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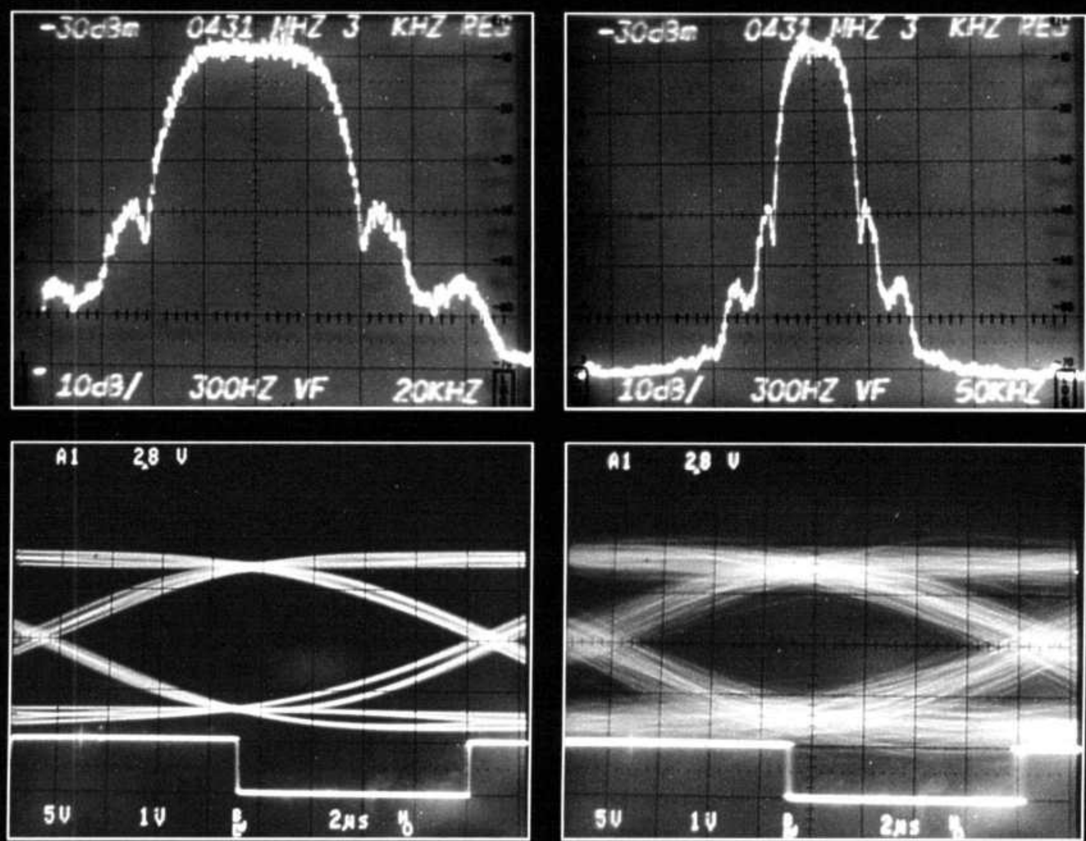

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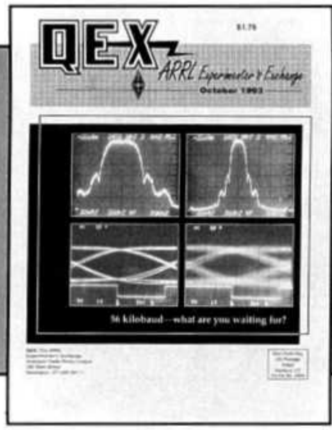

About the Cover:
56 kpbs: It's not just for breakfast any more! (WA4DSY photos, from *6th ARRL Computer Networking Conference.*)

140

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Features

Columns

October 1993 QEX Advertising Index

Purpose of *QEX*:

1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio field
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 correspondence for publication in *QEX* should be marked: Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking

In Conference

The 12th ARRL Digital Conference, known as the Computer Networking Conference in previous years, is history. Held in Tampa, Florida, on September 11, it was attended by about 50 people, who heard presentations and discussion of a number of topics of interest to today's amateur experimenter in—and implementer of—digital communications.

The proceedings of the conference (available from ARRL for $12 plus $3 shipping) contains the following papers: "Using Airborne Digital Repeaters in Emergency Communications," by Gary Arnold, WB2WPA; "A Prototype TNC-3 Design Approach," by Bill Beech, NJ7P, Doug Nielsen, N7LEM, and Jack Taylor, N7OO; "DSP Implementation of 300- and 1200-Baud FSK Packet Demodulators," by Jon Bloom, KE3Z; "The ARDS Project," by Roy V. Ekbert, WLIQ, and Martin R. Schroedel, K9LTL; "Usage of TCP/IP Over ROSE in the Tampa Bay Area," by Chuck Hast, KP4DJT/TI3DJT; "Unique Identification Signals for CCIR 625," by Michael J. Huslig; "Network Information Services," by Brian A. Lantz, KO4KS; "Networking ANSI Color Graphics," by Brian A. Lantz, KO4KS; "Interfacing Between ROSE and TCP/IP," by Thomas A. Moulton, W2VY; "Packet Radio in Disaster Situations," by Robert Osband, N4SCY; "The HUB 5/29 IP Routing Experiment," by Paul Overton, GØMHD, and Ian Wade, G3NRW; "A Low Cost Transceiver for 19.2 kb/s FSK Transmission in the 23 cm Band," by Wolf-Henning Rech, N1EOW/DF9IC; "Low Cost Entry Into Packet Radio Using BayCom," by Christopher C. Rendenna, KB2BBW; "Improved TNC Interconnections," by Donald A. Rotolo, N2IRZ; "Packet Tracker—A Graphical Packet Tracking Program," by Mark Sproul, KB2ICI; "Mail Tracker—A Graphical Mail Tracking Program," by Keith Sproul, WU2Z; "The Client/Server Bulletin Board System (csbbs)," by Jim Van Peursem, KEØPH, and Bob Arasmith, NØARY; "RMAILER: The Next Generation," by Frank Warren, Jr, KB4CYC; and, "SoftWire," by Aaron Wohl, N3LIW.

These papers represent what people are working on—or at least writing about—in amateur digital communications today. But do they represent what the amateur digital community *should* be working on? The answer is yes, but only in part. Conspicuous by their absence are substantial developments or applications that don't relate directly to packet radio. Where is the work on digital voice and image communications and/or processing? Where are the advanced techniques for using digital modes under poor propagation conditions? Where, in short, are we going with digital communications in Amateur Radio?

Packet radio as we amateurs practice it is almost 15 years old; the Vancouver Area Digital Communications Group packet developments on which our developments are based were in 1979! It's time—past time—to advance.

It's not as though we have no technology available to us. The WA4DSY 56-kbps RF modem, introduced at the 6th conference, in 1987, remains viable and available. (See this month's "Digital Communications" column for details.) So we can send faster bits. Isn't it about time we did so, and used those fast bits for some useful and interesting real-time communications?

This Month in *QEX*

Spectrum analysis is a valuable design tool, but spectrum *analyzers* tend to be expensive. Bruce Pontius, NØADL, presents some alternative approaches to "Narrowband Spectrum Analysis with High Resolution."

Our 6-part series on ALE continues with part 4, by David Sutherland and Dennis Bodson, W4PWF.

There's nothing quite as handy as a dip meter. Larry Cicchinelli, K3PTO, describes "A Dip Meter with Digital Display" that you can build.

Merle Rummel, W9LCE, has performed a number of computer simulations of troposcatter links using typical high-performance amateur stations. His results, and his methods, are documented in "Troposcatter—DX CQ."

Finally, Harold Price, NK6K, turns his "Digital Communications" column over to Barry McLarnon, VE3JF, to explain how you can get on 56 kbps.—*KE3Z, email: jbloom@arrl.org (Internet)*

Narrowband Spectrum Analysis with High Resolution

Sure, you can do spectral measurements with a multi-thousand dollar instrument. But what if you don't have one?

Bruce E. Pontius, NØADL

With digital modes becoming popular for data transmission over radio, and digital modulation techniques in use for voice as well, analysis of baseband modulation and radio-frequency occupied bandwidths is of considerable interest. Unfortunately, this usually requires the use of very expensive high-resolution spectrum analyzers.

This article describes some relatively simple and low-cost circuits which provide high-resolution spectral analysis, both at baseband, or audio frequencies, and over narrow ranges of radio frequencies.

Typically, low-cost spectrum analyzers have low resolution, low frequency stability, and limited dynamic range, so they are not suited for narrowband analysis and cannot separate (resolve) signals less than a few kilohertz apart. A typical lowest resolution band width (RBW) of 1 kHz is barely good enough for modulation analysis in FM systems and is not suitable for narrowband systems such as SSB. A spectrum analyzer with 300 Hz or better RBW, good frequency stability and dynamic range of more than 70 dB can cost as much as $28,000, but all of the other features that might come with it are not absolutely necessary for use as a narrowband analyzer. Older used units with required RBW may be available for under $4000, but maintenance could be a problem. The arrangement shown here can provide, with care exercised in use, good enough resolution, stability and range to perform detailed analysis of complex signals from audio to radio frequencies.

Eventually, digital signal processing techniques will provide the tools for the applications described in this article, but the performance reasonably available to the Amateur Radio enthusiast at this time is marginal. A computer-based audio spectrum analyzer is described in the 1993 edition of *The ARRL Handbook for Radio Amateurs*, based on an article in January 1992 *QST*.[1] This approach is certainly worth looking at for signals with limited bandwidth. However, I had all of the equipment and materials on hand to build the special purpose analog analyzer described in this article, which gives very good performance at low cost, so I chose this approach.

Spectrum Analyzer Basics

Chapter 25 of the 1993 *Handbook*, and earlier editions, contains a good discussion of spectrum analyzer basics, specifications and applications. As pointed out there, a spectrum analyzer can be described as simply an electronically tuned receiver which measures voltage or power as a function of frequency. In fact, meaningful spectral measurements can be obtained using a well characterized and calibrated communications receiver. Measurement of intermodulation distortion in SSB transmitters is a subset of general spectral measurements and is readily accomplished with just a communications receiver and maybe a variable attenuator.[2]

If a communications receiver with a good S-meter circuit and display meter is available, it can be used for making general spectral measurements. The S-meter circuit should have a well-behaved response and have an easily readable meter with adequate scale markings. The receiver serves as the IF, detection and display sections of a manually tuned spectrum analyzer with a mixer and separate local oscillator serving as the front end to afford handling signals at frequencies outside the receiver's range, such as audio-modulation frequencies. Fig 1 is a block diagram of such a system.

I had at least two candidates on hand with fairly well marked S meters; a Kenwood TS-440S and a Drake R7. Various signals were used to check relative responses to voltage and power. Calibration curves for CW signals are shown in Fig 2. The Drake has a well-behaved logarithmic response usable over a 90-dB range, from 60 dB

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[1]Notes appear on page 12.

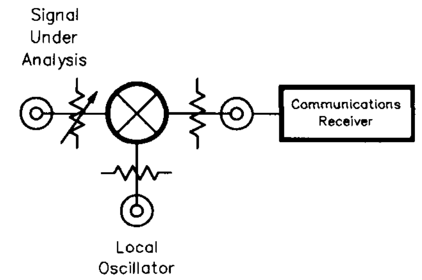


Fig 1—Simple narrowband spectrum analyzer using a communications receiver.

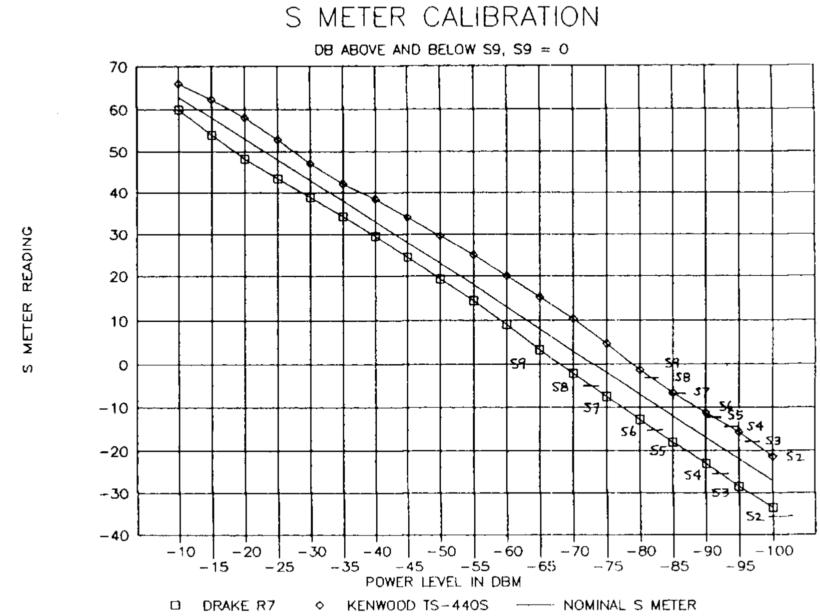


Fig 2—Response curves for ideal and actual receiver S meters.

over S9 to almost 35 dB below S9. It exhibits a consistent 5 dB per S unit below S9. These receivers would be suitable for measuring discrete signals such as intermodulation (IM) products, but problems could arise in interpreting the values for some other types of signals. Also, most receivers have much less IM dynamic range for close-in signals which are within the second IF passband than they exhibit for signals spaced 20 kHz apart, where the IM dynamic range is usually measured.

One of the problems is that each different receiver's handling of complex signals and noise is not well defined, so it might be difficult to compare data taken in different locations with different equipment. It would be better and possibly easier to measure the power in the signals directly. In the case of noise signals, an important signal category due to the use of broadband noise in all sorts of measurements (bit error rates, or BER, for example), we want to know the actual power in the slice of spectrum under test, not simply a circuit's particular response to the signal voltages and the attendant interpretation of peak and average power.

The true RMS values of signals are desired, which means that crest factors and bandwidths are important. W. Sabin built a true RMS reading device, described in his *QST* article on SSB/CW receiver sensitivity measurements, to properly handle mixed signals and noise.[3]

A simple approach which I use for measuring the power in baseband complex signals and noise is shown in Fig 3. The local oscillator and a mixer with known transfer characteristics translate the signal (in this case an audio or baseband modulation signal) up to a frequency where a thermistor-mount power meter can measure the actual power (above 9 MHz). This makes a broadband measuring instrument. This approach is particularly useful when setting the ratio of two complex signals, such as in BER measurements, where bit errors are measured as a function of data signal-to-noise ratio. The data signal in the audio range varies in amplitude and frequency as does the noise signal. The actual power in the waveforms is measured reliably by the thermally responsive power meter. This arrangement provides for measurements of wideband signals which may exceed an RMS voltmeter's range.

For the reasons discussed above, and because of its accuracy and readability, the power meter is a good detector and indicator for the spectrum analyzer. These units are available for reasonable prices on the surplus market and they are quite valuable around the lab for power measurements from 9 or 10 MHz through X band.

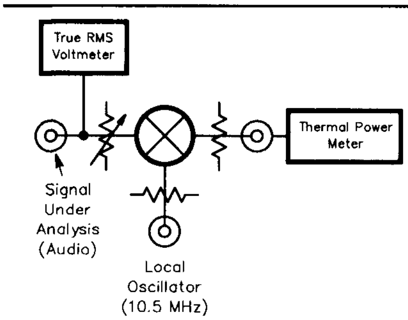


Fig 3—A simple approach to measurement of complex baseband signals or noise. An HP431B power meter with HP478A thermistor mount is used by the author. A true RMS voltmeter such as the Fluke Model 87 can be used at the imput.

Other types of detectors could be substituted here, especially for measurements using a narrow filter. A logarithmic AGC amplifier circuit and meter, or a receiver, can be used if a power meter is not available. One of the cellular IC chips with an RSSI circuit could be used as well.

Circuit and Equipment Description

The block diagram in Fig 4 shows the arrangement for a narrowband spectrum analyzer with high-stability and good resolution bandwidth, coupled with 70 dB or more of dynamic range for closely spaced signals.

The signal or signals under test are attenuated properly and inserted into the mixer's IF port. A stable, low-noise, finely adjustable source serves as a local oscillator to translate the signal to the frequency range of the crystal filter. A variable attenuator sets levels and provides part of the level readouts. The linear amplifier has two tuned stages and an untuned stage and provides the gain required to drive the power meter, as well as providing bandwidth limiting to reduce wideband noise. An oscilloscope, recorder or other device can be attached to the power meter's output connector to supplement the panel meter.

Stable, Low Noise Generator

The local oscillator-generator can be a VFO operating at around the center frequency of the crystal filter. Suitable circuits are described in Note 1

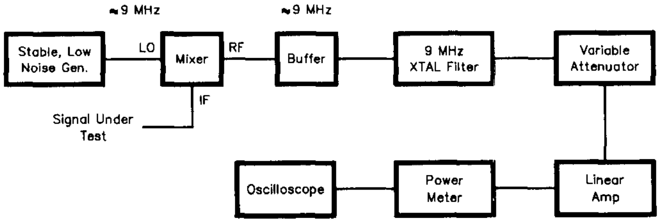


Fig 4—Narrowband spectrum analyzer (NBA).

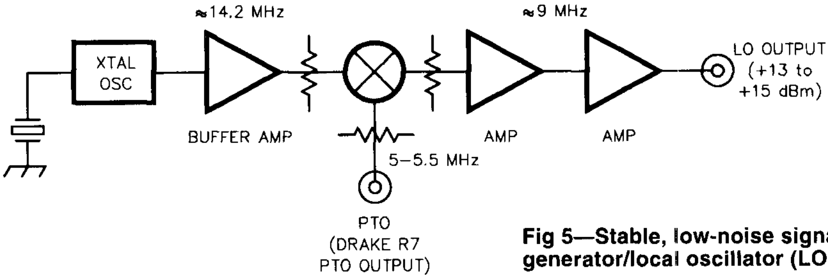


Fig 5—Stable, low-noise signal generator/local oscillator (LO).

and in the book *Solid State Design For The Radio Amateur*, available from the ARRL.[4] Also, Note 5 includes a 10-MHz VFO which would probably be satisfactory when moved down to the crystal filter's frequency of 9 MHz. The tuning range need be only 100 or 200 kHz to provide enough operating flexibility. This would allow a range from dc to 200 kHz or plus and minus 100 kHz around an RF frequency which was translated down to the analyzer's input. The output level should be +13 to +15 dBm.

