does not need to compensate for rapidly changing reactance values, it shows a reasonable SWR bandwidth. However, the design is sensitive to the capacitor value within very close tolerances. The resonant impedance (50.7 Ω) of the model using 4.11 pF capacitors changed to an impedance of 45.4 – *j* 39.1 Ω simply by using a 4.0 pF capacitor value. However, Zimmerman uses an interesting technique involving parallel transmission line for his loops. See "Uniform Current Loop Radiators," *QEX*, May/Jun, 2006, pp 45-48. By cutting alternative positions on the wire length, he allows the facing wires to form the capacitors. Field adjustment consists of slowly widening the gaps until you achieve the desired capacitance.

If you prefer a more symmetrical arrangement, you might increase the number of capacitors to eight, placing them at corners and at the center of each side. Without altering the loop size, the capacitor size increases slowly as you add capacitors. For eight capacitors, models suggest a value of 4.7 pF for each one. The feed point remains centered between two capacitors. In addition, the radiation performance does not change. The chief hurdle in constructing a constant-current loop is still obtaining the correct capacitance value, a matter for careful construction.

Conclusion

We have looked as some of the basic options for horizontally polarized omnidirectional antennas, including turnstiled elements, halos of various shapes and the constant-current loop.[1] Our concern has been less to look at specific construction ideas than to see the basic principles, as well as the limitations and challenges, presented by each class of antenna. Which one you may decide to build will likely rest as much on local shop skills as on basic needs for the antenna. Commercial versions exist for some of the antennas discussed, with halos especially popular. For the inveterate antenna builder, there are many additional options. Next time, we shall examine a few larger arrays and the stacking question.

[1]Models for the antennas discussed in these notes are available at the ARRLWeb site. Go to **www.arrl.org/qexfiles** and look for 11x07_AO.zip. All models are in *EZNEC* format.

Out of the Box

Raymond Mack, W5IFS
17060 Conway Springs Ct
Austin, TX 78717-2989

NEW FILTER DESIGN SOFTWARE

Applied Computational Sciences has developed a new program: a linear circuit simulator, including design and synthesis tools suitable for HF through microwaves. The program includes a large library of standard parts for analysis of FET and bipolar circuits, along with S parameter models for active and passive components, including transmission lines. The design and synthesis tools allow ideal transmission line elements to be converted into real implementations on circuit boards.

There are two versions of the software: *LINC2* and *Micro*LINC, which is the Professional, full featured version. *Micro*LINC is also referred to as *LINC2 Pro* on the ACS Web site. The software allows synthesis of low noise amplifiers (LNAs), multistage amplifiers, filters, impedance matching circuits, attenuators, transmission lines and more.

The price for *LINC2* is $175 and *LINC2 Pro* is $495. As of this writing in early September 2007, the price of $495 for *LINC2* Pro includes *LINC2 Filter Pro* and *Amp Pro* software packages. That is listed as a savings of $1590. For more information, contact:

Applied Computational Sciences
1061 Dragt Pl
Escondido, CA 92029
Phone: 760-612-6988
e-mail: ***LINC2*@appliedmicrowave.com**
Web: **www.appliedmicrowave.com**

NEW SOURCE FOR ONE PULSE PER SECOND GPS

We have presented several articles and product announcements regarding using the one pulse per second output of a GPS receiver to train a VCXO to become an accurate frequency standard. The first was an article that used the Hewlett Packard Z3801A modules, which were briefly available as surplus pulls from cell phone sites.[1] Another article, uses any modern GPS receiver with a 1 pulse-per-second output signal.[2]

There are sources of brand new receivers, but they have not been low priced. SiRF is a company that makes a very nice line of GPS receiver chips. They haven't been available in small quantities until now. Tyco has designed a line of GPS receivers for the rest of us. They are priced in the range of $29 to $60. They also have evaluation modules that begin as low as $130 and go to $300 each. The evaluation modules include the material and software to make a complete functioning GPS receiver. These modules and evaluation kits are available and in stock at Mouser Electronics.

