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About the Cover:

Brandon Taylor, of Bethel Park, PA, fearlessly pilots a soapbox racer equipped with NK6K's single-board computer instrumentation.

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Features

Columns

August 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

Nonamateur Packet Connection

In recent months, several efforts aimed toward using nonamateur means of interconnecting amateur packet-radio networks have begun or continued. Some of these efforts use donated capacity on commercial common-carrier networks. Others use Internet or commercial packet-switched computer networks. Both NET/ROM and TCP/IP systems have been connected via such networks.

Connecting two packet nodes together through a landline connection is not exactly new, of course. It was happening almost as soon as packet radio began in the amateur service. And "wormholes" between distant networks have been around for some time, too. So, the mere existence of these connections isn't news. What *is* new are the organized efforts to make these connections part of workaday amateur packet communications on a regular basis. It is the case in some networks now that you can hardly tell whether the system you are communicating with is local or thousands of kilometers distant.

Few would argue the benefits of such networking, but there are those who question the means. Is it a good thing for amateurs to be transporting their data partly via nonamateur means? Is that "real ham radio?" Or is it an abomination that further erodes the purity of the "amateurness" of ham radio? How you feel about the issue may depend on what you want from packet radio. Those who want to use packet as one of the communications tools Amateur Radio provides tend not to care so much how the data gets there. Those who feel they are advancing the amateur state-of-the-art often feel that using nonamateur means is "cheating."

The ARRL Digital Committee, in a meeting held in March of 1993, expressed enthusiasm for the use of nonamateur means of extending the amateur network. The members of the Committee commented, however, that the long-term goal should be to develop an all-amateur network so as not to rely solely on other means of interconnecting amateur networks. For now, at least, we need to get the bits where they are going. Being excessively fastidious about the medium by which they get there provides no benefit and unduly restricts our potential communication capability.

This Month in *QEX*

Noise blankers and strong signals don't mix, and Mark Mandelkern, KN5S, shows one way to keep them separate with "Evasive Noise Blanking."

DSP filters are becoming commonplace. Russ Ward, WA4ZZU describes a simple design technique for one class of filter in "Basic Digital Filters."

Part 2 of our 6-part series on ALE concerns generation of test signals for ALE systems, as described by David Wortendyke, NØWGC, et al.

Over the years, the search for a broadband 80-meter antenna has led to many strange configurations. A simple transmission-line approach is presented by Hugh Fallis, N4HF, whose computer-model designs are presented in, "Broadband Dipole and Unipole 3.5-to-4.0-MHz Antennas."

Finally, *QEX* columnist Harold Price, NK6K, describes a project he did recently using an off-the-shelf single-board PC in his "Digital Communications" column.—*KE3Z, email: jbloom@arrl.org (Internet)*

Evasive Noise Blanking

Independent tuning of the noise-blanker channel provides crunch-proof noise blanking.

Mark Mandelkern, KN5S

For the low-band HF DXer, breaking the pile-up may be the biggest problem. But on the higher HF bands, and especially for competitive weak-signal SSB/CW work on the VHF bands, hearing signals through that ubiquitous man-made noise often becomes the crucial determining factor for success. Efforts to eliminate the noise take various avenues, from replacing the neighbor's doorbell transformer to persuading the power company to fix the lines, or escaping to a mountaintop for a contest. But after all other efforts, we are reduced to dependency on our noise blankers. Alas, were it only so easy as pushing a button! The cold, cruel reality is that noise blankers do sometimes blank the noise, but they often also add more trouble than they remove.

A typical response from a nearby station, when asked "Is my signal troubling your receiver," is "Well, you're not overloading my receiver, but you're crunching my noise blanker." Modern strong-receiver technology has just about eliminated the problem of receiver overload, once a serious matter. But modern noise blankers are notorious for folding up under the stress of strong signals on nearby frequencies. Strong signals cause two problems with noise blankers. My experience shows that one of these has by far the worst effect, and I've recently built a transceiver that avoids the problem.

Let us first briefly review the operation of a noise blanker. A block diagram of a typical blanker is shown in Fig 1. There are three main sections. The noise amplifier amplifies the noise pulses to a level sufficient for the detector and pulse-forming network, which produces a control signal, called the blanking pulse. This blanking pulse is routed to the receiver gate, where it disables the receiver during the noise pulse. Sounds simple, right?