I used the Drake R7 as a tunable VFO since it had the desired stability, low noise and spurious, and continuous, smooth tuning. Of course the 5- to 5.5-MHz VFO output had to be shifted up to the 9-MHz range of the KVG crystal filters used. Fig 5 is a block diagram

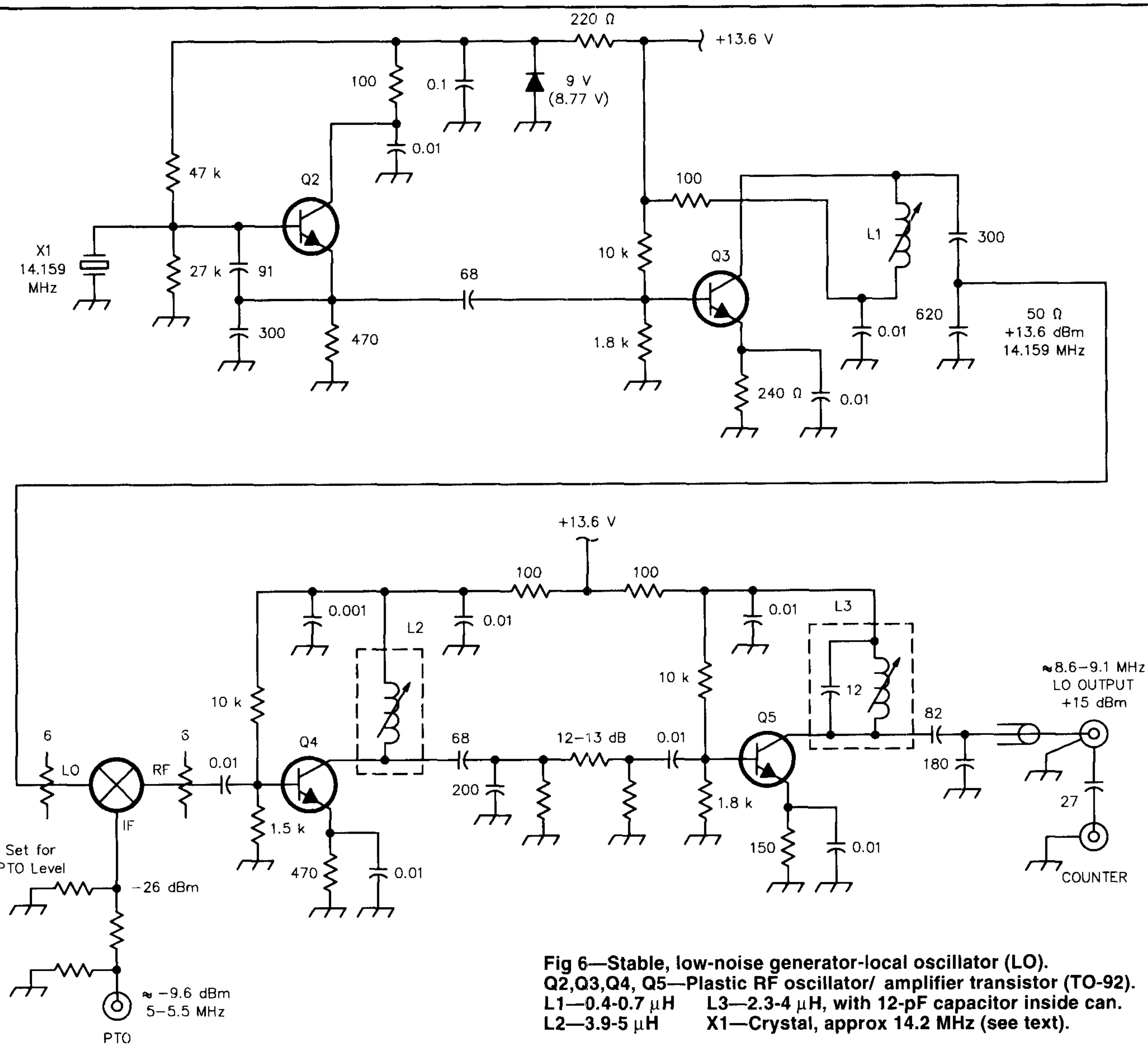


Fig 6—Stable, low-noise generator-local oscillator (LO).
Q2,Q3,Q4, Q5—Plastic RF oscillator/ amplifier transistor (TO-92).
L1—0.4-0.7 μH L3—2.3-4 μH, with 12-pF capacitor inside can.
L2—3.9-5 μH X1—Crystal, approx 14.2 MHz (see text).

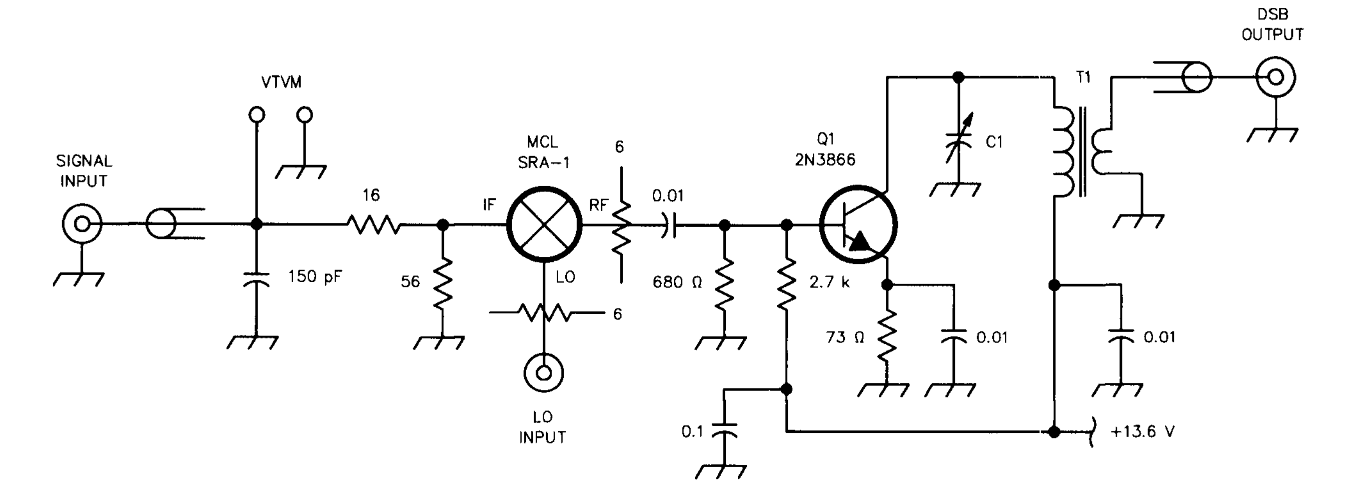


Fig 7—Signal mixer and buffer amplifier.
C1—Arco 404
T1—Primary 26 t no. 26, secondary 7 t no. 26, on T-50-2 core.
Q1—2N3866

of the LO source using the Drake PTO. Fig 6 shows schematics and parts values. I used a 14.159-MHz crystal left over from a previous project which provided a suitable frequency coverage range. The exact frequency does not matter since a counter was used to read the LO frequency. The receiver's readout could be used, with the operating frequency computed for each reading (done in a spreadsheet on your computer).

Referring to Fig 6, the desired signal power levels are given in the diagram. The 14-MHz signal for the mixer LO can be between +13 and +15 dBm. The PTO signal is inserted into the mixer's IF port. The level at the IF port is set at around –26 dBm by the attenuator. Two stages of amplification (at Q4 and Q5) bring the level up to the range of +13 to +15 dBm, with the interstage attenuator used to set the output level.

Signal Mixer and Buffer Amplifier

The signal mixer and buffer amplifier are shown in Fig 7. Signal levels become very important here in order to keep circuits operating well below compression to reduce unwanted intermodulation products. A VTVM is used to set and monitor the input level to the analyzer. Once all circuits are properly set up, the output level on the power meter becomes the indicator for operation within the proper ranges with the

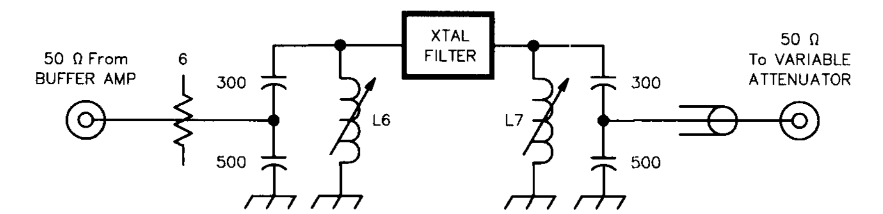
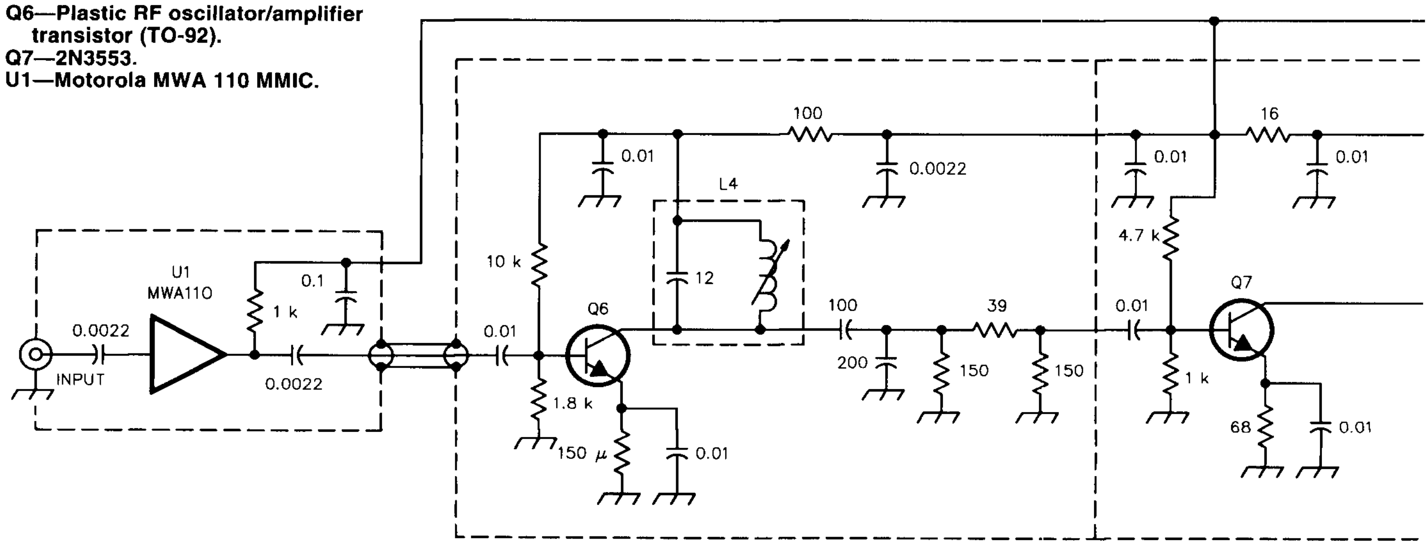


Fig 8—Crystal filter.
L6,L7—1.6 μH nominal.

Fig 9—Three-stage linear amplifier.
L4,L5—2.3-4 μH, with 12-pF capacitor inside can.
Q6—Plastic RF oscillator/amplifier transistor (TO-92).
Q7—2N3553.
U1—Motorola MWA 110 MMIC.



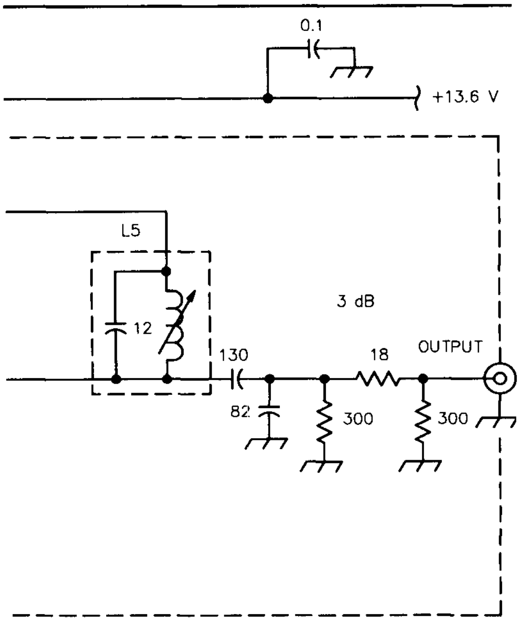
VTVM used as a check and input level monitor.

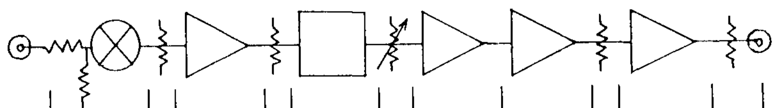
The mixer IF input is padded and matched to 50 ohms with a resistive attenuator. (The mixer has an 80- to 100-ohm input resistance at the IF port.) The resistor pad results in a 50-ohm input resistance with a 3.4-dB loss. Attenuators are used on all ports for predictable performance.

A fairly strong amplifier follows the mixer to build the signal level while keeping intermodulation to the range of 60 dB less than a desired signal (> 60 dB below one tone of a two-tone signal for example). This is the stage where intermod can cause distortion and errors in the measured results for a test signal. The output level from this amplifier stage should not be allowed to go above −2 dBm peak, or −8 dBm on a single tone or CW signal. In the case of a two-tone audio signal with two equal tones, the maximum levels are −2 dBm peak, −5 dBm average, and −8 dBm for each tone. Then IM will stay below 60 dB down.

These numbers refer to maximum-strength discrete signals. In actual use, the signals are usually spread across a band of frequencies with no individual components within several dB of the average power. This reduces the intermodulation even further to give an IM range of well over 70 dB for signals spread over a bandwidth of a few kHz or more. A further discussion of this is contained in the section on applications and measurements.

The diagram in Fig 10 shows the desired signal levels throughout the analyzer.





< 4 mv	Mixer	−6	Post Mixer Amplifier	−6	9 MHz Crystal Filter	V	Amplifier	Amplifier	−6	Amplifier	−3
Gain (Av) (dB)	(−3.4) (−6.4) −9.8	−6	43.7	−6	−9	−52	+14	+36	−6	+35	−3
Signal Level (dBm) Per Tone	−39.3 −35.9 −45.7		−51.7 −8		−14 −23		−75 −61	−61 −25		−31 +4	+1
3rd–Order Output Intercept (dBm) Per Tone	~+4		+22		+6		+6	+16		~+22	
1 dB Input Compression Point (dBm)	−5		~+16				−2.5	+8		~+16	

Fig 10—Gain distribution for the narrowband spectrum analyzer.

These circuits are satisfactory for the intended purposes. However, higher dynamic range could be achieved with the circuits given in Note 6, which use a high-level mixer and a high-bias, push-pull IF amplifier. Circuit complexity is greater, however.

Crystal Filter

I used KVG XF-9NB and XF-9D filters. These are nominally 500-Hz and 5-kHz filters, respectively. The narrow filter was used for modulation and occupied bandwidth measurements, and the 5-kHz filter was used for measuring the power in a 5-kHz FM channel near a transmitter frequency with data modulation. Since all interfaces are 50 ohms, other filters can be inserted for other purposes, such as a 2.4-kHz filter for use in determining interference to an SSB channel.

A 6-dB attenuator, shown in Fig 8 before the filter, provides improved broadband interstage matching and ensures stability of the mixer buffer amplifier.

The match into the MIC MWA110 amplifier on the output of the filter, shown in Fig 9, after the variable attenuator, is good enough over a broad bandwidth that a pad is not required even when the variable attenuator is at zero dB.

A variable step attenuator can be built as described in the ARRL *Handbook*.

Three Stage Linear Amplifier

About 80 dB of gain is required to get a level sufficient to drive a thermistor-mount power meter. In Fig 9, two tuned stages yield 71 dB of gain, or 62 dB with a 6-dB attenuator between stages and a 3-dB pad on the output. (This output pad provides a dc return for zeroing the power meter into—also, the power meter mount is not 50 ohms at the low frequency of 9 MHz.) A 14-dB gain monolithic amplifier increases the overall gain to a satisfactory level of 76 dB, as shown in the gain distribution table in Fig 10. The output stage could be the same as the mixer output buffer amplifier in Fig 7. The amplifier shown in Fig 9 was available from another project and was used here rather than building another one.

An on-off switch is convenient for turning the power to the amplifier off while zeroing the power meter. An on-off switch for the stages before the crystal filter is also useful while checking levels.

The amplifiers and the other circuits should be well shielded to prevent signal leakage across the crystal filter and coupling from the output to the input. Stability depends on shielding.

Components Used

Small value capacitors can be ceramic or silver mica. Silver mica were available to me, so I used them. Coupling and bypass capacitors are all ceramics and are not critical. All resistors are ¼-watt carbons. The attenuators shown as resistor symbols are Mini-Circuits units but could be made with ¼-watt carbon resistors.

Gain Distribution and Signal Level Chart

Gain distribution and desired signal levels are shown in Fig 10. The overall gain from input to output is about 37 dB. Variations from the gain distribution shown are acceptable as long as the maximum levels are not exceeded, which could result in nonlinear operation.