Tyco Electronics
www.tycoelectronics.com/gps/
Mouser Electronics, Inc
1000 North Main St
Mansfield, TX 76063
Phone: 800-346-6873
Local: 817-804-3888
Fax: 817-804-3899
Web: **www.mouser.com/catalog/631/29.pdf**

HITTITE ANNOUNCES 75 DB LOG DETECTOR

Hittite has announced a new HMC600LP4 SiGe logarithmic detector with 75 dB dynamic range that covers the frequency range 50 MHz to 4 GHz. The part can be used in two modes. If one ties the Vset pin to Vout the part operates as a logarithmic detector with 19 mV/dB output. If one ties the Vset pin to an external voltage, the part can operate as an AGC circuit. Hittite claims a 75 dB useful range, but the linear range (< 1 dB error) is only about 65 dB (–62 dBm to +4 dBm) below 900 MHz. For frequencies from 1900 MHz to 4000MHz, the linearity is reduced and the linear dynamic range decreases with frequency to –5 dBm to –45 dBm.

Parts can be ordered directly from the Hittite Web site with a minimum order of 10 parts. The HMC564LC4 is $9.96 each. An evaluation board is available for $392.00.

Hittite Microwave Corporation
20 Alpha Rd
Chelmsford, MA 01824
Phone: 978-250-3343
Fax: 978-250-3373
Web: **www.hittite.com**

HIGH VOLTAGE SURFACE MOUNT RESISTORS

Allowable working voltage is a frequently overlooked resistor parameter. The working voltage for "normal" surface mount resistors varies from 25 V to 150 V depending on construction and power level. Even radial lead through hole parts are only rated to 200 V at ½ W (again depending on construction). HDK America has introduced a line of high voltage surface mount resistors suitable for high voltage applications. The HCR50 series resistors handle 1500 V with ½ W dissipation. Packaged in a 2010 case, they are available from 3 Ω to 16 MΩ. The temperature coefficient is ±200 ppm/ºC and operate from –55 to 125°C.

KOA Speer has also added high voltage surface mount resistors. The HV73 series is available in 350, 400, and 500 V ratings. They are available in values from 10 kΩ to 51 MΩ and tolerances of 0.5, 1, 2, and 5%.

HDK America
Barrington, IL
Stuart Tanaka
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KOA Speer Electronics
Bradford, PA
Dawn McGriff
Phone: 814-362-5536, ext 266
e-mail: **dmcgriff@koaspeer.com**

NEW SOURCES FOR UHF AND MICROWAVE SEMICONDUCTORS

This is a double-purpose announcement. The first is that the group that was originally the Hewlett-Packard semiconductor group, then Agilent semiconductors after the spin-off, is now known as Avago Technologies. The other news is that Mouser Electronics is now a distributor of the Avago line of RF parts, including silicon and GaAs low noise transistors and ICs. Many of the microwave parts you have seen in *QEX* articles are available from stock. As a bonus, Mouser provides direct links on their Web page to Avago design information on each part. Clicking on the datasheet link takes you to an Avago Web page, with datasheet and application information.

If you cannot find the parts you need at Mouser, you can try Digi-Key. They also stock Avago parts. You will find links on their Web page to Avago data sheets, as well.

Avago Technologies
350 W Trimble Rd
San Jose, CA 95131
Phone: 800-235-0312
Web: **www.avagotech.com**

Mouser Electronics, Inc
1000 North Main St
Mansfield, TX 76063
Phone: 800 346-6873
Local: 817-804-3888
Fax: 817-804-3899
Web: **www.mouser.com**

Digi-Key
701 Brooks Avenue South
Thief River Falls, MN 56701
Phone: 800-344-4539
Local: 218-681-6674
Fax: 218-681-3380
Web: **www.digikey.com**

[1]Bill Jones, K8CU, "Using the HP Z3801A GPS Frequency Standard," *QEX*, Nov/Dec 2002, pp 49-53.