Now, what goes wrong? First, the blanking pulses modulate any signal in the passband, turning them on and off rapidly. The effect is to create noise sidebands on either side of the signal, just like key clicks from a transmitter without proper key shaping. But this effect is minor compared to the second problem: A strong nearby signal—one outside the 2-kHz filter but *within* the 25-kHz filter—will enter the noise amplifying channel and create false blank-

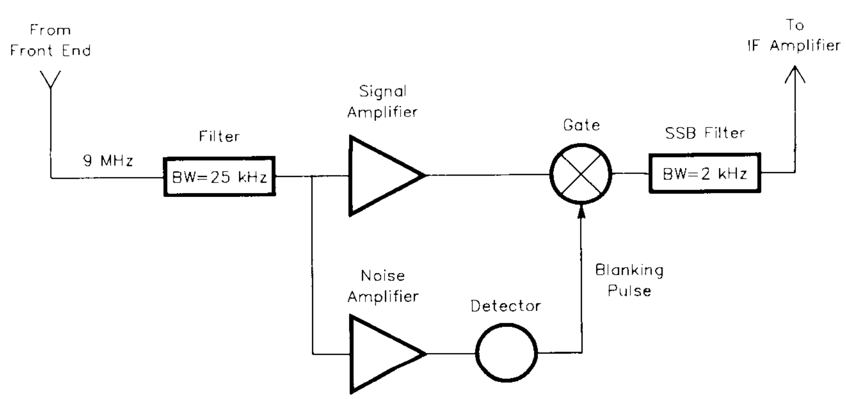

Fig 1—Block diagram of a typical modern noise blanker.

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Las Cruces, NM 88005

ing pulses which tend to quiet the receiver, eliminating the desired signal. We say the receiver is "crunched."

I know of three noise blanker systems which avoid this problem. First is the Collins system of the 1950s. Second, in 1967 I built a copy of the Collins system which I adapted to a Heath SB-301 and a 6-meter converter, for noise-free 50-MHz reception. Third is the system I recently built into my new home-brew transceiver. I've tried to design *everything* into this new radio. It covers all the bands from 1.8 to 144 MHz, has dual VFOs, all-quartz reference oscillators (no microprocessor or synthesizer—and no birdies or phase noise!), IF shift, RF speech clipping, panel-adjustable CW offset, dual receive—the works! But most important, I wanted to get rid of the infuriating power line noise—without being crunched. The result is the evasive noise blanking system described here.

The Collins System

The Collins system is shown in Fig 2; it was designed mainly for mobile operation with the KWM-2. It differs from the modern blanker in that the noise is picked up at 40 MHz by a separate antenna. In a mobile installation, the vehicle's broadcast radio antenna would be used. The main disadvantages of this system were twofold. First, the noise heard at the operating frequency might not also appear at 40 MHz. The second disadvantage occurred in a receiver used in a home station with an antenna having some gain (a 10-meter beam, for example), and a separate 40-MHz whip for noise pick-up. Even if the noise were also at 40 MHz, still the whip might not hear noise heard by the beam. No blanking would occur. A separate 40-MHz rotatable beam would be ideal, but I don't know if anyone actually did this.

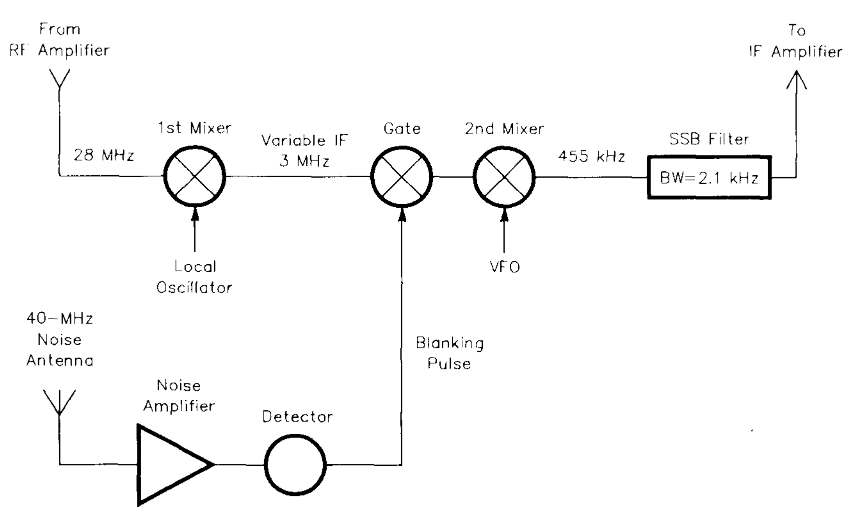

Fig 2—Block diagram of the Collins blanker. The noise amplifier limits the bandwidth of the noise channel to about 500 kHz.

A Six-Meter System

The system I built into my SB-301, shown in Fig 3, was adapted from the Collins circuit.[1] The gate was built for the Heath first IF frequency at 8 MHz. More important, the noise amplifier was built for 51 MHz, and the 51-MHz noise was taken from a tap after the RF amplifier in the 6-meter converter. Thus the noise amplifier had the advantage of using the large 6-meter beam and could hear noise that was no louder than the weakest 6-meter DX signal. This system avoided both disadvantages of the Collins system, and the results were spectacular.