At the levels shown, the mixer is well out of the excessive IM generating range. The post mixer buffer amplifier is the limiting stage. Third-order intermodulation performance of this stage

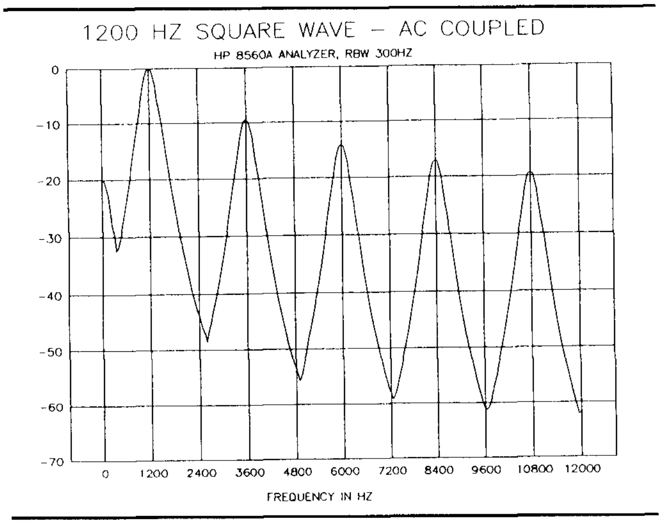


Fig 11

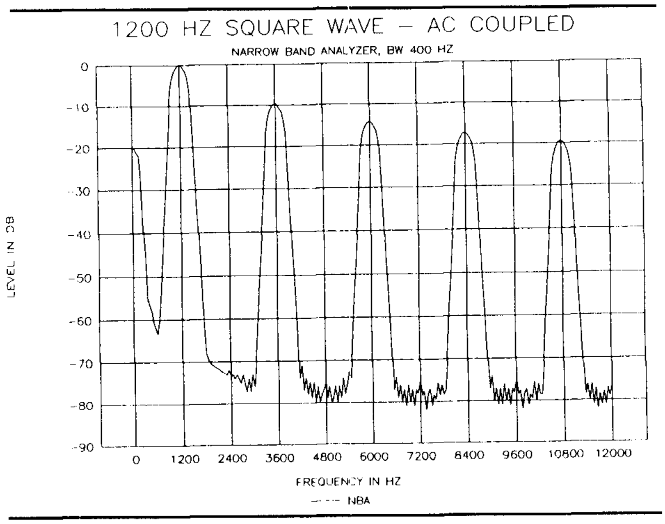


Fig 12

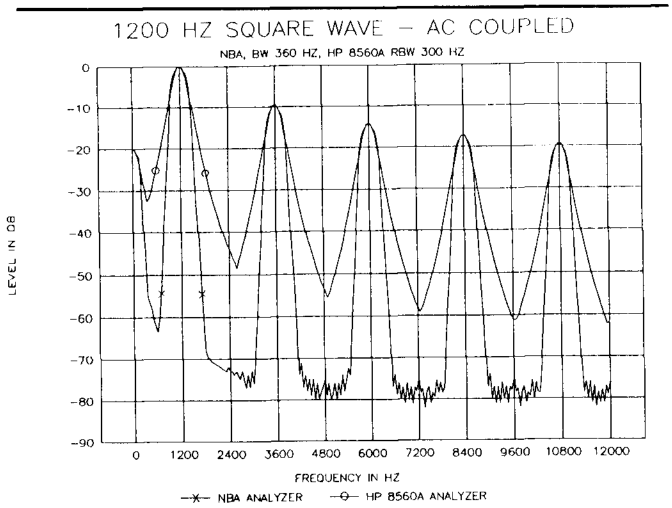


Fig 13

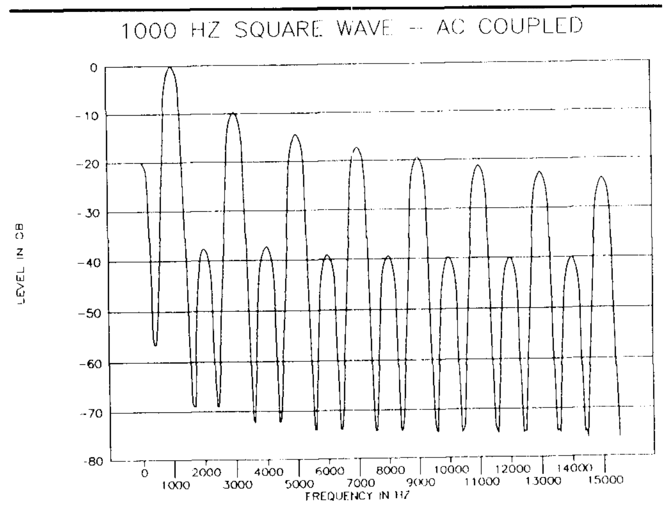


Fig 13A

was measured using a two-tone signal. Third-order products were 60 dB below each tone at –8 dBm per tone output, or 66 dB below peak at –2 dBm peak output (average power as read on a thermistor-mount power meter is –5 dBm, with both side bands included). This is the maximum output level for maintaining good dynamic range.

When using the instrument, input power can be measured approximately with a VTVM at the mixer input (at the connector as shown in Fig 7) and can be measured accurately with the power meter at the output of the post mixer buffer amplifier. The more important value is the output of the post mixer buffer amplifier, which should be held below –5 dBm average power.

A single sinusoidal signal at, say, 1 kHz, at a level of 3.58 millivolts RMS, is inserted at the input and results in the levels shown in the chart.

The variable attenuator can be used to set the output level at the power meter to around 0 to 1 dBm. This gives about 52 dB of range with the attenuator alone and, with about 20 to 25 dB of range contributed by the power meter, total range is 70 dB even with signals which are spread out over frequency. The meter gets hard to read at low levels due to noise fluctuations but the eye and brain make a quite usable video-bandwidth filter.

Applications and Examples

Applications and example measurements are discussed in the following sections.

Measurements of Known Signals

Performance of the narrowband spectrum analyzer can be examined by using known signals and comparing results.

A symmetrical square wave at 1200 Hz was inserted into an HP 8560A spectrum analyzer, resulting in the display shown in Fig 11. The same signal was inserted into the NBA analyzer described herein, resulting in the data displayed in Fig 12. (Measurements were made out to the 15th harmonic but the figures show out to the ninth.)

Table 1 lists the levels of the harmonics of the signal measured with the circuits described in the NBA. Values measured are very close to those for an ideal square wave.

Table 1

Square Wave Harmonic Levels

Fundamental	0	reference	0
Third	–9.7		–9.54
Fifth	–14.2		–14.0
Seventh	–17.2		–16.4
Ninth	–19.3		–19.1
Eleventh	–21.2		–20.8
Thirteenth	–22.7		–22.3
Fifteenth	–24.0		–23.5

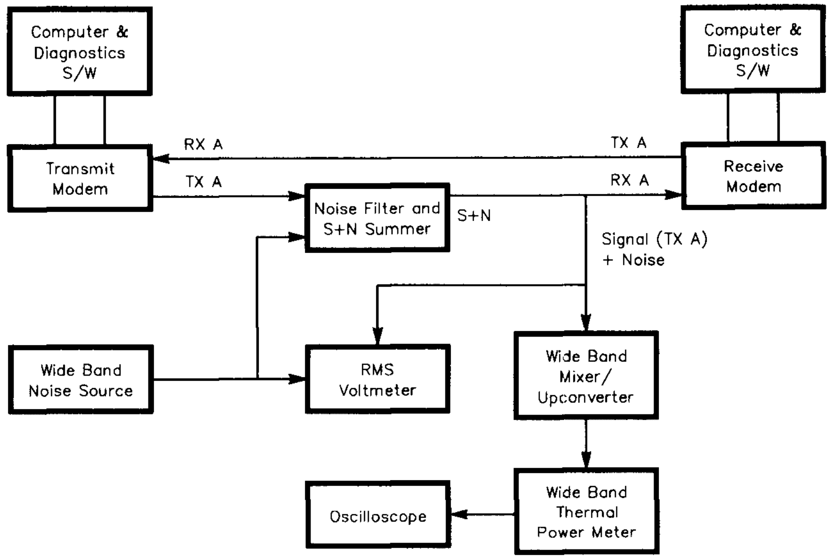


Fig 14—Setup for bit-error rate measurement.

Fig 13 is a combination of the HP 8560A data and the NBA data. The shapes of the pips are the shapes of the IF filters used in the instruments. Near the top the HP pip is narrower but lower down it is wider due to the use of fewer poles in the filter response. The filter response in the HP is carefully designed to accommodate fast electronic sweeping of the frequency whereas the filter in the NBA is an 8-pole communications filter and the NBA must be manually tuned (but costs thousands of dollars less).

There is ripple in the response of the 8-pole filters used in the NBA, but the passband ripples have been smoothed out in the presentation of the data. The peak response in the middle of the filter's band is always used for level measurements. As one tunes a CW signal through the filter's bandwidth, three or more peaks are noted, but the higher, middle one stands out sufficiently to afford accurate measurements (care must be exercised to avoid misreadings). Noise or data signals have energy at many frequencies and the power meter simply reads the average power in the filter's bandwidth, so the ripple does not matter. The NBA with the XF-9NB filter is equivalent to a 360-Hz RBW analyzer. I rounded that off and call it 400-Hz RBW.

The capability of the NBA to resolve signals is demonstrated by the plot in Fig 13A of a 1000-Hz squarewave which is not quite symmetrical, resulting in even harmonics between the odd harmonics. When the signal was dc coupled, there was a high value of energy at zero frequency which is resolved by the NBA. Signals only 500 Hz apart can be resolved.

Of course, narrower RBW settings are available on the HP but are not generally used in the applications discussed here. (RBW settings below 200 Hz or so use digital signal processing to process and display the signals.)

Measurements Of Digital Modem Signals

Some interesting modems using DSP-based signal generation, modulation and demodulation, and incorporating error correction, were tested using the equipment described above.

First, the setup in Fig 14 was used to measure bit-error rates in the presence of controlled noise. The wide-band mixer/upconverter of Fig 14 is the arrangement shown in Fig 3 and is very useful in making easy and accurate measurements of the wide-band noise signals and modulation signals, especially when noise and data are combined.

Modems were then connected to the signal-under-test port on the NBA as shown in Fig 4. Proper impedance (resistance) matching and attenuation must be used between the modem outputs and the mixer IF input. The level can be set by using an unmodulated mode generating a CW sinewave signal out of the modem. Set the attenuators so the output of the NBA is 0 to +1 dBm as discussed. The VTVM on the input in Fig 7 should read 4 mV or less.

After the input level is set, modulation is turned on using either a random digital signal or by sending a long text file. The pattern does not change noticeably between a random data stream and a text file.

The output level of the analyzer will drop several dB when modulation is applied. With a CW signal, all of the energy is at essentially a single frequency and all of the energy is within the passband of the NBA (or within the filter's passband). When modulation is applied the spectrum spreads out as shown in Fig 15. The voltage level stays the same but the signal is spread over a range of frequency. The power in any one slice of the spectrum is then about equal to the ratio of the modulation bandwidth to the slice (filter) bandwidth. A familiar example is frequency modulation of a transmitter, which will be discussed below when we look at RF spectra.

The input level should be left where it was set for the CW signal. This reduces the measurement range by 7 dB or more but ensures remaining below levels where intermodulation products could affect readings. This still affords plenty of range for modulation analysis.

The NBA PTO or local oscillator is then carefully tuned through the signal and the levels are recorded. The eyes and brain are the video filter in this simple machine, so just estimate the value from the fluctuating meter. If enough data points are taken, the results can be more accurate than that which the expensive conventional spectrum analyzer provides. The recorded numbers can be transferred to a computer math graphics program or to a computer spreadsheet program for processing and plotting.

DGMSK Modulation

For a special case of GMSK modulation called differential GMSK, or DGMSK, the spectrum of the 4800-bit/s modem baseband modulated signal looks like Fig 15. (GMSK is Gaussian

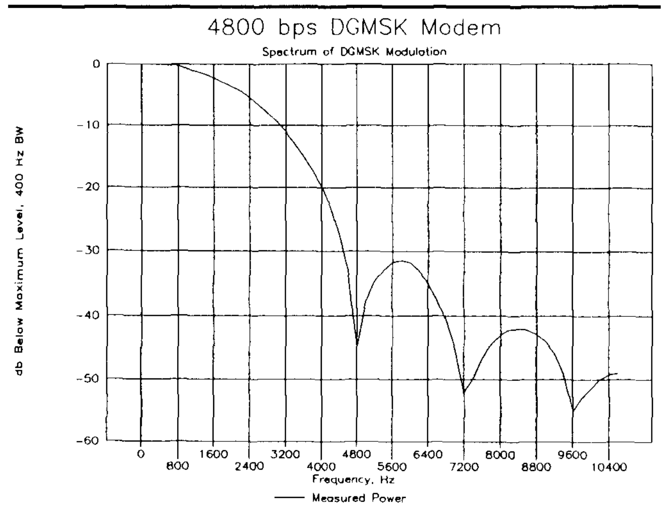


Fig 15

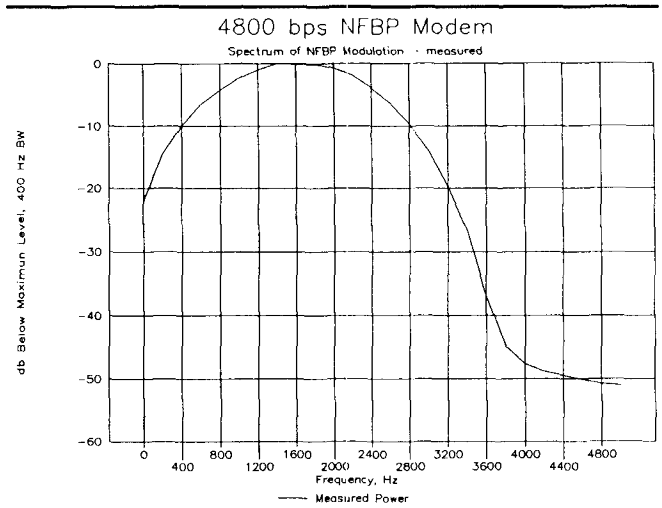


Fig 16

[shaped] minimum shift keying.) Notice that the spectrum goes down to dc or zero frequency, indicating the requirement for dc response in the transmitter's modulation network. Also note the multiple nulls and peaks extending out in frequency, with a null at the bit rate and repeating at one half of the bit rate thereafter.

NFBP Modulation

Nyquist-filtered bipolar modulation, NFBP, from another modem running at 4800 bit/s is shown in Fig 16. This modem output is well filtered beyond about 4000 Hz and does not require transmitter modulator response down to dc, as shown by the spectrum plot.

Some local oscillator feedthrough at about 20 dB down in the NBA prevents a null at dc. If it was desired to more accurately explore the area of a signal near zero frequency, the mixer balance could be improved to reduce local oscillator feedthrough. This is not usually required.

Radio Frequency Spectrums

Radio frequency spectra and occupied bandwidths were measured using the NBA with the test setup of Fig 17.

E. F. Johnson DL3470 UHF FM transceivers were used for these tests, with frequency deviation set to 4 kHz. The same units were used for each modem's testing. These transceivers were designed specifically for data use. The modulation transfer characteristic curve is shown in Fig 18. Frequency deviation is very linear with audio modulation voltage, as measured by observing carrier and sideband nulls, which is a more accurate measurement technique than some transfer calibrated modulation meters. (That good old communications receiver comes in handy again here.) The lower curve (also almost a straight line) is the dc response. Time and temperature variations have some effect on the dc measurements. In summary, the DL3470 modulation is a very linear function of voltage applied and is constant with modulation frequency out to more than 10 kHz.

Transmitter deviation must be set carefully for best BER performance with most amateur transmitters because non-linearities in the transfer functions, especially at high baud rates, can cause errors. It is best to keep deviation below 4 kHz with most equipment. Also, occupied bandwidth increases rapidly with increased deviation, which can cause interference to other spectrum users. When analyzing various modulations we found that increasing deviation from 4 kHz to 5 kHz typically causes a 20-dB increase in the RF sidebands out at ±7.5 kHz from the carrier. The NBA described here is quite useful in gaining a quantitative understanding of these types of issues.

The UHF transmitter signal is properly attenuated (exciter only used) and translated down to some frequency in the range of 25 to 75 kHz for insertion into the signal-under-test port of the NBA. The exact frequency is not important since the NBA has at least a few hundred kHz of flat-response tuning range. The NBA PTO or local oscillator is then carefully tuned through the signal and the levels recorded.

Fig 19 is the radio frequency spectrum of the DGMSK modulation and Fig 20 shows the spectrum of the NFBP modulation. The unmodulated carrier level was used to set the input level to the NBA for a reading on the power meter of 0-dBm output. With modulation applied, the signal spreads out with the total power remaining the same. For example, the unmodulated carrier was 7 or 8 dB above the 0 reference level in the diagram in Fig 19.

Of course, wider spans can be made but taking data gets to be tedious (it is easy with the $28,000 digital storage analyzer and a plotter). Power in the

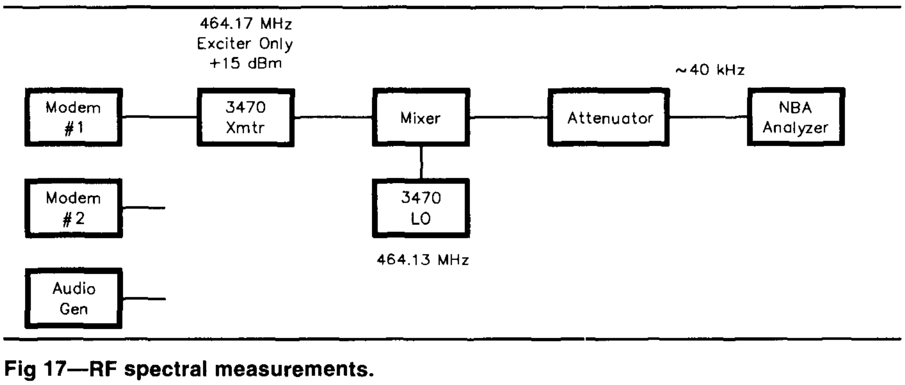


Fig 17—RF spectral measurements.