[2]Bertrand Zauhar, VE2ZAZ, "A Simplified GPS-Derived Frequency Standard," *QEX*, Sep/Oct 2006, pp 14-21.

Letters to the Editor

Calibration of a Homebrew Noise Source (*QST* May, 1994)

My article, "A Calibrated Noise Source for Amateur Radio," in the May 1994 issue of *QST* described the theory, design and construction of two noise sources for the frequency ranges 0.5 to 500 MHz and 1.0 to 2500 MHz. That circuit used special noise generating diodes from the Noise/Com Company in Parsippany, New Jersey. The noise source is still viable project after all these years. For a few years after the article appeared, Noise/Com provided a lab calibration service to Amateurs and experimenters for an especially low cost, as arranged by Gary Simonyan, the Noise/Com CEO. Mr. Simonyan has since retired.

This service, at the low price, was discontinued because it was not considered viable for the Company. Recently, however, Noise/Com has decided to *resume* the special offer, on a trial basis, for a fee of $100, as compared to the customary charge of $600. This service is restricted to domestic customers, because of Customs complications that have been experienced for these projects by Noise/Com. Suitable noise diodes are also specially-priced for Amateurs and experimenters.

The Noise/Com calibration can subsequently be transferred with good accuracy to other homebrew units by Amateurs and experimenters who have the necessary lab equipment and skills.

Contact Noise/Com by phone or e-mail for this service, and to request an RMA (Return Merchandise Authorization) for the calibration. For the special amateur pricing offer, contact:

Al Sebolao, Sales Manager, Noise/Com
25 Eastmans Rd
Parsippany, NJ 07054
Phone: (973) 386-9696
E-mail: **asebolao@wtt.bz**

The *QST* article is on the 1990-1994 *QST View* CD ROM.[1] I can also e-mail a PDF copy of the article for those who don't have access to it. Contact me at **w.sabin@mchsi.com**. The noise source project was described in *The ARRL Handbook* for many years. Circuit boards are available from:

FAR CIRCUITS
18N640 Field Court, Dundee, Illinois 60118
Phone: (847) 836-9148
E-mail: **farcir@ais.net**
Web: **www.cl.ais.net/farcir**

Circuit board template details are on the ARRL Web site. Search for "sabinns.pdf." A copy of this is also available from **w.sabin@mchsi.com**.
— 73, William E. Sabin WØIYH, 1400 Harold Dr SE, Cedar Rapids, IA 52403; **w.sabin@mchsi.com**

[1]1990-94 QST View, ARRL Order no. 5749. *QST View* CDs are available from your local ARRL dealer, or from the ARRL Bookstore. Telephone toll-free in the US 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

Tech Notes (May/Jun 2007)

Dear Editor,

Congratulations to the Tech Notes article, "Electromagnetic Radiation: A brief Tutorial." This note, written by Doug Smith, was excellent. I am looking forward to more good presentations like this, and hopefully from Doug.
— 73, Ulrich Rohde, PhD, N1UL, 52 Hillcrest Dr, Upper Saddle River, NJ 07458; **KA2WEU@aol.com**

Dr. Rohde,

Thanks for the note. I am pleased that you enjoyed Doug's Tech Notes article. I also hope that he will submit other articles for publication.
— 73, Larry Wolfgang, WR1B, QEX *Editor*; **lwolfgang@arrl.org**

A DDS Based QRSS (and CW) Beacon (Sep/Oct 2007)

Hello Larry,

I received an e-mail from a *QEX* reader about a possible error in the schematic diagram of Figure 1 in my article; I verified that he is right, Q2 and Q3 emitters are missing the connection to ground.
— Best Regards, Matteo Campanella, IZ2EEQ, SS Dei Giovi, 41/A, Binasco, Italy 20082; **iz3eeq@arrl.net**