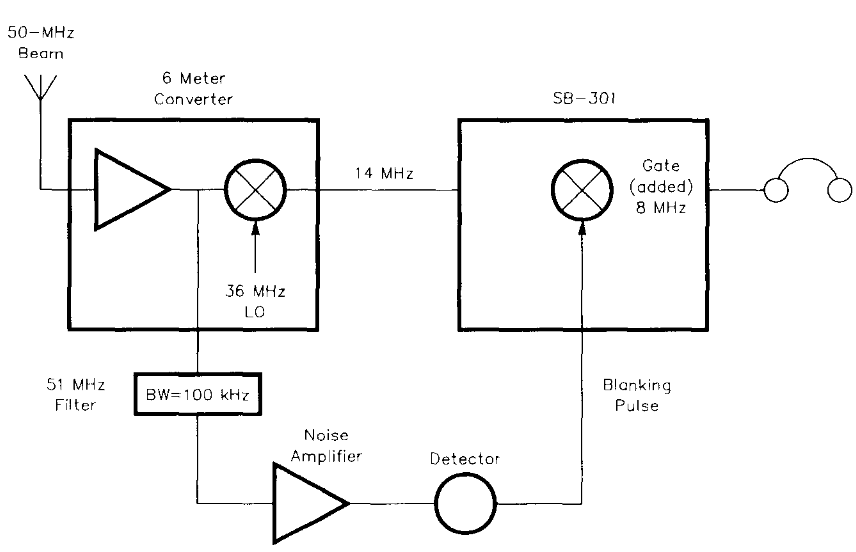

Fig 3—Block diagram of the blanker for 6 meters built for a Heath SB-301 in 1967. The noise channel operates at 51 MHz, while the desired signals are at 50-50.5 MHz.

The Modern System

After many years of missing the old Heath, I began a 3-year project of building a transceiver, the last piece in a completely home-brew station. The design goal for the blanker in the new radio was that it should provide the advantages of the old 51-MHz system on all bands. The transceiver tunes 1-MHz segments, and a natural extension of the old system would be to listen for noise at the top of the band in use. But now a third problem could be foreseen. While such a fixed frequency noise amplifier was okay for 6 meters, where I could predict where signals would be found, it might not work so well on other bands. (Is there a local repeater on 145 MHz?)

The solution was a tunable noise amplifier—to *evade* the strong signals. Working 24.9 MHz? Just tune the noise amplifier to any clear frequency between 24.0 MHz and 24.8 MHz. One panel knob does the tuning. Once a clear

[1]Notes appear on page 6.

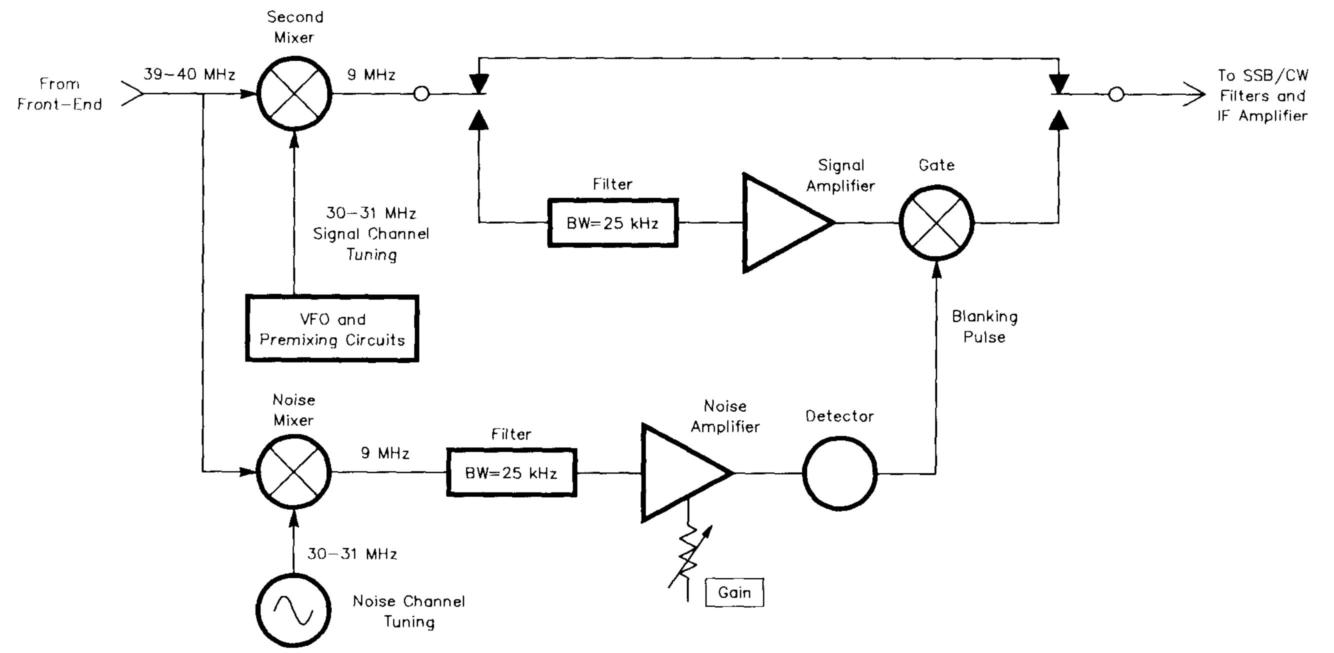

Fig 4—Block diagram of the evasive noise blanker.