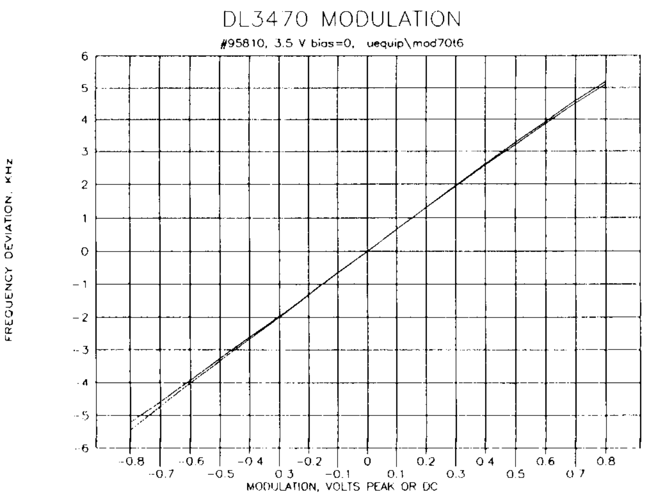


Fig 18

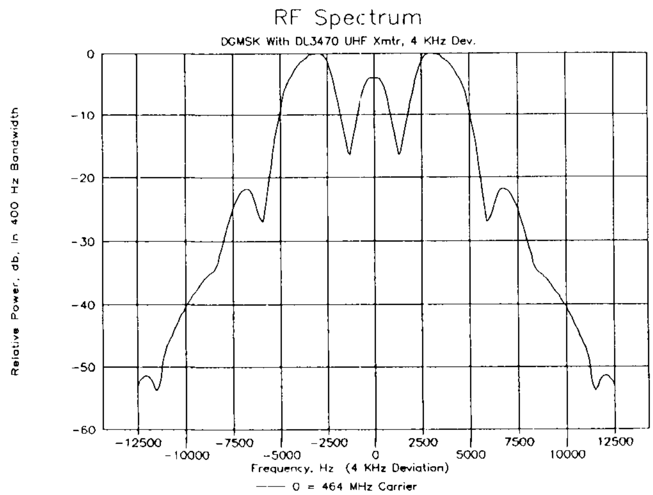


Fig 19

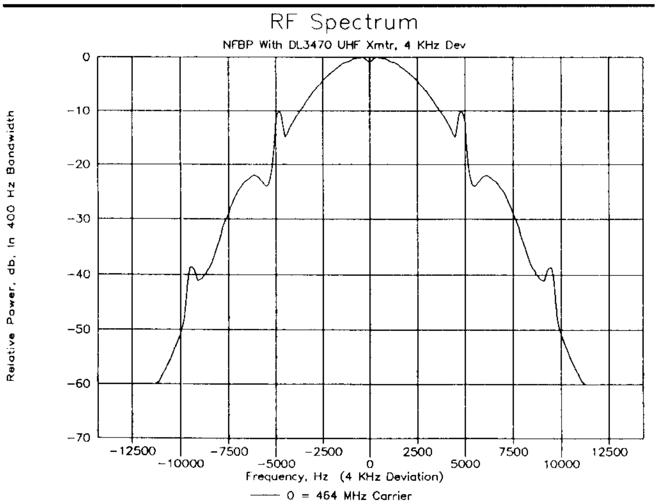


Fig 20

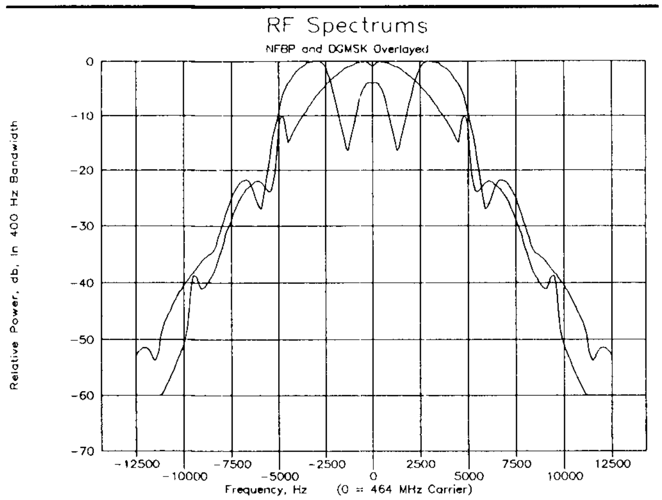


Fig 21

adjacent channel can be measured using a 5-kHz filter in the NBA and letting the power meter do the power integration over the adjacent channel receiver's bandwidth.

Comparisons can be made of different patterns and occupied bandwidths once the data is put into the computer. Fig 21 is a plot of the two RF signal spectra overlayed, while Fig 22 shows the corresponding audio range base-band signals.

Another example application is the measurement of an SSB transmitter output spectrum shown in Fig 23. The transmitter was modulated with broadband noise and the output plotted with the NBA. The inside curve, marked with an X, is the response curve of the SSB filter used in the SSB generation circuits. This filter was designed for use in a controlled-carrier SSB transmitter handling digital signals and is somewhat wider than a voice-only transmit filter. The curve marked with an O is the transmitter output when driven fairly hard. Notice the typical third-order IM products in the humps on the sides of the main signal. The curve marked with the triangles is the spectrum resulting when the output is reduced by 9 dB. The

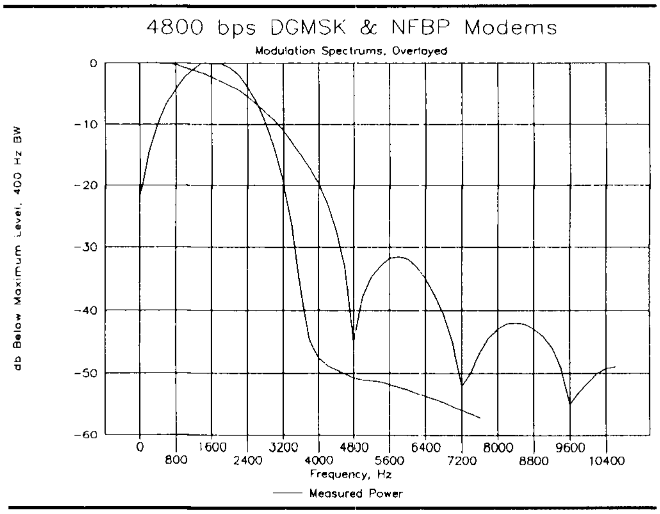


Fig 22

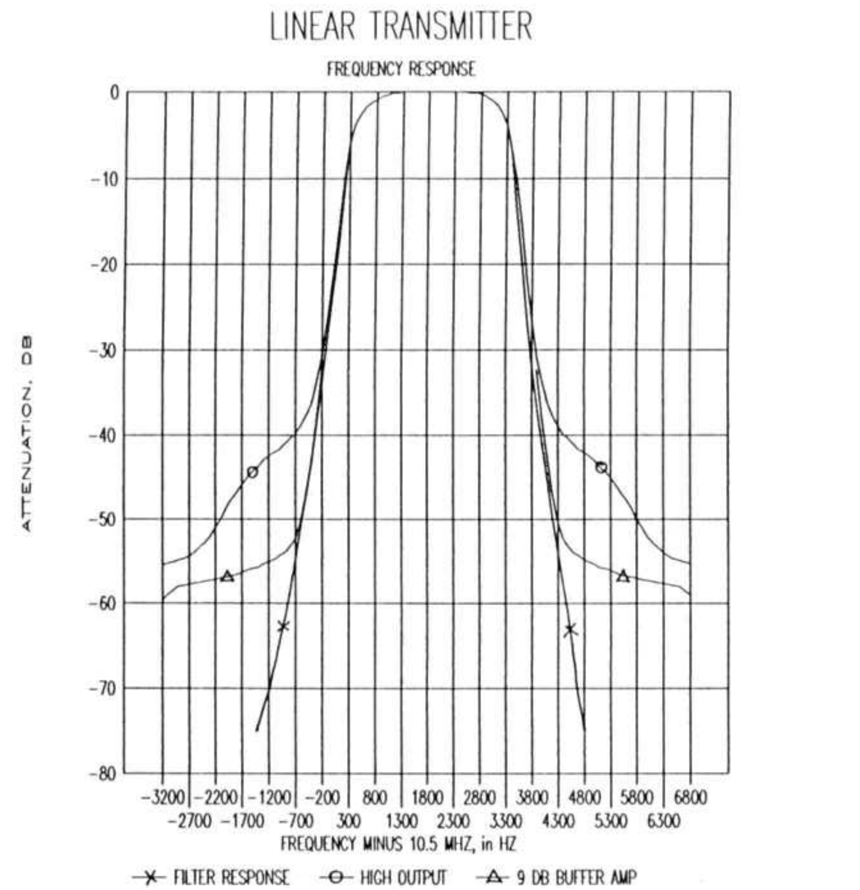


Fig 23

third-order products drop down considerably.

A conventional two-tone signal IM plot could be measured as well, but this plot in Fig 23 shows the actual signal which results from data modulation, which is much like noise with many types of modulators.

The lack of a narrow video filter is most noticeable when working with these noise signals because the meter fluctuates quite a bit at lower signal levels. Just guess at the average value, and, as these plots demonstrate, the results will be satisfactory if enough data points close enough together are taken. Fluctuations can be averaged out or smoothed out during data processing and analysis, resulting in a good plot.

The narrowband analyzer described here has proven to be accurate and useful for spectral analysis of a wide range of signals, although the low cost and relative simplicity require careful use and some time and effort as compared with an expensive commercial analyzer.

Notes

[1] *The ARRL Handbook for Radio Amateurs*, 1993 and earlier editions.

[2] Gonsior, M., W6FR, "TX IMD Performance And Measurement Techniques," *Communications Quarterly*, Summer 1991, pp 61-65.

[3] Sabin, W.E., WØIYH, "Measuring SSB/CW Receiver Sensitivity," *QST*, Oct 1992, pp 30-34.

[4] *Solid State Design For The Radio Amateur*, published by the ARRL.

[5] Makhinson, J., "High Dynamic Range Receiver," *Communications Quarterly*, Nov 1990, pp 75-76.

[6] Makhinson, J., "A High Dynamic-Range MF/HF Receiver Front End," *QST*, Feb 1993, pp 23-28.

About the Author

Mr. Pontius is a graduate in Electrical Engineering from the University of Washington and has been involved in the development of semiconductors for communications applications and development of radio equipment and systems for thirty years. He played major roles in the development of early cellular equipment, trunking radios and systems, and narrowband radio equipment.

He served as Vice-President Engineering at E. F. Johnson Company from 1973 until 1988, and worked with UPS for a time after that on their narrowband data radio system. Mr. Pontius is now President of TRM Associates and is working with a number of companies in the data radio field.

The Growing Family of Federal Standards for HF Radio Automatic Link Establishment (ALE)

Part IV: Network Simulation for the Radio Amateur

Complex network designs need to be thoroughly tested. The first step of testing is simulation.

David Sutherland and Dennis Bodson, W4PWF

Introduction

An important question stems from the recent series of *QEX* articles on the Status and Future of HF Digital Communication, and that is "How can radio amateurs use computer simulation modeling to investigate questions in the field of Amateur Radio?"[1]

The purpose of this article is to interest the radio amateur in computer simulation modeling as an experimental method. It is not our intent that the radio amateur could read this article and immediately program a simulation. However, it is our contention that simulation modeling is accessible to the radio amateur. The most important requirement is a thorough knowledge of what the investigator is trying to simulate. So, why would you even consider simulation modeling?

Suppose you have an idea for a structure of networking protocols to speed up transmission of digital messages, to identify the propagating frequencies, and to provide error correction. Suppose the idea is intended to work with hundreds of radio stations operating around the clock, over many frequencies. Imagine the frustration of trying to coordinate several hundred other amateurs in enough 24-hour tests to gain reliable performance measurements of this new radio networking idea! Now imagine trying to do it all over again after fixing all the problems and bugs which were revealed by the first try. This sounds as much fun as banging your head against a brick wall, and it is.

How can an idea like this ever get tested enough to get the kinks out? Computer simulation modeling is an answer. The authors develop and use computer simulation models of ALE HF radio digital networks to determine network performance under proposed networking protocols and networking features. The idea is to use simulation in lieu of over-the-air testing so that development of federal HF radio standards can proceed, although a feature will be included in the standard only after being proven in over-the-air testing. Over-the-air testing can be expensive, labor intensive, and time consuming for radio manufacturers, standards developers and users. Even so, the authors do not propose that simulation modeling replace over-the-air testing. The predictions given by a simulation are always approximations to expected network performance (see Note 1).

The costs of over-the-air testing can be reduced by simulation. Most of the major bugs should be revealed prior to expensive and time consuming over-the-air testing. Here is a simple example.

Suppose the stations in a very busy HF ALE digital network have long queues of digital messages waiting for transmission. Suppose every station handles each message immediately after transmitting the previous message. It's easy to imagine the situation where every station is attempting to transmit, but no station can receive since no station is listening for incom-

[1]Notes appear on page 18.

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ing messages. A simple solution is for each station to pause for a short period of time after completing a transmission and listen for incoming traffic. A computer simulation revealed this problem before it got on the air which would have added "EM pollution" to the already crowded HF spectrum. That's one problem that did not make it to an over-the-air test, and the solution was also verified by simulation.

Prerequisites

What does an amateur need in order to design a simulation program? The prerequisites are divided into two areas: equipment (hardware and software) and knowledge (skill and experience).

An AT class personal computer (PC), 286 Central Processor Unit (CPU), is basic. You don't need a workstation or a mainframe. A 386 or 486 (CPU) based PC is preferred. A floating point coprocessor is a plus if there is a great deal of floating point operations in the simulation program. The more random access memory (RAM) the better, but only if the simulation program uses the extra RAM.

The authors run simulations on a 486-50 MHz PC, with an EISA bus (32-bit), a SCSI 450-Mb hard drive with a fast caching controller card to speed up input and output (I/O), and 16-Mb of RAM. This is an expensive setup. The same simulations run at home on a 386-33 MHz PC with an 80-Mb hard drive and 4 Mb of RAM or on a 286-25 MHz PC (16-bit) with a 40-Mb hard drive and 2 Mb of RAM. They run more slowly at home, of course. As PC prices continue to fall, simulations will be more feasible for the amateur.

The amateur simulator needs to know a high level basic programming language and must have the language software (compiler/interpreter). High level means closer to English than machine language. BASIC, Pascal, FORTRAN, and C all qualify. Knowledge of data and file structures and efficient program design techniques are desirable, but are not absolutely necessary.

Knowledge of basic statistics is needed to analyze results and will help in deciding on the structure of the input data.

The most important prerequisite is a thorough knowledge of radio systems and the network operating procedures that are being modeled.

Simulation Characteristics

The authors have developed a discrete event simulation model for ALE HF Radio networks.[2] This will be the basis for describing some of the procedures for designing a network simulation.

Network simulations are generally discrete event simulations, which usually are stochastic as opposed to deterministic. Stochastic refers to randomly generated input data. Since the input data is randomly generated, the output data is also random in a sense. Specifically, there will be variability associated with the answer due to the randomness of the input. The relative extent of the variability gives an indication of how accurate the answer is. A deterministic simulation has no random components.

Network simulations are generally dynamic as opposed to static. A dynamic simulation evolves over time and may measure some performance parameter as a composite over the entire simulated time of the computer run or the measurement may be made periodically over the course of the simulation run. A static simulation does not evolve over time but generally simulates a situation with time fixed. Monte Carlo simulations, strictly speaking, are static simulations; however, some authors call anything with a random component "Monte Carlo."

Network simulations are generally discrete in that the discrete states or conditions of channels and stations change instantaneously in the model whether or not these changes are actually instantaneous. Engineers who do simulation call some components of the simulation finite state machines. A simple light switch can be described as a finite state machine. It has two states, on and off, and is always in one of those states. When the switch is moved from one state to the other, we assume that the state changes instantaneously. A simulation would assign a particular time for the event indicating the change of state of our light switch, and would not consider the real time it takes our hand to move the switch from one position to the other.

This abstraction is at the heart of simulations. A simulation does not include every detail, but abstracts some of the detail. In our light switch example, the details of the electricity moving in the wires and the light bulb element heating up are left completely out. The state of the light switch could also be extended to include the state of the light.

If the detail is important it should be included in a simulation. For example, the effects of the propagation condition of our radio channel on an individual bit in a packet of data could be left out if we are not considering such effects in our simulation. So we might consider the frequency either open or closed (like our light switch) and assume that the message is received as sent if the channel is open. However, if we are simulating the performance of a forward error correction scheme, then the effect on each individual bit must be considered.

A simulation model gives performance estimates of a system which cannot be described by a single equation or a small set of simple equations. Usually the situation is so complicated that it may not be easily described analytically or there are just too many interactions to consider.

Assumptions and Limitations of the Simulation Model

Simulation results are meaningless without a listing of the assumptions and limitations of the model. Assumptions and limitations indicate how far the model deviates from the real system. An example of a simplifying assumption is that a digital message will be received by another station, as transmitted, if the radio channel between them has an SNR of 5 dB or better. An example of a limitation may be that a simulation covers daylight hours only. Another limitation could be that the simulation is run only five times with different sets of randomly generated input data and that statistics are based on these five runs.

Assumptions and limitations serve to simplify the simulation model and its development. If, for instance, propagation time is not important to your simulation, then you might assume that the signal propagation time is effectively zero. This means that every time a message is transmitted you do not have to add the propagation time to the simulated time that the channel is busy with this transmission. An alternative is to assume that the propagation time is included in the time it takes for a message to be transmitted. Either way, computing time is reduced by this minor simplification. The size of the program is reduced, and the model is not so complex.

Assumptions are not intended to hide or ignore features of the real system which are important to the model. The above statements about propagation are moot if multipath delay is important. Assumptions need to be reasonable and reflect the system being simulated in its normal state.

A Random Number Generator

Pseudo-random number generators are important parts of discrete event simulations. They are used to provide a stream of numbers whose statistical characteristics are similar to a true sequence of random numbers which are uniformly distributed between zero and one (the probability of a specific number between zero and one occurring is the same as for all other numbers in the same range). The pseudo-random number generator actually generates a long sequence of integers which are normalized to produce the pseudo-random numbers. Pseudo-random numbers will be called random numbers in the rest of this article. We don't go into detail on random number generators but will provide one which should get the interested experimenter started.

$$Z_i = a Z_{i-1} \bmod m \qquad \text{Eq 1}$$

$$U_i = \frac{Z_i}{m} \qquad \text{Eq 2}$$

The integer sequence generator is defined recursively by Eq 1, where a is 630,360,016, m is 2,147,483,647 (or 2^{31}-1) and i is an integer between *1* and m*-1*. The *mod* operator is the usual modulo operator (m mod n is the integer remainder of integer m divided by integer n). The uniformly distributed random numbers, U_i, on the interval (0,1) are obtained from Eq 2. This generator is widely used in applications requiring uniformly distributed, random numbers.[3,4] The last integer of the sequence seeds the next integer and therefore is the seed for the next random number. This generator has a full period. That is, the integer stream covers every integer between 1 and m-1 before it starts over again. Any integer in this range can be used as a starting point (initial seed).

Methods of doing large integer arithmetic are generally available in most modern programming languages. If not, the same generator can still be used by using synthetic division and multiplication techniques (see Note 3).

An alternative is to use the random number generators available as a library routine within programming languages or within other software. Exercise caution in using unknown random number generators, they may not be as random as you think and should be tested. However, this type of generator is a good place to start and may provide what you need.