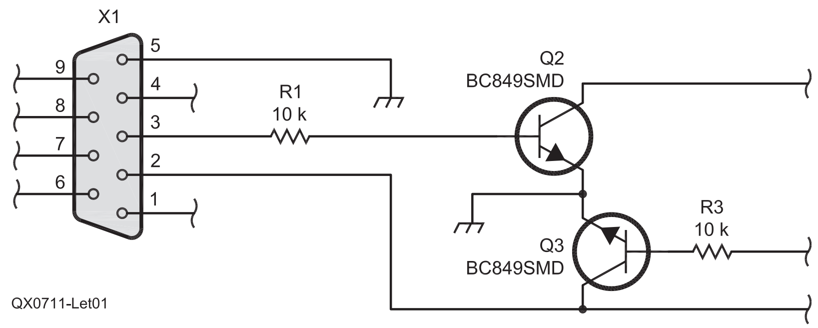

Figure 1 — In Figure 1 of "A DDS Based QRSS (and CW) Beacon," on page 25 of the Sep/Oct 2007 issue of *QEX*, the ground connection between the Q2 and Q3 emitters was omitted. This partial diagram shows the correction.

Hi Matteo,

Thanks for bringing this schematic error to our attention. Shown here as Figure 1 is a partial schematic of Figure 1 from the original article. We've added that missing ground connection. We apologize for the omission.
— 73, Larry, WR1B

Empirical Outlook (Sep/Oct 2007)

Hi Larry,

You asked for feedback, so let me add two thoughts that were triggered by your "Empirical Outlook" and one of the Letters to the Editor in the Sep/Oct issue.

First, I say "Amen" to K9OSC's letter. From that letter, I conclude that he and I have very similar views of what we expect and what we receive from *QEX*. I believe that he and I may be representative examples of a target audience that should be considered when you are selecting articles for publication. Now, I do not expect all articles to be aimed at this group. But, if you see one that fits, think of us!

Second, based on your description, Letters to the Editor really deserves a new column name. If *QEX* is the "forum for communications experimenters," that section of the magazine is the heart of the forum! And, I think your comments about its use should be included at the top of that page, in every issue.

Finally, I liked Issue No. 244 and I am looking forward to Issue No. 245!
— 73, Larry Keith, KQ4BY, 231 Shenandoah Trail, Warner Robins, GA 31088; **kq4by@arrl.net**

Hi Larry,

Thank you very much for sharing your thoughts. My intention is to strive for a mix of more and less technical articles for

QEX. I have been talking with a number of potential authors about some of their work, and will continue to encourage them to write the articles they may have thought were not technical enough for *QEX*.

I'm not quite ready to change the focus of the Letters column, but I do believe there are some ways to incorporate those "letters" that share a particular concept or describe ongoing projects. At the moment, my thought is to use the Tech Notes column to collect and share those types of correspondence. This month's column came from just such a letter.

I am excited about the article line-up in this issue, and I hope you like it as much as the last one. We have more excellent material in store for the new year, and a few surprises that I hope you will enjoy.
— 73, Larry, WR1B

Turbo Delphi Explorer (Sep/Oct 2007)

Dear Editor,

I just received the Sep/Oct 2007 issue. The Delphi article on page 44 was most interesting. For those who are not familiar with *Pascal*, there is available a very good document describing the language. Marco Cantu, an author of tech books on software, has written a paper on Pascal that would be very helpful to Pascal beginners. This is available at **www.marcocantu.com/epascal**. You might want to look around his Web site, as he has written a number of books on *Delphi*.
— Bill Goodwin, WB8BER, 2295 Knotwood Dr, Holt, MI 48842; **wb8ber@arrl.net**

Larry,

Thanks for the forward. Bill directs others to a great site. Marco Cantu is very well known. Readers should also check out the following in Marco's Web page: Under Site Menu, try the "Code Repository" and "Essential *Delphi*" articles. There are lots of coding examples, many of which will work with "*Explorer*" unchanged. There is also a free e-book on the *Delphi* Pascal language (not specifically for "*Explorer*" but almost everything written there will apply just fine). There are many more Web sites that are useful to beginners and experts alike. Just Google "*Delphi* hints" or "*Delphi* programming." I Wish I had thought to include a reference in the original article to Marco's site and it was nice of Bill, WB8BER, to point it out.
— 73, Steve Gradijan, WB5KIA, 1902 Middle Glen Dr, Carrolton, TX 75007; **steve_jg@msn.com**