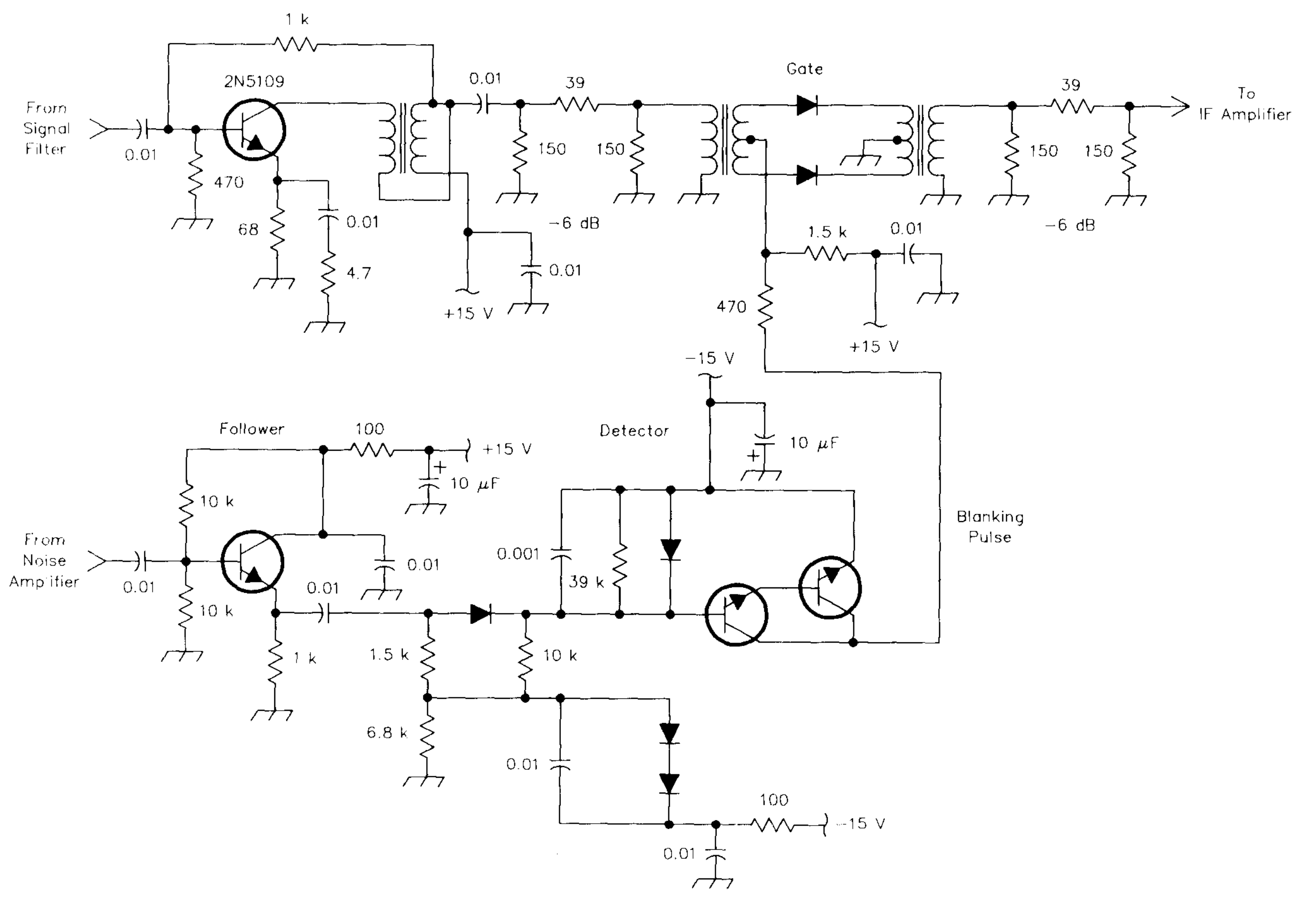

Fig 5—Schematic diagram of the noise detector, pulse-forming network, and receiver gate used in the author's home-brew transceiver.

spot is found, it seldom needs attention. Another knob adjusts the noise blanker gain, which is set according to the ambient noise level. Any noise above this level at the noise channel frequency will cause a blanking pulse. This control is a push/pull type for on/off blanker switching. Both the noise amplifier and the receiver IF at the gate have bandwidths of 25 kHz. This width prevents the noise pulses from being excessively broadened, as they would be with sharp filters. Some noise blankers use a somewhat narrower bandwidth. I had the 25-kHz filters in the junk box, but some experimentation might be called for here. The usual sharp SSB/CW filters come only after the gate. (I have 2-kHz, 400-Hz, and 200-Hz IF filters available.)