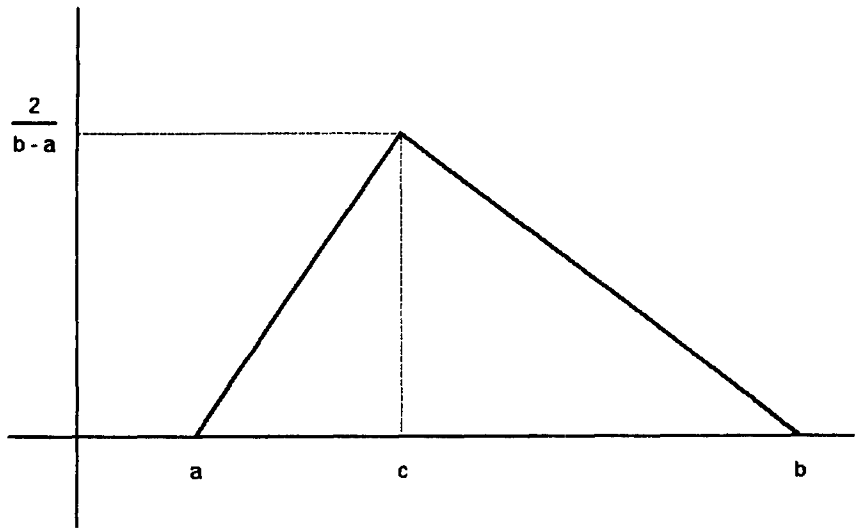

Fig 1—Triangular distribution.

Modeling the Input Data

To model input data it is necessary to statistically describe each of the input data parameters. A data parameter of the message traffic, for example, could be arrival time, length of message, originating station, destination station, etc. The statistical distribution, the mean (average), and the standard deviation of the input data parameters must be known. From this information the parameters which characterize the input can be generated using random number generators (see Note 3).

In the absence of knowledge about the input data parameter in question then the following procedure utilizing a triangular distribution can be used (see Note 3).

The example parameter is the random length, in minutes, of a voice conversation on a frequency. Let a be a reasonable estimate of the minimum length of a conversation, let b be a reasonable estimate of the maximum length of a conversation, and let c be a reasonable estimate of the most likely length of a conversation. Note that c is not the average but the mode. Now let the percentage of calls with length c be $2/(b\text{-}a)$. This value is used to normalize the distribution; that is, the probability of a conversation having length greater than or equal to a but less than or equal to b is 1. The triangular distribution of conversation lengths is shown in Fig 1. The average conversation length is $(a+b+c)/3$.

The length of the ith conversation is given by Eq 3, where U_i is the next uniformly generated random number in the random number string from Eq 2.

$$L_i = \begin{cases} \sqrt{cU_i} & 0 < U_i \le c \\ 1 - \sqrt{(1-c)(1-U_i)} & c < U_i < 1 \end{cases} \qquad \text{Eq 3}$$

If I can guess at maximums and minimums, then why can't I just guess at the mean (average) and then use that average to model each conversation? This means that every conversation in the simulation will have the same length, which could lead to simulation failure. For example, consider a simple factory welding process on an assembly line. The process takes 5 minutes, on average. Assume that the average interarrival times (time between arrivals) of the materials to be welded is 6 minutes. Having the average interarrival times greater than the average process time ensures that we will have no backups, doesn't it? A simulation of the process using the measured average time for the process to model each of the welding processes, and using the average interarrival times to model the arrival of each set of materials, will indicate that this is true.

The real system could incur backlogs hours long. What happens? By using only the averages to model processes or arrival times the statistical variation is ignored. There is possibly another mistake made in the simple example above. The simulator may not know enough about the input data and may not realize the variations possible in the welding process or the delivery process of material.

Notice one more thing in this example. By determining and using the statistical mean, variance, and distribution to model the input data in the welding process above, the simulation designer has abstracted all the variables mentioned into one statistical model. It might seem complicated to deal with the statistics, but it is simpler since the model for the welding process could encompass much more detail.

Simulation Events

Managing the occurrence of events is the most important part in developing the simulation program. The simulation events are either randomly created input data, they are events which are created by the simulation program, or they are housekeeping events. In an HF radio network simulation, example events are: message arrival, by which messages arrive at a particular radio station for transmission; message departure, which handles the completion of the transmission of a message to another radio station; call attempt, which models an attempt by a radio station to link with another radio station; and hourly housekeeping events, by which the program compiles statistics and resets variables and event counters after each hour of simulation time.

The event records are kept in a queue or in a sorted list of records. Each event record contains all necessary information about the event: event type, event time, radio station, call signs, frequency, etc. The simulation will read the next event to occur and call the appropriate subroutine of the simulation program to accomplish the event. Any new events created by this action are assigned an event record and the record is placed in its sorted position in the event queue.

The queue or stack of event records could be a linked list (linked by pointers) or an array (vector) of records. In either case the event records are kept in sorted order, by time of occurrence with the next event at the top of the stack or front of the list.

Sorting routines are available in programming software. For additional sorting methods, see Notes 5 and 6.

State Variables and State Changes

As the events occur, the associated simulation program subroutines change the state variables. The state variables, together, completely describe the simulated system at any point during the simulation.

An important state variable is the simulation clock. In a discrete event simulation the events generally drive the clock. The first action of an event subroutine is usually to advance the clock to the time of the event. Another important example is the state of each frequency or channel. In this case the state variable could indicate, 0-idle or 1-busy like the finite state machines mentioned previously (the light switch).

State variables are used in decision making by the simulation program. They are also handy debugging tools.

Queues and Stacks

Messages (actually records characterizing the messages) waiting at a station for transmission are normally held in a queue which is sorted by arrival time. When a radio station is finished receiving a message from another station or finished transmitting a message, the first message in the queue is taken out of the queue for transmission. The rest of the messages gain one position in the queue. It's like waiting in the checkout line at the grocery store. This is a first-in first-out queue.

Event records are held in a stack. This is the same as the queue mentioned above except that the events are held in the order of occurrence in the simulation and not in the order of arrival to the stack. There are two common ways to set up each of these structures. One is a linked list and the other is an array (or vector).

A linked list uses pointers and can take advantage of computer dynamic memory allocation. That is, the records which characterize the messages in a stack only take up memory in the computer when they actually exist. When the message transmission has been completed by the simulation, then the record is deleted and the allocated memory is available for other use.

Definitions and descriptions of these structures are given in the documentation of programming software. Methods of using these structures can be found in a text on data structures, see Notes 5 and 6, for instance.

Propagation Model

All the previous details are, in a sense, just bookkeeping. Now comes the interesting (hard) part—the radio channel (frequency) propagation model. How do you model the effect of the radio channel on the messages you transmit? The answer depends on what you are trying to measure. For instance, if you are measuring the delay of messages caused by the overhead for a particular networking protocol, the propagation model may be an on/off model: another finite state machine. That is, the channel is either open or closed. No attempt is made to model the degree of channel reliability. The authors are using the HF propagation prediction program IONCAP for this type of simulation model.

On the other hand if you are testing an error detection code requiring ARQ and retransmission, the effects of the channel on individual bits is important. The propagation model must then indicate whether an individual received bit is good or is in error. A channel mode that uses bit error statistics may be in order here.

If one does not have a propagation prediction program in hand, such as IONCAP, then a good way to model a channel is by using your own experience or statistics that you have compiled. For example, you probably know which frequencies propagate at night or in daylight hours only. You are probably familiar with the effects of the sunspot cycle and what propagation performance you can reasonably expect with a certain sunspot number. And you probably have some experience of seasonal variation. You can include this information in your propagation prediction model.

A simulation program does not necessarily need a complicated mathematical model for simulation of radio channel propagation. A simulation designer recently assumed that five out of the ten available HF frequencies were open at any one time.[7] However, if you report your results to others, you should detail the propagation model. Limitations, assumptions, or restrictions that are placed on the propagation model are limitations to the simulation.

Examples

The following examples are generalizations of simulation events in the simulation the authors use to model HF ALE Radio networks (see Note 2). The program was developed using top-down design methods; however, general flow chart model diagrams can describe the logical flow.

A message arrival event is detailed in Fig 2. Most events begin by advancing the simulation clock. The first decision is a determination of the state of the transmitting radio station. If the station is busy (transmitting or receiving another message, for example), the newly arrived message is placed in that station's message queue. If the station is idle then the list of available channels (frequencies) are sorted by the recorded (or predicted) channel quality. An event record is created for a call event on the first channel in the sorted list. The event is placed in the proper position in the event queue. Program control returns to the main program which pops the next event from the top of the event queue and calls the appropriate subroutine to accomplish the new event.

A call event is illustrated in Fig 3. A call is an attempt by a radio station to link with another station over one of the available frequencies (channel *i*). As before, the call event begins by advancing the simulation clock. The state of both channel *i* and the receiving station is determined. If both the receiving station and the channel are idle (not busy) and the two-way propagation path between the stations is open, then the two stations link (message transmission begins). The state of both stations and channel *i* is changed to busy. A departure event is created and placed in the event queue. The departure event occurs when the message transmission is finished and accomplishes the return to the idle state for channel *i* and for the two stations.

If the radio station is busy or if channel *i* is busy or closed (propagation not possible), the channel state is recorded (busy or closed). If channel *i* is the last channel in the sorted list of channels then the message is placed on the end of the message queue to await a retransmission attempt. Otherwise a new call event is created for the next channel, *i*+1, and the event is placed in the message queue. Control returns to the main program.

Many details have been left out of these descriptions for the sake of brevity and clarity. For instance, the propagation states of the channel in the call event procedure are determined by the implementation of the propagation model. Another omitted detail is how a newly created event is placed in sorted position in the event queue.

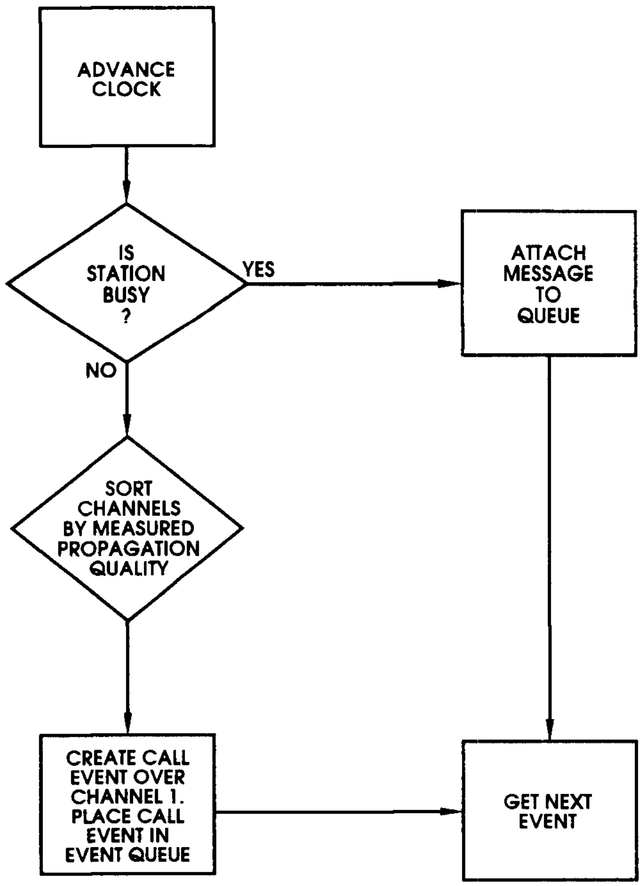

Fig 2—Arrival event flow chart.

Example Results

The authors have recently completed a simulation study of the effects of sounding on an ALE HF radio network (see Note 2). Sounding is the periodic broadcast of self identification information which may be monitored, recorded, and evaluated by other radio stations. The state of the one-way HF propagation channels may then be determined. The channel quality information may then be used by radio stations in the network to select a channel that is most likely to provide a link. However, the overhead of sounding may be detrimental to network performance: A sounding station cannot receive incoming traffic and a channel being sounded upon is unusable by other stations.

The conclusion of the simulation study is that sounding is generally detrimental to network performance. Any efficiency due to the additional channel information obtained by sounding is outweighed by the interference to other traffic caused by the periodic sounding.

An exception to our major conclusion is that in poor propagation conditions, with low traffic rates, sounding may enhance some aspects of network performance. This is illustrated in Fig 4, which shows hourly call success rates for a twenty-four-hour period. Call success rates are shown: (1) with each station sounding, on each available frequency, once every hour, and (2) with stations not sounding. Each data point represents the average call-success by hour for twenty-five runs of the simulation. The success rate is better without sounds during the first 13 hours of the period. In the second part of the twenty-four-hour period, during more adverse propagation conditions, sounding enhances the call success rate. This is due to the identification, by sounding, of the few, 2 or 3, propagating HF channels. Therefore, the radio stations are not wasting time trying to link over the channels that are not usable.

These results and conclusions from a simulation study show some of the possibilities of simulation of radio networks.

Conclusion

This article has presented simulation modeling as a tool for experimentation and investigation. This is intended to serve as a starting point for

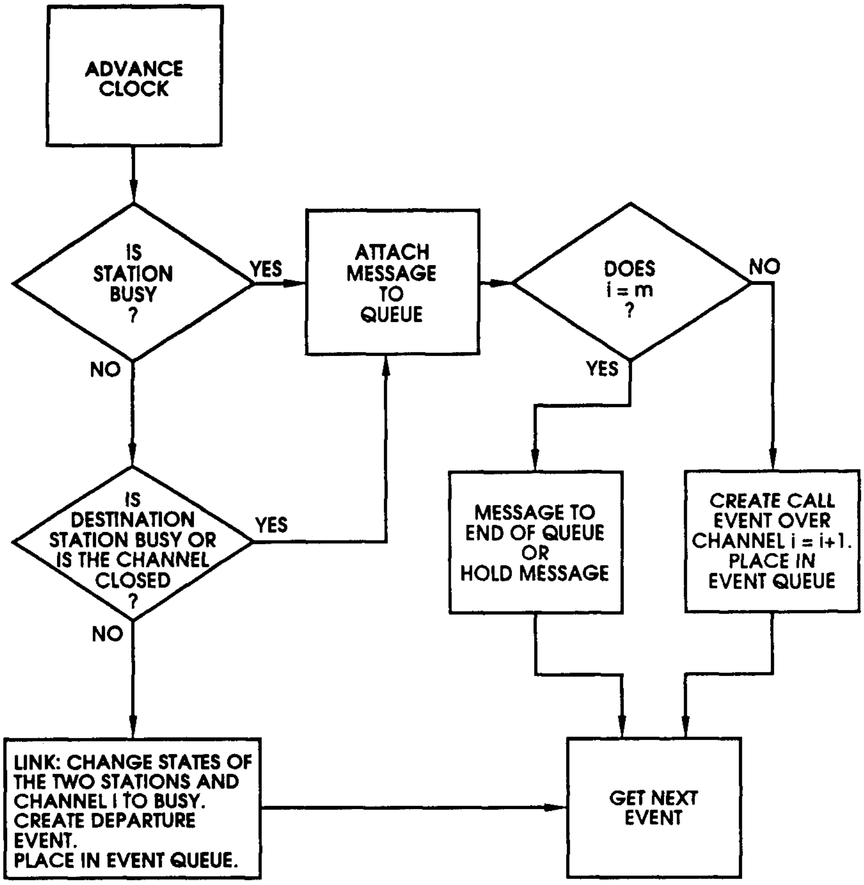

Fig 3—Call event flow chart.

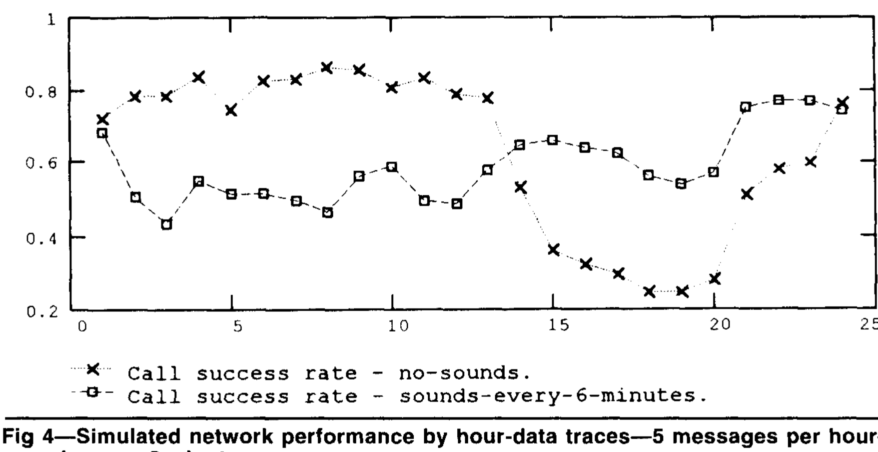

Fig 4—Simulated network performance by hour-data traces—5 messages per hour—sounds every 6 minutes.

the prospective simulation designer. Further information and concepts on simulation modeling are available by consulting the books in the Notes.

The book by Law and Kelton (Note 3) is a good reference and text for general simulation. It is self contained and well referenced. The book gives algorithms for many example procedures, including input data generation and random number generation. It requires some knowledge of statistics and probability but contains a comprehensive review chapter of these subjects. Dr. Law teaches a short course on simulation modeling every few months at George Washington University. The book by Bratley, Fox, and Schrage (Note 4) is a mathematics text that requires knowledge of statistics and probability. The Horowitz and Sahni books (Notes 5 and 6) contain detailed explanations, techniques and algorithms on linked lists, searching, sorting and manipulation of data with pointers to data structures.

Acknowledgments

This work was supported by the National Communications System (NCS) and the National Telecommunications and Information Administration/Institute for Telecommunication Sciences (NTIA/ITS).

Notes

[1]Wickwire, K., "The Status and Future of High Frequency Digital Communication, Part III: Simulating the Performance of HF Digital Networks," *QEX*, No. 126, August 1992, pp 12-17.

[2]Sutherland, D. (1992), NTIA Report 93-291, NCS Technical Information Bulletin 92-21, *Simulated effects of sounding on automatic link establishment HF radio network performance.*

[3]Law, A.M., and Kelton, W.D., (1991) *Simulation Modeling and Analysis*, 2nd Ed, (McGraw Hill, NY).

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[5]Horowitz, E., and Sahni, S. (1987), *Fundamentals of Data Structures in Pascal* (Computer Science Press, Rockville, Maryland).

[6]Horowitz, E., and Sahni, S. (1983), *Fundamentals of Data Structures* (Computer Science Press, Rockville, Maryland).