Using Gain-Probability Data to Compare Antenna Performances (*QEX* Sep/Oct 2007)

Dear Larry,

I read with interest the article "Using Gain-Probability Data to Compare Antenna Performances" (*QEX* Sep/Oct 2007). I developed a similar approach to antenna evaluation back in 2000 (**www.qsl.net/g3cwi/hfantenna.html**). It is difficult to measure if the approach actually works in practice, but it does offer an interesting insight into the desirability of matching antennas to aspirations!
— Regards, Richard Newstead, G3CWI; **g3cwi@btconnect.com**

Dear Richard,

Thank you for your note, bringing your Web site to my attention. Indeed, your work is very similar to mine. In fact, if I had seen it earlier, it would have saved me a lot of work and research!

In many ways, your program is far superior to my spreadsheet approach to the problem, and that would have been my next step in the development of the idea. I think *QEX* readers will appreciate this addition to the concept.

Where you have included a weighted gain, I have considered probabilities of gain into certain regions. Both approaches are valuable and similar in results.

I am glad that someone else had the same idea. It gives credibility to the power of the approach and extends the state of the art. You obviously don't need convincing that this is a powerful approach to antenna performance comparisons.

Thank you for your good work.
— 73, Fred Glenn, K9SO, 320 Castlewood Ct, Palatine, IL 60067; **k9so@arrl.net**

An Unusual Vector Network Analyzer (Sep/Oct 2007)

Dear Readers,

As a number of you noticed, we had a layout glitch with Dr George Steber's article in the Sep/Oct issue. At the end of that article, on page 23, Note 7 was cut off before the end, and Notes 8 and 9 were omitted. We apologize for this omission. Printed here are all of the Notes for that article, for your complete reference.

Notes

[1]**cp.literature.agilent.com/litweb/pdf/5965-7917E.pdf**
[2]**www.national.com/appinfo/adc/files/ABCs_of_ADCs.pdf**
[3]**www.ar.com.au/~softmark/**
[4]**www.control-systems-principles.co.uk/whitepapers/frequency-response-analysis2.pdf**
[5]**www.k6mhe.com/n7ws/AN77-3.pdf**. Caution, this is a 13 MB file.
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[9]**www.rohde-schwarz.com/www/downcent.nsf/ANFileByANNoForInternet/F7E4CBCA92015B8AC1256B490052825E/$file/1ez35_1e.pdf**
— 73, Larry, WR1B

Tune SSB Automatically (Jan/Feb 1999)

Robert Dick called to update his contact information and ask that anyone with knowledge of any way his techniques were used in radios to please contact him. He wrote about this technique to automatically tune an SSB signal, but never implemented it himself. He has been told that a software defined radio may have been developed that used his technique or something similar, and is interested in hearing from anyone who actually used the method. He never attempted to patent the concept and has no financial interest, but would like to know that if the idea has ever been tried. If you have any knowledge of such use, please contact Robert Dick, 157 High St, Sanford, ME 04073; **gr2rojad@aol.com**
— 73, Larry, WR1B QEX

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Tech Notes

Upcoming Conferences

Miscellaneous Articles

In the next issue of QEX

Next Issue in *QEX*

Cornell Drentea, KW7CD, continues his series about the design of his *Star-10* transceiver. In this part, Cornell gives us a detailed look at the transmit/receive half octave filters, the 75 MHz first IF and the conversion to the second IF at 9 MHz. He also explains the operation of the master reference unit, which produces the stable signal that serves as a reference to phase lock all of the local oscillators.

Jack Smith, K8ZOA, describes the Z100, a CW, RTTY and data mode tuning indicator. Using a PIC18F2420 programmable microcontroller and a 24 LED display, this project will have you tuned to zero beat in no time. It's another must-read issue!

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