The block diagram of my evasive noise blanker is shown in Fig 4. The radio uses dual conversion on all bands from 1.8 to 144 MHz, with high-side first injection. The IF frequencies are 40 MHz and about 9 MHz. The first IF tunes from 40 to 39 MHz. The noise blanker samples the first IF and gates the second IF. When the noise blanker is off, relays switch it completely out of the circuit. The amplifier ahead of the gate replaces the gain lost in the filter and gate, so there is no change in overall receiver gain when the noise blanker is switched in or out.

The evasive noise blanker can be adapted to any IF. Any of the *Handbook* circuits can be used for the noise mixer, oscillator, and amplifier, since these are non-critical.[2] For the noise detector and receiver gate, you can copy the circuit from your favorite commercial radio. The circuit I used is shown in Fig 5. The noise amplifier has no AGC and the operator must use the noise amplifier gain control on the panel. This may be a disadvantage compared to some modern "automatic" types, but my hunch is that it may actually be an advantage. There is also room for experimentation here.

One point which does not show up in the schematic should be mentioned. My radio has no AGC applied to the front end. One reason for this is out of consideration for the blanker. When AGC from the narrow IF is allowed to reduce the gain ahead of the noise channel, some weird effects result. Noise may be actuating the AGC, then the blanker gain is increased, the noise is driven out of the IF, the AGC voltage drops and the front-end gain increases. This has a positive feedback effect which may saturate the noise detector. The operator experiences a hysteresis type of blanker latch-up. This problem is avoided by applying AGC only to the IF amplifier.

Adapting the Evasive Noise Blanker

Perhaps the most likely adaptation of this system would be as a stand-alone device at 28 MHz for use between VHF converters and an HF radio. It might have a noise channel at 29 MHz, and blank the entire IF. Or it might have a tunable noise channel and also use down/up mixing to blank only a tunable 25-kHz segment. It sounds like a lot of work, but serious operators will stop at nothing to hear the DX!

Another possible variation would be to simply take a 28-MHz tap off the VHF converter output. Tune, filter, amplify, and detect the noise, and then (here's the rub) find the spot in your receiver's original noise blanker gate at which to apply the blanking pulse. Still other variations would involve two receivers.

On the lower bands there is a special problem. The noise is usually of atmospheric origin rather than the man-made pulse noise on which noise blanking has half a chance. And there may be signals in the entire 1-MHz segment. Narrow bandwidth antennas, as on 160 meters, are another limitation. The next step is to try using a VHF band to feed the noise channel while operating a lower HF band. This just means separate band switches for the operating channel and the noise channel.

This evasive noise blanker has been in constant use for over a year. Results on the VHF and higher HF bands have been excellent. I can use the blanker as close as I wish to loud stations. I hope you have fun trying the evasive blanker; please write to me about your results. There is still much room for experimentation to fight this noise problem. New noise sources are being invented faster than better blankers!

Notes

1 *Fundamentals of Single Side Band*, 2nd edition, Collins Radio Company, Cedar Rapids, Iowa, 1959.

2 *The 1993 ARRL Handbook for Radio Amateurs*, Newington, CT, 1992

Basic Digital Filters

Use a simple BASIC program to design DSP bandpass filters.

Russ Ward, WA4ZZU

Digital Signal Processing (DSP) is not magic, nor, as J. H. Karl confirms, does it require a high priest.[1] DSP involves two computer programs of only moderate complexity, but most DSP writing would lead you to think it's fearsomely complicated. This article will show you how to easily design one type of digital filter and will present some ideas about implementation.

All digital filter systems contain three parts: input, processing, and output. The filter input converts the analog signal to a digital one. Assume we want an audio filter up to 3 kHz. The analog input signal is chopped into pieces (sampled) at a rate of twice the audio bandwitdth, or 6 kHz. Each piece is given a digital value that corresponds to the amplitude of the signal. The maximum possible digital value sets the precision, in bits. For example, there are 256 possible values in an 8-bit conversion. So our audio signal consists of 6000 values per second, each value being 8 bits (1 byte). The output section of the filter does a digital-to-analog conversion of the byte stream, resulting in the analog audio signal. One chip may do both conversions.[2]

Appropriately manipulating the 6-kHz byte stream with a computer gives a digital filter. Fig 1 shows a finite impulse response (FIR) digital filter. The output of the filter depends on a chosen number of past values which are temporarily stored in the boxes labelled "delay." The bytes are shifted right at a 6-kHz rate. The number of delays is an important design parameter; more delays mean steeper filter skirts. Each output byte is the sum of all the delay values, each multiplied by a corresponding factor, h_n. Since Fig 1 has M delays, there will be $M + 1$ multiplications and M additions per output byte. The h factors are mostly small numbers and are somewhat critical. Determination of the h factors is what must be done to design a digital filter.