[7]Johnson, E. (1992), New Mexico State University, private correspondence. □□

A Dip Meter with Digital Display

There's nothing like a dip meter for tuning antennas. You can stop squinting at that faded old dial after building this meter with digital display!

by Larry Cicchinelli, K3PTO

One of the main difficulties I have had over the years in building my dipole antennas is that of cutting them to the correct length. This article describes a dip meter with a 3-digit digital display which I designed and built to aid me in solving this problem.

I have a grid dip meter which I built from a kit and which, for quite a while, worked quite well. One of its shortcomings was that of alignment. Even though the printed frequency chart was relatively accurate, it still needed calibrating for each band. Since I was trying to decide on my next construction project anyway, I decided to build a solid-state replacement for it.

The device is a dip meter with a 3-digit frequency display. The analog portion of the circuit consists of an FET oscillator, voltage doubling detector, dc offset circuit and amplifier. The digital portion of the circuit consists of a high-impedance buffer, prescaler, counter, display driver, LED display and control circuit.

Circuit Description

The dip meter has four distinct functional blocks each of which I will describe in detail.

The RF oscillator is a standard Colpitts using a common junction FET, Q1, as the active element. Its range is about 1.7 to 45 MHz. A 200-pF tuning capacitor was selected because I wanted a 2:1 tuning range. In order to get a 2:1 frequency range the capacitor must have a 4:1 range. The sum of the minimum capacitance of the variable, the capacitors across the inductor, and the strays must therefore be in the order of about 70 pf. The values of L1 were determined experimentally by winding the coils and observing the lower and upper frequency values. The coils I made cover the following frequency ranges: 1.7-3.1 MHz, 2.8-5.9 MHz, 5.6-11.9 MHz, 9.7-20.7 MHz, and 19-45 MHz. Because I experimentally determined the coil sizes and am using old coil forms of unknown parentage, I've not provided winding details. Experiment!

I mounted the tapped capacitors inside the coil forms so that their values could be different for each band if required. The frequency spread of the lowest band is less than 2:1 because the tapped capacitor values are larger than for the other bands. The frequency spread of the highest band is greater than 2:1 because its capacitors are smaller.

The analog display circuit begins with a voltage doubler for the detector in order to get higher sensitivity. It drives a dc offset circuit, U1A. R1 is used to insert a variable offset which will get subtracted from the detector voltage. This allows the variable gain stage, U1B, to be more sensitive to variations in the detector output voltage. Q10 was added to U1C in order to allow its output to get closer to ground. The resistor in series with the meter should be chosen so that the current is limited to a safe value. For example, if a 1-mA meter is used, the resistor should be 8.2 k.

The prescaler begins with a high-impedance buffer and amplifier, Q2 and Q3. If you are going to use the meter for the entire frequency range described, care should be taken in the layout of both the oscillator and this buffer/amplifier circuit. The digital portion of the prescaler is a divide-by-100 circuit consisting of 2 divide-by-10 devices, U2 and U3. The devices I used were selected because they were available. Any other similar devices may be used as long as the reset circuit is compatible. Q5 is a level translator, shifting the 5-V signal to 9 V.

The first part of the digital display block is the oscillator circuit of U1C, which is used to create the gate time for the frequency counter. R3 adjusts the oscillator to a frequency of 500 Hz, yielding a 1-ms gate. The best way to set this frequency is to listen for the dip meter output on a communications receiver and adjust R3 until the display agrees with the receiver. Once this calibration has been made for one of the bands, all the bands are calibrated.

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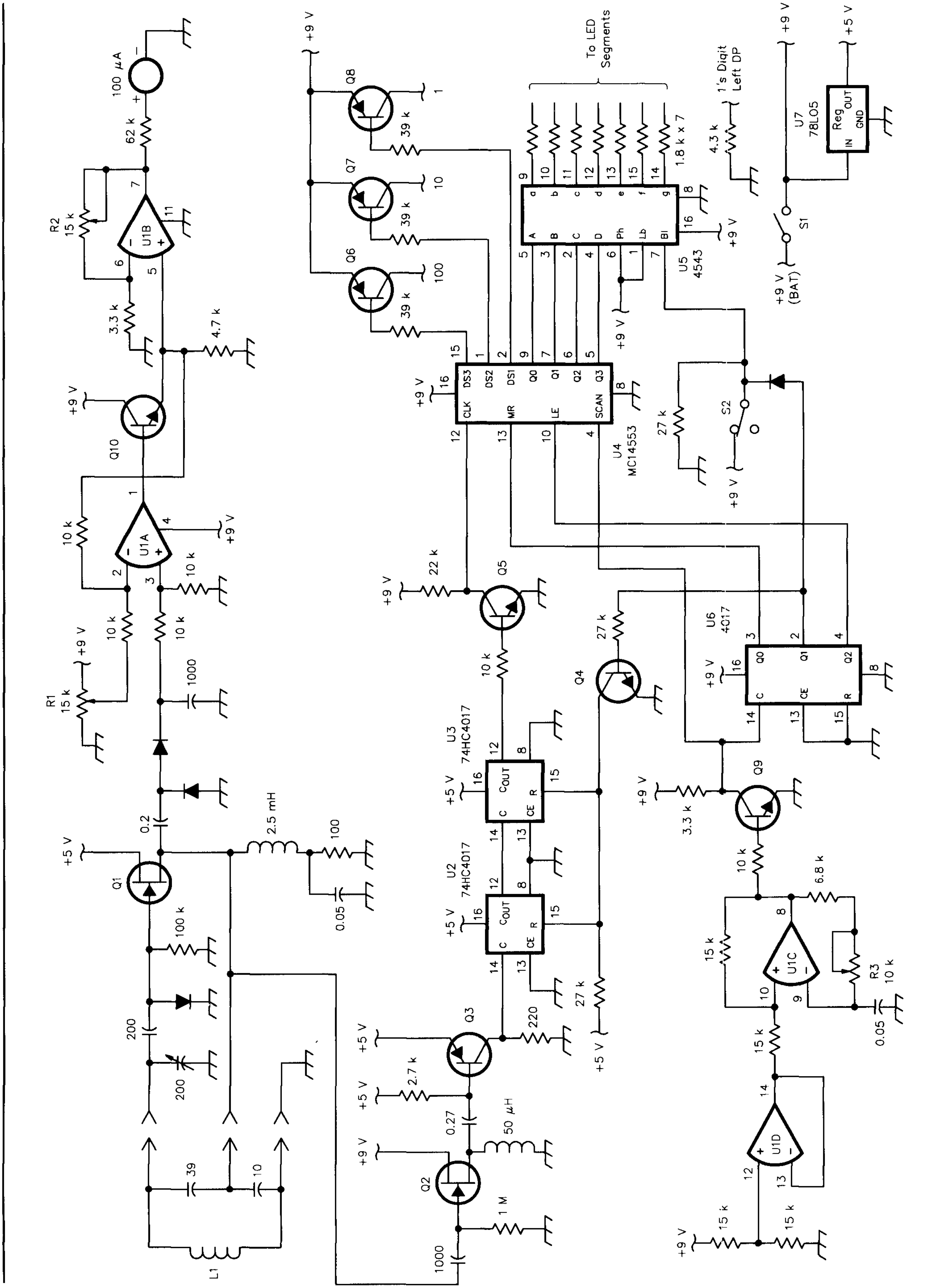
Lansdale, PA 19446



To LED Segments
1's Digit Left DP
100 µA
62 k
R2 15 k
U1B
3.3 k
4.7 k
Q10
10 k
U1A
R1 15 k
1000
0.2
Q1
2.5 mH
100
0.05
100 k
200
39
10
L1
Q8
Q7
Q6
39 k
Q5
22 k
U3 74HC4017
U2 74HC4017
Q3
220
27 k
2.7 k
0.27
50 µH
Q2
1 M
U4 MC14553
U5 4543
1.8 k x 7
4.3 k
U7 78L05
S1
+9 V (BAT)
S2
Q4
U6 4017
Q9
3.3 k
6.8 k
U1C
R3 10 k
15 k
U1D
+9 V
+5 V

U1D is used to give a low-impedance voltage reference for U1C. I added Q9 because the output of the oscillator has a small "glitch" which can cause the counter to trigger incorrectly. I decided to use this type of oscillator, as compared to a crystal oscillator, mainly for simplicity. This circuit is fairly stable, easy to adjust, and has a low parts count. Since the frequency is displayed with 100-kHz resolution, I felt that the "compromises" were well worthwhile.

The digital system controller, U6, is a divide-by-10 counter which has 10 decoded outputs, each of which will go high for one period of the input clock. The Q0 output is used to reset the frequency counter, U4. The Q1 output is used to enable the prescaler and disable the display, and the Q2 output latches the count value into the frequency counter. Since the prescaler can only count while Q1 is high, it will be enabled for only 1 ms. Normally a 1-ms gate will yield 1-kHz resolution. Since the circuit uses a divide-by-100 prescaler, the resolution becomes 100 kHz.

The frequency counter, U4, is a 3-digit counter with multiplexed BCD outputs. The clock input is driven from the prescaler, hence it is the RF oscillator frequency divided by 100. This signal is present for only 1 ms out of every 10 ms. The digit scanning is controlled by the 500-Hz oscillator of U1C. U5 is a BCD-to-seven-segment decoder/driver. Its outputs are connected to each of the three common-anode, seven-segment displays in parallel. Only the currently active digit will be turned on by the digit strobe outputs of U5, via Q6, Q7 and Q8. The diode connected to the blanking input of U5 disables the display while U4 is counting.

U7 is a 5-V regulator. I could probably run the entire circuit using 5 V, but I wanted to use a single battery for the circuit and I wanted to drive the LEDs from 9 volts.

S2 is used to enable the LED displays once the unit has been adjusted for a dip.

The circuit draws about 20 mA with the LEDs off and up to 35 mA with the LEDs on. The current will vary depending on the frequency displayed.

Many of the resistor values were used because they were available and could just as easily be other values. For instance, the four 10-k resistors around U1A could have been 15 k. For the op-amp circuits, the resistor ratios are what is most important. The resistor at the collector of Q3 should not be varied. The 0.27-µf capacitors I used are monolithic and have good high-frequency characteristics over the range of the meter.

Most of the parts can be purchased from Digi-Key.[1] The only IC which they do not have is the 4543, which I purchased from Hosfelt Electronics.[2] The 74HC4017 may be substituted with a 74HCT4017. The circuit is built on a 4-inch-square perf board with places for up to 12 ICs and is housed in a 7×5×3-inch mini box.

Operation

I finally did use the dip meter to help me adjust the length of my new antenna. The dipole I built is a 40- and 20-meter parallel dipole made with good quality TV twin lead. I originally cut both sections too long. Using the meter I was able to adjust them to the proper length before raising the antenna to its final height. The only difficulty I had was that the LEDs were not bright enough to be viewed in direct sunlight. I had to adjust the frequency tuning for a dip and then go inside to view the frequency. *A word of caution*: Do not measure the antenna frequency at the end of the transmission line. I found that I could only get accurate results with a single loop of wire right at the feed point of the antenna.

I start with the gain, R2, set fully clockwise for maximum sensitivity. With R1 set fully counter clockwise, +9 V, the output of the offset circuit, U1A and Q10, is at ground. As R1 is rotated, the voltage on the arm approaches, and then becomes less than, the detector output. At this point the meter will start to deflect upward. I usually adjust R1 so that the meter reads about center scale. This allows for variations in the output level of the RF oscillator. As L1 is brought closer to the circuit being tested, the meter will deflect downward due to energy being absorbed by the circuit. For best results you should use the least amount of coupling possible to the circuit being tested. If you overcouple, the oscillator frequency will be pulled.

The RF oscillator and the dc offset circuit are variations of a dip meter described in the 1986 ARRL *Handbook*.

Notes

[1]Digi-Key, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel 800 344-4539.

[2]Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952, tel 800 524-6464. ☐☐

Fig 1—Schematic diagram of the dip meter.
Q1,Q2—MPF102 JFET transistor
Q3,Q6,Q7,Q8—2N3906 PNP transistor
Q4,Q5,Q9,Q10—2N3563 (or any general purpose NPN)
U1—LM324
U2,U3—74HC4017
U4—MC14553
U5—4543
U6—4017
U7—78L05
All diodes are 1N914 or similar
All resistors are 1/4 W, 5%

Troposcatter—DX CQ

Ever wondered what troposcatter propagation can do? The author's computer simulations provide some intriguing answers.

By Merle C. Rummel, W9LCE

Troposcatter communication is based on the scattering of a radio beam by the various particles in the air such as water vapor, dust and air molecules. This permits reception of radio communications off the straight line-of-sight beam. Normally, this is useful mostly in the VHF frequencies, but it is progressively being used in the microwave frequencies.

This article is based on a computer simulation using all known required data to determine a maximum distance of reception of microwave troposcatter signals. The base station was selected to be approximately equivalent to an EME station, ie, the highest possible transmitter power, the least line loss, the best available antenna on both ends, the lowest noise front end, preamp—mounted at the antenna—and SSB bandwidth (2200 Hz) voice signals. Since most troposcatter is at VHF and above, and since the author has been a Technician-class licensee for some 40 years, the frequencies chosen were those amateur bands available with a Technician license: 6 meters, 2 meters, 222 MHz, 432 MHz, 900 MHz, 1.3 GHz, 2.4 GHz, 3.3 GHz, 5.6 GHz, 10.3 GHz and 24 GHz, with particular interest in 432 and above. The paper assumes similar stations on both ends of the communication path. The computer program written originally came from Bob Atkins' article in *Communications Quarterly* (1991), but was translated to True BASIC, then revised several times to achieve the accuracy and detail needed to obtain the results desired for this study.

I've communicated using this mode on 6 meters and have some familiarity with the quirks experienced in its use. But I am sure there are readers who have much wider experience and could profitably add to the accuracy of these results. All known effects were included which would modify and limit the transmission (see Appendix).

Troposcatter signals in the UHF and microwave bands have a similar range if they have a similar transmitter power. While there is increasing loss at higher frequencies, there is increasing gain in the same capture area of the antenna. The two factors balance each other. For the constant 100-W transmitter, Table 1, note the increasing ERP, especially of the cylindrical parabola with its constant 64-square-foot area. Since it is assumed that the receiving antenna has the same area, the increased gain compensates for the increased "loss" due to frequency.

In troposcatter transmission the most power is found in direct line-of-sight from the original beam. However, as a result of collisions of the beam with molecular substances in the air, some of the energy is scattered. The path with the least angle of scatter, or bend, between the transmitting station and the receiving station will give the strongest signal, and thus the greatest range. Transmission distance versus take-off angle for a given power is shown in Table 2. According to this table, a station on a high hill, with only lower hills on its horizon, will give very distant contacts. The higher the station has to elevate its beam to cross its horizon, the shorter its range will extend. The range drops off rapidly. Note in this table that the ERP (output of the antenna) is held constant, and the mileages are, as a result, reduced with higher frequency. (I will not guarantee the accuracy of the –1° mileage; it worked out this way on the computer program!)

Table 3 shows the effect of air pressure on the range of the troposcatter signal. The lower the pressure, the greater the mileage. This is probably because the increased pressure results in more obstructions in the same space, hence reducing the power in any one path. As shown, the change in path length for a given pressure change is about constant.

In Table 4, several patterns can be observed that are present on all further charts. There is a peak range, normally close to freezing at low humidity (with

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Boston, IN 47324

Tables begin on page 25.

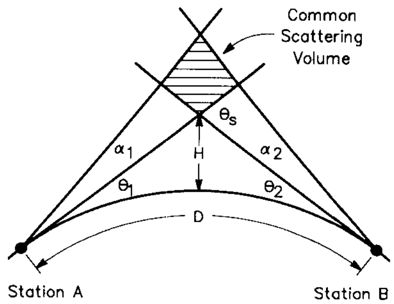

Fig 1—Troposcatter profile.

more range at lower humidity and less range at higher). Note, though, the slight dropoff in ranges below freezing.

Compare Table 9 with Tables 10 and 11 and note how the range changes as air pressure increases and decreases. Higher air pressure means less range; lower air pressure means increased range.

Comparing Table 12 to Table 9 shows the rapid decrease in range with a slight increase in the elevation of the beam. (The elevation value was chosen for author's site.)

In Tables 16-20, note the increasing antenna gain due to decreasing wavelength and constant reflector size (capture area).

These tables also show how high heat and humidity drastically reduce range at high frequencies. In Table 20, note the available distances during dry, cool weather, even at low power.

Comparison of Table 21 with Tables 22 and 23 shows very similar results to those at 432 MHz for varying air pressure. Because of the sharp vertical beam pattern, it is possible that no Earth noise will intrude on the received signal, and the 310-K signal loss might be relaxed. I could not find information to substantiate this, and we have not begun testing this data at our club, so 310 K was retained.

In Table 24, increased elevation of the antenna beam causes considerably reduced ranges from those of Table 21. This is consistent with our findings at 432 MHz.

Note the 10- to 18-mile increase in the ranges of Table 26 compared to Table 25 with the cylindrical parabolic antenna. This is due to the 3-dB higher gain of the dish in Table 26.

The large antenna of Table 27 gives a respectable range even with a low-power Gunn diode transmitter.

Table 29 shows a cold weather sport. Find something better for those summer days! I have been looking for some high-power diodes that would work as push-push frequency/power doublers for this band—anyone know of a source? If I find one, I'll stick one of our 10-W, 10-GHz power transistors on it!

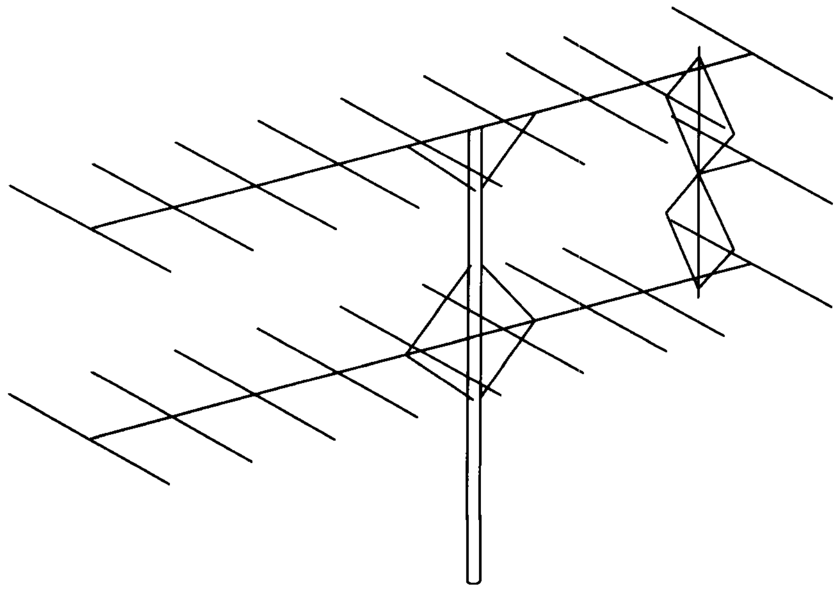

Fig 2—Twin quad antenna.

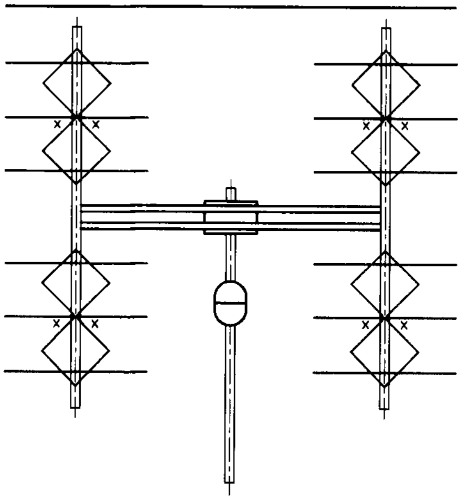

Fig 3—Twin quad antenna.

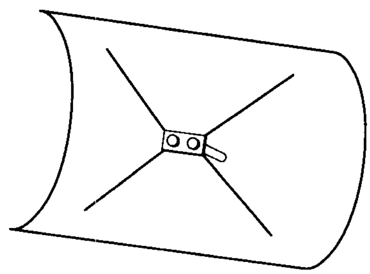

Fig 4—Cylindrical parabolic antenna.

Receiver noise figure values have a significant relationship to troposcatter distances, as shown in Table 30. Note that for these arbitrary values for the 432-MHz band, the difference of 1 dB in the noise figure is equal to 10 miles. Similarly, 1 dB is about 6 or 7 miles for the 10-GHz band. This is a good reason for trying to obtain the lowest noise figure front ends possible on the preamp.

Conclusions

There is real possibility of a communication range of about 300 miles through all the microwave bands using troposcatter communications. This is especially true if high-quality communication equipment is used (moon-bounce equivalent). Most of these stations will need to be home brew, with the best of construction techniques and special attention paid to low-noise reception. The worst problems are beyond the control of the amateur, but are addressable: the path needs to be as near to 0° above the horizon as possible (to get the nearest direct angle and, hence, most scatter power); the atmosphere should have the least amount of water moisture in it (low temperature—around freezing, and low percent humidity). Consistent microwave DX communication is there during all but the hot, humid summer days.

About the Author

The author is a college professor of physics and electronics. He has been chief engineer for WKOI, Ch 43 TV, Richmond/Cincinnati. This was a presentation made to the Midwest VHF-UHF Society meeting, February 26, 1993.

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Bob Atkins, KA1GT, "Radio Propagation by Tropospheric Scattering," *Communications Quarterly*, Winter 1991, p 119.

DJ9HO, et al, *The UHF Compendium Parts 1, 2, 3 & 4* (now if only they would bring out a microwave Part 5).

Appendix

To meet the stated limits, several choices were made:

• The transmitters were determined to have the maximum possible power on each band. On the VHF bands, this power was to be a full kilowatt output. [This includes the novel SSB method, used in Europe, SBFM, or the clipping of the amplitude component of the SSB signal so that only the frequency change is transmitted; there is no difference required in the reception of this signal and it is not obvious to the receiving station that this transmission method is used (so they say—the author is adapting a transmitter to this method to try it). The signal thus would be able to use Class C, or better, amplifiers, not be concerned with linearity, and so have high efficiency—75% or better, close to a 3-db gain in power.] The use of a 2C39 amplifier (water cooled) gives good power on 1.3 and 2.4 GHz (see the ARRL *Handbook* and *QST*), a 1977 article by W9ZIH ("Video modulated Four-Tube Amplifier for 1270-MHz Television," June 1977, *Ham Radio*) describes a 4-tube 2C39 cavity amp at 1.3 GHz that would result in a full kW). Some 10-watt transistors are available for 10 GHz, but it is recognized that most microwave depends on the Gunn diode, and 0.1 watt is pretty good output with these. The transmitters should be mounted at the antenna, or have low-loss leads (open-wire twin lead at VHF frequencies).

• The antennas were limited to two types: for the low VHF/UHF frequencies, the dual-quad Yagi (from Weiner, *UHF Compendium*) was chosen, using 20-foot pipe for the booms for 50 MHz and 10-foot pipe for 144, 222, and 432 MHz. For the microwave frequencies, a cylindrical parabolic reflector (re: 1961-66 *73* Magazine, "Big Sail Antenna" and related articles) was chosen, size 8'x 8'. This allowed high gain (3-db less than the full parabolic) from a sharp vertical pattern, with reasonable beam width (fan pattern) in the horizontal, to permit CQ-type contacts. (Normal amateur communications in the microwave bands—without all the problems of prearranged time and frequency and dish alignment.) Since reflectors are sensitive to $\lambda/10$ irregularities, the use of the reflector at 10 GHz means that the surface must be accurate on the parabolic curve to 3 mm (1/8 inch) (1 mm at 24 GHz). This means a solid sheet of metal (aluminum) fixed to a rigid set of ribs. This is nearly impossible for a full dish; even a 2-meter satellite dish may not meet this requirement. (These are built for 3.5-GHz. Don't use a perforated dish above 3.5 GHz!) So the simple curve of the cylinder antenna is almost demanded—and gives the benefit of a fan type horizontal beam. (I wonder how aluminum covered insulation board would work? 2 sheets at 4x8?) The feed (horns) for transmit and receive are mounted on the same base, but are separate units. To avoid transmitted power reflecting into the receiving system, a wedge isolator was assumed to be mounted on the reflector directly below the transmitter horn/receiver horn junction (or, with more loss, the transmitter feed can be a pipe coming up from the reflector between the two systems). This causes a slight loss of directed power to the distance but permits full-duplex operation. The antenna was considered to be mounted on a stable base, at a height of 30 feet (a tripod made of 3 old telephone poles! for the cylindrical parabolic). Wish I didn't have to move!

• The pre-amps: GaAsFETs can give close to 30X (0.5 dB) front ends at the UHF frequencies, and the new HEMTs may do 1.0 dB µF at frequencies beyond those in this study. 0.5 dB was chosen as the input noise figure below 1 GHz and 1.0 dB for those bands above. It must be acknowledged that 1.0 dB is a reach, with typical values now nearer 5.0 dB. Preamps are mounted at the antenna and a separate coax feed line run for receive.

•Formulas

index of refraction:

$$ns = \frac{77.6m}{t} + \frac{3.73 \times 10^5 e}{t^2}$$

based on water vapor pressure

$$e = h^{(9.4051 - (2353/t))}$$

with: m = pressure (millibars)
h = percent humidity
t = temperature (K)

refractive index loss $\quad 1n = 0.2(ns - 310)$

(310 is the temperature of the Earth in Kelvins)

scattering angle (degrees) $\quad sa = 0.005d_{miles} + 2t$ (t in K)

3-db beamwidth

$$bw = \frac{27000}{g/10}$$ (g = ant gain)

coupling loss

$$l = 2.5 + \frac{1.8sa}{bw^2} - 0.063\left(\frac{sa}{bw^4}\right)^2$$

(slight distortion of results on cylindrical reflector)

take-off angle

$$toa = \frac{h_2 - h_1}{d} - \frac{57.3d}{2(8475)}$$

(all measurements are in meters)
h2 = height of hills
h1 = height of ant
d = distance

gain of long Yagi $\quad gi = 10.1 + 6 \log b1/\lambda$
b1 = boom length (feet)

(book says 10.1 should be 12.1, I couldn't do it!)

stacked Yagi $\quad g = gi + 2.7\ 1n\#$
\# = number of booms

parabolic dish $g = 10\log\left[0.55\left(\frac{\pi d^2}{\lambda}\right)\right]$

d = diameter

cylindrical parabola $g = 10\log\left[0.75\left(\frac{\pi d^2}{\lambda}\right)\right]$

d = distance

temperature (K) $T = \frac{5(t-32)}{9} + 273.15$

t = temp (°F)

air pressure (millibars) $pr = 33.86p_i$

p_i = barometric pressure (inches)

power level (dBW) $p = 10\log W$

W in watts

receiver bandwidth (dBW) $bd = (10\log bw) - 20$

bw = 2200 Hz (SSB)

receiver input noise level (dBW) $pr = NF + bd - g - 184$

NF is noise figure in dB

Transmitter (erp) $erp = p + g - \text{line loss}$

Line loss in dB

Path Loss $lp = erp - pr$

Total Path Loss

$tpl = 53.45 + 20\log d_{mi} + 30\log f(MHz) + 1 + 10\ tsa + 1n$

[freespace path loss (base) ------------- 32.45 dB]
[scatter loss (base) ------------------------ 21 dB]
------------]
53.45 dB]

for distance: $lp = tpl(loop)$

Table 1—Troposcatter—Distance Vs Frequency

MHz	Transmitter *100 W*	*ERP*	
50.5	429 miles	33	Twin-
144	366	36	Quad
222	333	37	Stacked
432	332	42	Yagi
900	262	43	
1300	273	46	
2300	289	51	8'x8'
3300	296	54	
5600	303	58	Cylindrical
10000	303	63	Parabola
24000	291	71	
47000	331	77	

Table 2—Troposcatter—angle to horizontal

distance in miles

ERP –45 dBW degree

MHz	–1°	0°	1°	2°	3°	4°
50.5	1115	840	670	460	268	111
144	964	742	528	329	158	44
222	902	681	471	278	118	27
432	807	591	386	204	69	19
900	703	491	295	131	33	4
1300	653	444	254	102	21	3
2300	576	373	193	63	10	1
3300	528	328	157	44	6	0.6
5600	458	266	110	24	3	0.6
10000	384	202	61	11	1	0.6
24000	279	119	28	4	0.6	0.6
47000	1204	69	1	0.6	0.6	0.6

Table 3—Troposcatter-Air Pressure Effect

inches Hg	miles 432 MHz	miles 10 GHz	32° Temp 45% Humidity
28.0	450	243	
29.2	426	225	
30.5	402	205	

Table 4—6-Meter Standard Configuration: 1-kW input

50.5 MHz	Ant—20' Twin Quad Yagi	
41.9 dBW	Xmtr Power	800 W
–182.9 dBW	Rcvr NF	0.5 dB
224.8 dB		
	Ant Gain	12.8 dB
Elevation 1°	Beamwidth	37°

T –10°	10°	32°	50°	70°	95°	P 29.2"
498 mi	521 mi	541 mi	551 mi	550 mi	525 mi	h 20%
495	516	528	527	505	431	45%
492	508	510	491	438	298	70%
490	502	497	468	394	218	90%

Standard configuration: 29.2-inches mercury; 1° elevation. Range for 0 dB S/N is shown. SSB readability is frequently considered to be +3 dB S/N which varies by maximum distance but is normally 20 to 30 miles less than the values shown.

Table 5—6-Meter Standard Configuration: 250 W

50.5 MHz	Ant—20' Twin Quad Yagi	
36.8 dBW	Xmtr Power	250 W
–182.9 dBW	Rcvr NF	0.5 dB
219.8 dB		
	Ant Gain	12.8 dB
Elevation 1°	Beamwidth	37°

T –10°	10°	32°	50°	70°	95°	P 29.2"
446 mi	470 mi	490 mi	499 mi	498 mi	473 mi	h 20%
444	465	477	475	453	381	40%
441	456	458	440	438	353	70%
438	451	446	417	345	177	90%

Table 6—6-Meter Standard Configuration: 20 W

50.5 MHz	Ant—20' Twin Quad Yagi	
25.8 dBW	Xmtr Power	20 W
−182.9 dBW	Rcvr NF	0.5 dB
208.7 dB		
	Ant Gain	12.8 dB
Elevation 1°	Beamwidth	37°

T −10°	10°	32°	50°	70°	95°	P 29.2"
340 mi	361 mi	380 mi	389 mi	388 mi	365 mi	h 20%
338	356	368	366	346	279	40%
334	349	351	333	285	164	70%
332	344	339	312	246	101	90%

Table 7—2-Meter Standard Configuration: dual twin, 10' booms; 250 W

144 MHz	Ant—10' Dual Twin Quad Yagi	
40.5 dBW	Xmtr Power	250 W
−186.6 dBW	Rcvr NF	0.5 dB
227.0 dB		
	Ant Gain	16.5 dB
Elevation 1°	Beamwidth	24°

T −10°	10°	32°	50°	70°	95°	P 29.2"
382 mi	405 mi	424 mi	433 mi	432 mi	408 mi	h 20%
380	400	411	410	389	320	40%
376	391	393	376	326	199	70%
372	386	381	354	285	130	90%

Table 8—222-MHz Standard Configuration: 10' dual twin quad; 250 W

222 MHz	Ant—10' Dual Twin Quad Yagi	
41.1 dBW	Xmtr Power	220 W
−187.7 dBW	Rcvr NF	0.5 dB
228.8 dB		
	Ant Gain	17.6 dB
Elevation 1°	Beamwidth	21°

T −10°	10°	32°	50°	70°	95°	P 29.2"
345 mi	367 mi	386 mi	395 mi	394 mi	370 mi	h 20%
343	362	374	372	351	284	40%
340	354	356	339	290	168	70%
337	349	344	317	251	105	90%

Table 9—432-MHz Standard Configuration: 10' quad twin quad Yagi

432 MHz	Ant—10' Quad Twin Quad Yagi	
51.1 dBW	Xmtr Power	800 W
−192.1 dBW	Rcvr NF	0.5 dB
243.2 dB		
	Ant Gain	22 dB
Elevation 1°	Beamwidth	13°

T −10°	10°	32°	50°	70°	95°	P 29.2"
397 mi	420 mi	439 mi	448 mi	447 mi	423 mi	h 20%
395	415	426	425	404	334	40%
391	406	408	391	340	212	70%
389	401	396	368	300	141	90%

Table 10—432-MHz Configured for 28 in. Hg, air pressure

432 MHz	Ant—10' Quad Twin Quad Yagi	
51.1 dBW	Xmtr Power	800 W
−192.1 dBW	Rcvr NF	0.5 dB
243.2 dB		
	Ant Gain	22 dB
Elevation 1°	Beamwidth	13°

T −10°	10°	32°	50°	70°	95°	P 28"
421 mi	443 mi	461 mi	470 mi	468 mi	443 mi	h 20%
420	438	450	447	425	353	40%
416	430	431	413	360	229	70%
414	425	419	390	319	156	90%

Table 11—432-MHz Configured for 30.5 in. Hg, air pressure

432 MHz	Ant—10' Quad Twin Quad Yagi	
51.1 dBW	Xmtr Power	800 W
−192.1 dBW	Rcvr NF	0.5 dB
243.2 dB		
	Ant Gain	22 dB
Elevation 1°	Beamwidth	13°

T −10°	10°	32°	50°	70°	95°	P 30.5"
371 mi	394 mi	414 mi	424 mi	424 mi	401 mi	h 20%
368	389	402	401	381	313	40%
365	381	384	368	318	194	70%
363	376	372	346	278	126	90%

Table 12—432-MHz Configured for elevation of 1.6°

432 MHz	Ant—10' Quad Twin Quad Yagi	
51.1 dBW	Xmtr Power	800 W
−192.1 dBW	Rcvr NF	0.5 dB
243.2 dB		
	Ant Gain	22 dB
Elevation 1.6°	Beamwidth	13°

T −10°	10°	32°	50°	70°	95°	P 29.2"
283 mi	304 mi	322 mi	330 mi	330 mi	307 mi	h 20%
281	300	310	309	290	226	40%
278	292	293	278	231	121	70%
276	287	283	257	195	69	90%

Table 13—432-MHz Standard Configuration: dual Twin Quad

432 MHz	Ant—10' Quad Twin Quad Yagi	
48.3 dBW	Xmtr Power	800 W
−189.4 dBW	Rcvr NF	0.5 dB
237.8 dB		
	Ant Gain	19.3 dB
Elevation 1°	Beamwidth	18°

T −10°	10°	32°	50°	70°	95°	P 29.2"
347 mi	369 mi	388 mi	396 mi	396 mi	372 mi	h 20
345	364	376	374	353	286	40%
341	356	358	341	292	170	70%
340	351	346	320	253	106	90%

Table 14—432-MHz Standard Configuration: 8 Twin Quad Yagi—10'

432 MHz	Ant—10' Quad Twin Quad Yagi	
53.8 dBW	Xmtr Power	800 W
–194.8 dBW	Rcvr NF 0.5 dB	
248.6 dB		
	Ant Gain	24.7 dB
Elevation 1°	Beamwidth	10°

T –10°	10°	32°	50°	70°	95°	P 29.2"
448 mi	471 mi	490 mi	499 mi	498 mi	474 mi	h 20%
445	465	478	476	454	383	40%
442	457	459	441	390	256	70%
440	452	446	418	347	180	90%

Table 15—432-MHz Standard Configuration; low power 25 W

432 MHz	Ant—10' Quad Twin Quad Yagi	
36.0 dBW	Xmtr Power	25 W
–192.1 dBW	Rcvr NF 0.5 dB	
228.1 dB		
	Ant Gain	22 dB
Elevation 1°	Beamwidth	13°

T –10°	10°	32°	50°	70°	95°	P 29.2"
258 mi	278 mi	295 mi	303 mi	303 mi	281 mi	h 20%
256	273	285	283	263	202	40%
253	266	268	252	208	103	70%
251	261	257	232	173	56	90%

Table 16—900-MHz Low Power; 8' Cylindrical Parabola

900 MHz	Ant—8' Cylindrical Parabola	
37.0 dBW	Xmtr Power	25 W
–192.6 dBW	Rcvr NF 1.0 dB	
229.7 dB		
	Ant Gain	23.1 dB
Elevation 1°	Beamwidth	11°

T –10°	10°	32°	50°	70°	95°	P 29.2"
189 mi	208 mi	223 mi	231 mi	231 mi	210 mi	h 20%
188	203	213	212	195	140	40%
185	192	198	184	145	60	70%
183	193	188	166	115	28	90%