Determination of the h factor is easily done with a computer. Program 1 is a BASIC program that does just that. I modified this program, which I got from the book, *Introductory Digital Signal Processing*, by P.A. Lynn and W. Fuerst (Wiley, 1989). This is a really good book, and I recommend it. My modifications make the program easier to use for hams.

This simple program has the usual

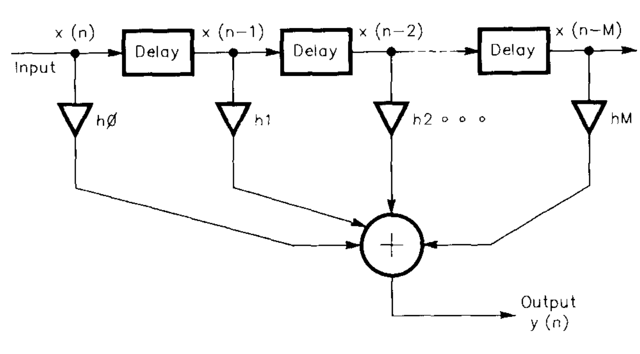

Fig 1

[1]Notes appear on page 8.

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quirks. It works only for an odd number of delays (also called taps). And it scrolls the factors fast if more than 25 delays are requested. To determine the number of taps, I use this rule of thumb: the minimum number of taps = 20,000/bandwidth (Hz). Note that the frequency plot generated by the program does not go below –50 dB. Being BASIC, it is slow; relax and watch the "–50" blink. The windows mentioned in the program are criteria which modify the *h* factors to obtain better filter shape; they are representative examples. Try them and watch the filter shape change.

The second program you will need uses the *h* factors to modify the input data stream. This second program is the actual digital filter.

For a filter with *M* taps, we have 6,000 × (*M* + 1) multiplications and 6,000 × *M* additions per second. The processing cannot be done directly in real time on a PC; it requires mathematical tricks. IW5BHY showed one trick to reduce processing, using factors which are all 0 or 1.[3] Another common trick is called a fast Fourier transform (FFT) and is beyond the scope of this article. If one is not concerned with realtime, the ideas in this article could be coded in BASIC. For example, you might try filtering a tape of the ZRO satellite receive test.

Speed is crucial to this second computer program. Real-time processing often requires assembler language, and often a special DSP chip is used. DSP chip assemblers are strange and expensive. But one can use the same computer for the above design program and for filter implementation. Sanderson detailed his SSTV DSP application at Dayton in 1993; it had 41-tap filter running on a 80486. Either seek out existing 80x86 code or convert one of the many available FORTRAN FFT programs. The assembler code is not very complex. IW5BHY gives non-FFT 80x86 code.

This article has described the basics of DSP filter design and has shown a BASIC language design program. It has also presented a few ideas concerning implementation of a DSP filter. Go, now, and do some DSP!

Notes

[1] *An Introduction to Digital Signal Processing*, J. H. Karl, Academic Press, 1989. Preface.

[2] "Low-Cost Digital Signal Processing for the Radio Amateur," D. Herschberger, *QST*, September 1992, p 43.