Table 17—1.3-GHz High Power; quad 2C39 cavity, water cooled

1300 MHz	Ant—8' Cylindrical Parabola	
52.2 dBW	Xmtr Power	400 W
–195.7 dBW	Rcvr NF 1.0 dB	
247.9 dB		
	Ant Gain	26.2 dB
Elevation 1°	Beamwidth	8°

T –10°	10°	32°	50°	70°	95°	P 29.2"
303 mi	324 mi	342 mi	350 mi	350 mi	327 mi	h 20%
301	319	330	329	309	245	40%
298	311	313	297	250	137	70%
295	307	302	276	213	80	90%

Table 18—2.3-GHz Standard Configuration; water-cooled 2C39 amplifier

2300 MHz	Ant—8' Cylindrical Parabola	
51.1 dBW	Xmtr Power	100 W
–200.7 dBW	Rcvr NF 1.0 dB	
251.8 dB		
	Ant Gain	22.2 dB
Elevation 1°	Beamwidth	4.5°

T –10°	10°	32°	50°	70°	95°	P 29.2"
265 mi	285 mi	303 mi	311 mi	310 mi	288 mi	h 20%
263	281	291	290	271	210	40%
260	274	275	260	215	110	70%
258	269	265	240	181	66	90%

Table 19—3.3-GHz Medium Power Xstr Power Amplifier

3300 MHz	Ant—8' Cylindrical Parabola	
37.3 dBW	Xmtr Power	2 W
–203.8 dBW	Rcvr NF 1.0 dB	
241.1 dB		
	Ant Gain	34.3 dB
Elevation 1°	Beamwidth	3°

T –10°	10°	32°	50°	70°	95°	P 29.2"
140 mi	156 mi	170 mi	177 mi	176 mi	158 mi	h 20%
138	153	161	160	145	98	40%
136	147	148	136	102	36	70%
135	143	140	120	78	15	90%

Table 20—5.6-GHz Gunn Diode Transmitter

5600 MHz	Ant—8' Cylindrical Parabola	
28.8 dBW	Xmtr Power	0.1 W
–208.4 dBW	Rcvr NF 1.0 dB	
237.2 dB		
	Ant Gain	38.8 dB
Elevation 1°	Beamwidth	2°

T –10°	10°	32°	50°	70°	95°	P 29.2"
72 mi	83 mi	94 mi	99 mi	99 mi	85 mi	h 20%
70	80	87	86	75	44	40%
68	76	77	68	43	12	70%
68	73	71	58	32	5	90%

Table 21—10-GHz, 10-W Power Amplifier

10000 MHz	Ant—8' Cylindrical Parabola	
58.9 dBW	Xmtr Power	10 W
–213.5 dBW	Rcvr NF 1.0 dB	
267.4 dB		
	Ant Gain	43.9 dB
Elevation 1°	Beamwidth	1°

T –10°	10°	32°	50°	70°	95°	P 29.2"
201 mi	219 mi	234 mi	241 mi	241 mi	221 mi	h 20%
200	215	225	223	206	154	40%
197	209	210	196	159	73	70%
195	205	201	180	130	37	90%

Table 22—10 GHz—Lower Air Pressure

10000 MHz	Ant—8' Cylindrical Parabola	
53.9 dBW	Xmtr Power	10 W
−213.4 dBW	Rcvr NF 1.0 dB	
267.3 dB		
	Ant Gain	43.9 dB
Elevation 1°	Beamwidth	1°

T −10°	10°	32°	50°	70°	95°	P 28"
221 mi	238 mi	253 mi	259 mi	258 mi	238 mi	h 20%
219	234	243	240	223	168	40%
216	228	228	213	174	83	70%
215	223	218	196	143	43	90%

Table 23—10 GHz—Higher Air Pressure

10000 MHz	Ant—8' Cylindrical Parabola	
53.9 dBW	Xmtr Power	10 W
−213.4 dBW	Rcvr NF 1.0 dB	
267.3 dB		
	Ant Gain	43.9 dB
Elevation 1°	Beamwidth	1°

T −10°	10°	32°	50°	70°	95°	P 30.5"
181 mi	199 mi	215 mi	222 mi	223 mi	205 mi	h 20%
180	195	205	205	190	140	45%
177	190	191	179	143	63	70%
176	185	183	163	115	31	90%

Table 24—10 GHz—Increased Beam Elevation

10000 MHz	Ant—8' Cylindrical Parabola	
53.9 dBW	Xmtr Power	10 W
−213.4 dBW	Rcvr NF 1.0 dB	
267.3 dB		
	Ant Gain	43.9 dB
Elevation 1°	Beamwidth	1°

T −10°	10°	32°	50°	70°	95°	P 29.2"
110 mi	124 mi	136 mi	142 mi	142 mi	129 mi	h 20%
108	121	128	127	114	74	45%
106	116	117	106	78	25	70%
105	113	110	93	57	10	90%

Table 25—10 GHz—Reduced Antenna Size—3' Cylindrical

10000 MHz	Ant—3' Cylindrical Parabola	
45.3 dBW	Xmtr Power	10 W
−204.9 dBW	Rcvr NF 1.0 dB	
250.3 dB		
	Ant Gain	35.3 dB
Elevation 1°	Beamwidth	3°

T −10°	10°	32°	50°	70°	95°	P 29.2"
105 mi	120 mi	132 mi	138 mi	138 mi	121 mi	h 20%
104	116	124	123	109	70	45%
101	111	112	101	73	22	70%
100	108	105	88	53	9	90%

Table 26—10 GHz—3' Parabolic Dish

10000 MHz	Ant—3' Parabolic Dish	
47.0 dBW	Xmtr Power	10 W
−206.6 dBW	Rcvr NF 1.0 dB	
253.6 dB		
	Ant Gain	37 dB
Elevation 1°	Beamwidth	2.3°

T −10°	10°	32°	50°	70°	95°	P 29.2"
123 mi	138 mi	152 mi	158 mi	158 mi	141 mi	h 20%
122	135	143	142	128	85	45%
120	130	131	120	88	30	70%
118	126	123	105	66	12	90%

Table 27—10 GHz—Gunn Diode Transmitter

10000 MHz	Ant—8' Cylindrical Parabola	
33.9 dBW	Xmtr Power	0.1 W
−213.4 dBW	Rcvr NF 1.0 dB	
247.3 dB		
	Ant Gain	43.9 dB
Elevation 1°	Beamwidth	1°

T −10°	10°	32°	50°	70°	95°	P 29.2"
75 mi	87 mi	98 mi	103 mi	102 mi	84 mi	h 20%
74	84	91	90	79	47	45%
73	80	81	72	50	13	70%
71	78	75	62	35	5	90%

Table 28—10 GHz—High Power—Maybe a Traveling Wave Tube

10000 MHz	Ant—8' Cylindrical Parabola	
63.9 dBW	Xmtr Power	100 W
−213.4 dBW	Rcvr NF 1.0 dB	
277.3 dB		
	Ant Gain	43.9 dB
Elevation 1°	Beamwidth	1°

T −10°	10°	32°	50°	70°	95°	P 29.2"
281 mi	300 mi	316 mi	324 mi	323 mi	303 mi	h 20%
279	295	306	304	286	228	45%
276	289	290	275	233	129	70%
274	284	280	256	199	77	90%

Table 29—24 GHz—Gunn Diode Transmitter

24000 MHz	Ant—4' Cylindrical Parabola	
35.4 dBW	Xmtr Power	0.1 W
−215.0 dBW	Rcvr NF 1.0 dB	
250.5 dB		
	Ant Gain	45.5 dB
Elevation 1°	Beamwidth	1°

T −10°	10°	32°	50°	70°	95°	P 29.2"
40 mi	47 mi	55 mi	58 mi	58 mi	48 mi	h 20%
39	45	50	49	41	22	45%
38	43	43	38	23	5	70%
37	41	39	31	15	2	90%

Table 30—Receiver Noise Figure Distances

NF (dB)	432	10000		
0.5	519	154	Humidity:	40%
1	514	151	Temp:	32°F
1.5	509	147	Air Pressure:	29.2"
2	504	143	Antenna Gain:	32 dB
3	494	136	Power:	100 W
4	484	130	Elevation:	1°
5	474	123		
6	464	117		

Digital Communications

Harold E. Price, NK6K

High-Speed Modems

There have been high-speed modems (higher than 1200 baud) in the amateur bands long before packet radio became popular. A Montreal group in the late seventies routinely used 9600 baud. A Los Angeles group, based at SRI, demonstrated 250 kbps using TV modules in the early eighties. Still, though, we lament the lack of high-speed modems for the common ham in the nineties. Part of this is a chicken-and-egg problem. Packet radio itself didn't take off until we reached a critical mass, that is, enough people or computers on the air so that you stood a reasonable chance of having something to connect to when you got on the air.

Another part is that individuals don't need high data rates if their packet usage is limited to point-to-point, keyboard-to-keyboard QSOs. Effective use of high speed usually means a network link, where several users are "multiplexed" on the same channel, or a file server link, where several hundred thousand bytes are downloaded at a time. Servers and networks require coordinated action between individuals or groups, and pooled financial resources to acquire, set up, and maintain computers, high-speed TNCs or adapter cards, high-speed modems, and UHF/microwave radios. This, in my opinion, is the main problem.

There is one other problem that I can do more about. I can provide a forum for high-speed advocates to tell the rest of us how to join them. This month, I want to let you hear what Barry McLarnon, VE3JF, has to say:

> I'm an admirer, and tireless promoter, of the GRAPES 56-kbps (or more) modem. It's amazing that this modem has been around for about 6 years now, and yet it is still *rara avis* in the packet radio world. It's not hard to build, it tunes up easily, and it requires no messy radio modifications—just standard VHF/UHF up/down converters. We've been using the modem here in Ottawa since 1988, and our 56-kbps full-duplex regenerating repeater has been in continuous operation since January 1990. The 56-kbps MAN is the backbone of our local network, linking our hub switch (Gracilis PackeTen) and Internet gateway to all the local nodes plus a handful of "power users." We also developed the Ottawa PI DMA interface board as an affordable way to drive fast modems like the GRAPES modem with PCs. 56 kbps ain't plug 'n play, but it ain't that hard either . . . What's everybody waiting for?

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71635,1174 (CompuServe)

Note that is not enough simply to have a modem that can send at 56 kbps. You also need a radio to match, and some device that can generate and accept data at that rate. Barry has been compiling a list of high-speed modem options and makes it available in a file called hispeed.003. It can be downloaded by anonymous ftp from hydra.carleton.ca, in pub/hamradio/packet. I've also placed a copy of the file, called HISPD3.ZIP on Compuserve in HamNet's DL9. The 56-kb portion of that file follows.

56-kb Options from Barry

The purpose of the following is to summarize the hardware options available for constructing medium- to high-speed packet links. Thus far, only 9.6, 19.2, and 56 kbps are covered. This material is intended to be a useful reference, but I make no claims as to its accuracy or completeness. Many details concerning model numbers and prices are missing, and I have very little information concerning equipment sources outside North America. If you have corrections, or suggestions on additional information to include in this survey, please send them to bm@hydra.carleton.ca (or ve3jf@ve3jf.#eon.on.can.noam).

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Equipment for 56 kbps

The GRAPES (WA4DSY) modem, is $250 in kit form. You also need to provide a box for it, plus a few interconnecting cables and connectors. It requires 5 V power (about 0.5A @ +5 V, 0.1 A @ –5 V). This is an RF modem with input and output (about 1 mW) in the 28-30 MHz band, designed for use in the bands above 220 MHz (occupied bandwidth is about 70 kHz at 56 kbps), using standard receive and transmit converters. The receive and transmit portions of the modem are separately crystal-controlled, and it can run full-duplex. It is not limited to 56 kbps—with suitable modifications, it can be made to work at 128 kbps or more.

Data Interfaces for 56 kbps

Ottawa PI Board (price: $120 US, $140 CDN). This board provides a DMA port which handles 56 kbps with ease, even with a 4.77-MHz XT-class machine. All you need to add is the cable to the modem. The main limitation of the board is that it does not support full-duplex operation, but full-duplex operation is rare (especially among end users). The PI also supports a low-speed port (you provide the modem and radio). The board can be used with any version of KA9Q NOS.

Gracilis TWIN-1E PackeTwin Card (price: $225.) Like the PI, it provides a DMA port for the 56-kbps modem and an interrupt-driven port for lower-speed modems. The DMA port supports full-duplex operation. The Kantronics 9600-bps modem can be piggybacked on the card.

Kantronics Data Engine (DE) (price: $299 with no modems). This is essentially a higher-speed TNC with two HDLC ports that can reportedly run at 56 kbps, and an RS-232 port that can run at up to 19.2 kbps. The standard firmware is G8BPQ, but there is now also a port of JNOS (JNOS40) by WG7J available. The DE appears to be more useful as a small standalone packet switch then as an interface for end users.

Gracilis PackeTen (price: around $1500). This is a full-blown packet switch that runs a custom version of KA9Q NOS. It is available in both stand-alone and PC bus versions. If you need more than two high-speed ports (more than one, if you need full-duplex), then this is really the only choice.

RF Equipment for 56 kbps

The RF equipment required depends on whether the links are half- or full-duplex. There are three basic configurations in use:

Half-Duplex Point-to-Point Links

An example is the Georgia backbone network. The usual RF equipment is a Microwave Modules (220, 430 MHz) or Sinclabs (220 MHz) transverter.

Full-Duplex Point-to-Point Links

Full duplex operation is significantly more complicated, but it is also highly desirable if you want to maximize the throughput of a backbone link. The GRAPES modem is inherently full-duplex, so it is only necessary to provide separate RF up- and down-converters. The two channels may be in-band or cross-band, using either separate antennas or duplexers. The only full-duplex point-to-point link I'm aware of is in Chicago—it uses PackeTen switches and operates in-band in the 70-cm band.

Multiple-Access Networks with a Full-Duplex Repeater

In this case, an in-band or cross-band 56-kbps repeater provides hidden-transmitter-free access to a channel (or rather, a pair of channels) by multiple 56-kbps stations. This might just be a LAN for the power users, but it also is an attractive means of linking a number of network nodes together, with less complexity than multiple point-to-point links (see the *10th ARRL Computer Networking Conference* proceedings for more details). As in the preceding case, separate receive and transmit converters are used, usually with separate antennas. (In principle, a transverter with "split" frequency operation could be used, but such things are hard to come by.) The stations in this network do not require full-duplex computer interfaces, but since the RF portions have full-duplex capability, it allows smaller TXDELAYs to be used than in the half-duplex case. It also allows users to observe the quality of their signals coming back from the repeater.

The first 56-kbps full-duplex repeater went on the air in Ottawa in January 1990. The repeater is cross-band (220.55 MHz in, 433.55 MHz out), so users must up-convert the modem's 28-30 MHz IF output to 220 MHz, and down-convert 432 MHz to the 28-30 MHz IF input.

430-MHz RF Equipment

[Note: Barry covers 220 MHz, 430 MHz, and 1.2 GHz in detail. For space reasons I'm only including the 430 version here—NK6K]

Transverters and Up-Converters

Down East Microwave DEM432 no-tune transverter, 50-100 mW output. This is a 3-board set, available in several forms, and there is an optional power amplifier that provides 15-W output. The local oscillator board normally has a single oscillator for standard half-duplex operation, but a second oscillator can be added on the board for half-duplex split or full-duplex operation. Some options and prices:

- DEM432B assembled and tested unit, including case, $275
- DEM432B DUAL as above, but set up for dual frequencies, $300
- DEM432K basic kit (no case or connectors), $155
- Second LO kit, $8
- 432PA 15W PA, assembled and tested, $180
- 432PACK 15W PA complete kit, $135
- 432PAK 15W PA basic kit (no case, connectors or heat sink), $75
- Enclosure to house both DEM432K and 432PA, $25
- DEM432-15S complete 15-W dual-frequency transverter, $395

Microwave Modules MMT432/28S transverter, 10-W output. Not readily available new, but quite a few used ones are on the market.

SSB Electronic TV 28-432 transverter (price: $310), 100-mW output. These units have no T/R switching, so that would have to be added externally for single-channel half-duplex operation. On the other hand, there are separate local oscillators provided for the receive and transmit converters, so this looks like a good choice for in-band full-duplex or half-duplex split operation.

Down-converters

Hamtronics ($49/$69/$99 for basic kit/kit with box/wired & tested). Quality of this unit is uncertain.

Microwave Modules MMc435.2 ($115). Current availability unknown.

SSB Electronic K7001-10 ($180). High quality, with a price to match.

There are other sources for units in the $100-$150 range, such as Lunar.

56-kbps Summary

The cost of a 56-kbps station is a bit hard to pin down, given all the variables. As an example, we'll consider a station for the Ottawa 56-kbps LAN. The modem kit and a PI board will set you back about $370. The rest depends on the choice of RF components. The total will vary from about $500 to $800. The "low road" is using the Hamtronics kits and scrounging up things such as boxes for them and the modem, homebrewing the antennas, etc. The "high road" is buying higher-quality assembled and tested gear, such as the Sinclabs transverter and the MM receive converter. If you can find some good used gear, the total should be closer to $650. Getting on 56 kbps is certainly a more challenging project than plug 'n play 9600, but the rewards are greater too.

Sources

Down East Microwave
RR 1, Box 2310
Troy, ME 04987
tel: 207 948-3741

DRSI (Digital Radio Systems Inc)
2065 Range Road
Clearwater, FL 34625
tel: 813 461-0204

Gracilis Inc
623 Palace Street
Aurora, IL
tel: 708 801-8800
fax: 708 844-0183
email: info@gracilis.com (Internet)

GRAPES Inc
PO Box 871
Alpharetta, GA 30239-0871
email: dug@kd4nc.atl.ga.us (Internet)

Kantronics
1202 E 23rd Street
Lawrence, KS 66046
tel: 913 842-7745

Maple Leaf Communications
Bob Morton, VE3BFM
RR 1
Everett, ON, Canada L0M 1J0
tel: 705 435-0689

MFJ Enterprises Inc
PO Box 494
Mississippi State, MS 39762
tel: 800 647-1800 (order)

Ottawa Amateur Radio Club
Packet Working Group
PO Box 8873
Ottawa, ON, Canada K1G 3J2
email: bm@hydra.carleton.ca (Internet)

PacComm Packet Radio Systems Inc
4413 N Hesperides Street
Tampa, FL 33614-7618
tel: 813 874-2980

SSB Electronic USA
124 Cherrywood Drive
Mountaintop, PA 18707
tel: 717 868-5643 □□