[3] "Digital Filter for EME Applications," A. Dell'Immagine, *QEX*, May 1992, p 3.

```
10 CLS
20 DIM SEE(200),H(200),W(320)
30 PRINT "Enter Center Freq (Hz)":INPUT A
40 W0=3.141593*(A/3000)
50 PRINT "Enter Bandwidth (Hz)":INPUT B
60 W1=.5*(B/3000)*3.141593
70 PRINT " Enter M (# Taps = 2M+1 )":INPUT M
80 PRINT"select window:"
90 PRINT"0=rectangular; 1=von hann; 2=hamming":INPUT X:CLS
100 REM
110 REM ********   compute window values   ********
120 IF X=0 THEN A=1:B=0:C=1:GOTO 150
130 IF X=1 THEN A=.5:B=.5:C=M+1:GOTO 150
140 A=.54:B=.46:C=M
150 FOR N=1 TO M:SEE(N)=A+B*COS(N*3.141593/C):NEXT N
160 REM
170 REM ********* compute impulse response values **********
180 H0=W1/3.141593:FOR N=1 TO M
190 H(N)=(1/(N*3.141593))*SIN(N*W1)*COS(N*W0)*SEE(N):NEXT N
200 REM ********* draw rectangular grid for plot *******
210 SCREEN 2,0
220 FOR K=0 TO 5
230 K20=(30+(20*K)):K64=64*K
240 LINE(0,K20)-(320,K20)
250 LINE(K64,30)-(K64,130):NEXT K:LINE(319,30)-(319,130)
260 LOCATE 18,1:PRINT "0"
270 LOCATE 18,7:PRINT "600"
280 LOCATE 18,15:PRINT "1200"
290 LOCATE 18,24:PRINT "1800"
300 LOCATE 18,31:PRINT "2400"
310 LOCATE 18,39:PRINT "3000"
320 LOCATE 5,42:PRINT "0 db"
330 LOCATE 7,42:PRINT "-10"
340 LOCATE 9,42:PRINT "-20"
350 LOCATE 12,42:PRINT "-30"
360 LOCATE 14,42:PRINT "-40"
370 LOCATE 17,42:PRINT "-50"
380 LOCATE 2,4
390 PRINT "FIR DSP FILTER FREQUENCY RESPONSE"
400 LOCATE 20,4
410 X = 2*M+1
420 PRINT X " Tap Filter  6 kHz Sample Rate"
430 LOCATE 22,1
440 REM
450 REM ********** compute frequency response  ********
460 FOR N=1 TO 320:FREQ=3.14159*(N-1)/320
470 W(N)=H0:FOR K=1 TO M:W(N)=W(N)+2*H(K)*COS(K*FREQ):NEXT K
480 LOCATE 17,42:IF N > 100 THEN PRINT "qrx"
490 LOCATE 17,42:IF N > 190 THEN PRINT "-50"
500 NEXT N
510 REM    normalise to unity ** convert to db and plot ***
520 MAX=0:FOR N=1 TO 320:IF ABS(W(N))>MAX THEN MAX=ABS(W(N))
530 NEXT N
540 FOR N=1 TO 320
550 DB=20*LOG(ABS(W(N))/MAX)*.4343:IF DB<-50 THEN DB=-50
560 IF DB=-50 THEN DB=-49
570 LINE(N,32-2*DB)-(N,30-2*DB):NEXT N
580 REM
590 LOCATE 22,1
600 PRINT " Hit enter for Filter Coefficients"
610 INPUT Y
620 REM if y=0 then goto 680
630 CLS
640 PRINT " Filter Coefficients"
650 PRINT"(corrected for unity maximum gain)"
660 Z=M+1
670 PRINT "h(" Z ")= " H0/MAX
680 FOR N=1 TO M
690 R=(M+1)-N
700 S=(M+1)+N
710 PRINT "H(" R ")=H(" S ")= " H(N)/MAX
720 NEXT N
730 REM *************************
```

Program 1—Used with permission of John Wiley and Sons, Inc.

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

Part II: A Compact Disc for Testing HF ALE Radios

An ALE test signal can be generated in a variety of ways, as the authors describe.

David Wortendyke, NØWGC, Chris Riddle, KBØHNM and Dennis Bodson, W4PWF

Introduction: Testing Methods

Federal Standard 1045 and companion Military Standard 188-141A define sophisticated adaptive HF radios that can perform automatic link establishment (ALE) with other ALE compatible radios whose addresses are known. All ALE radios procured by the US Government must comply with certain mandatory features in these two standards. This ensures that regardless of vendor, all ALE radios will interoperate successfully. Hence the systems must be tested to determine if the radios meet the mandatory requirements of the standard.

These two government standards for HF Radio ALE have been described in open literature for several years.[1,2,3,4] The interoperability testing and performance evaluation techniques initially used by the government were complex and required the transportation of test equipment to each manufacturer's location; this was a very time consuming and expensive task.[5,6] Because the goal of the testing was to determine only the interoper-ability and performance of each unit, compliance testing against the standards was not done. The scope of the two federal standards is very broad, with a rich set of features. Both standards have something for everyone. The features are either mandatory or optional, and most radio manufacturers tend to focus only on the mandatory features to minimize the product cost.

For the initial government procurement (drug interdiction program) in 1990, the government provided a test team and equipment to perform interoperability and performance testing at six different vendor locations. Interoperability normally implies that two different systems can perform the same functional tasks, and one is considered the "standard" system. A quick review of the three-way handshaking required by the radios to establish a link is shown in Fig 1. After a preliminary checkout of two identical systems provided by a vendor, one was placed on line with an antenna, in order to test the selected

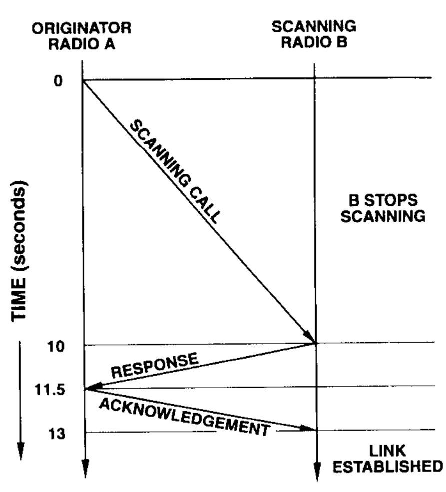

Fig 1—Three-way handshake used by ALE radios to establish a link.

[1]Notes appear on page 14.

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functions over-the-air with a system in Boulder, Colorado. While this method worked for the most part, it certainly was not a very satisfying procedure. The performance testing required the use of government-furnished equipment—including a portable, narrowband, propagation channel simulator (Watterson model). Because performance is measured between two radio systems, and the ALE technique uses the bidirectional handshake shown in Fig 1, two channel simulators were required—each costing about $60,000! A typical test set up is shown in Fig 2.

The interoperability tests using one standard radio, and the performance tests using two of the vendors' radios with two channel simulators, were not once-in-a-lifetime occurrences. As each new standard in the family of adaptive radio standards is issued, there must be at least two more sets of tests. The first is the "Proof of Concept" prior to the release of the standard, and then the "Acceptance" testing as described in note 5. After the 1990 testing, it appeared obvious that there must be a better way to perform the tests, with a technique that could reduce the transporting of equipment to each location. The remainder of this paper considers a method to solve many of the difficulties of this testing problem.

The Need for a Standard Set of Test Stimull

The need for a standard method of testing new ALE radios prompted the High Frequency Industry Association (HFIA) to organize their efforts under the Armed Forces Communications and Electronics Association (AFCEA) and propose a set of 12 types of tests which cover all the mandatory features of the two government ALE radio standards. The document was approved at their meeting on January 7, 1993, in San Diego, California. One of the features of the testing is the need for a standard test tape or audio compact disc (CD) with the various types of calls, emulating an ALE radio, encoded as ALE modem tones on the tape or CD. Table 1 lists the 12 types of tests as itemized in the document that was approved. The first 11 tests may fit on one or two CDs, but the 12th test for performance could require a dozen CDs if fully implemented. Each of the 12 types of tests will have multiple calls placed by the emulated radio recording. Generally, these calls will use different combinations of addresses for the calling tape and responding radio. Except for two cases, all the tones will be at the required level (0 dBm into 600 ohms) to be considered a "clean tone" (599). The 12th test measures the ability of the radio to link in a disturbed environment with noise, multipath and fading. Table 2 presents the various channel conditions required by the government standards for measurement of performance, established by how many links will occur with 100 attempted calls. The clean tones are used to determine the interoperability

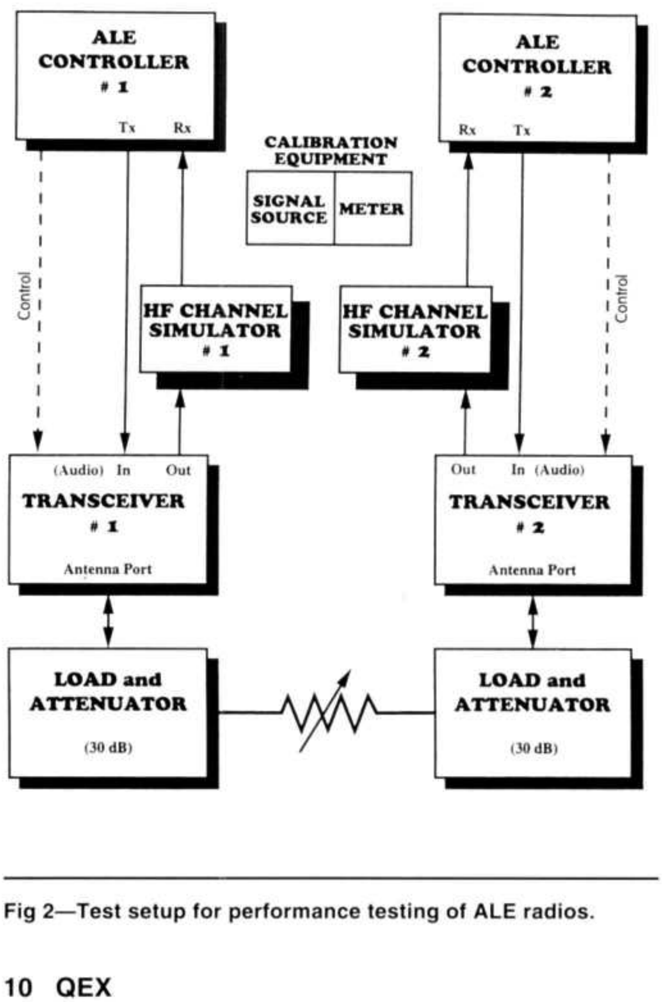

Fig 2—Test setup for performance testing of ALE radios.

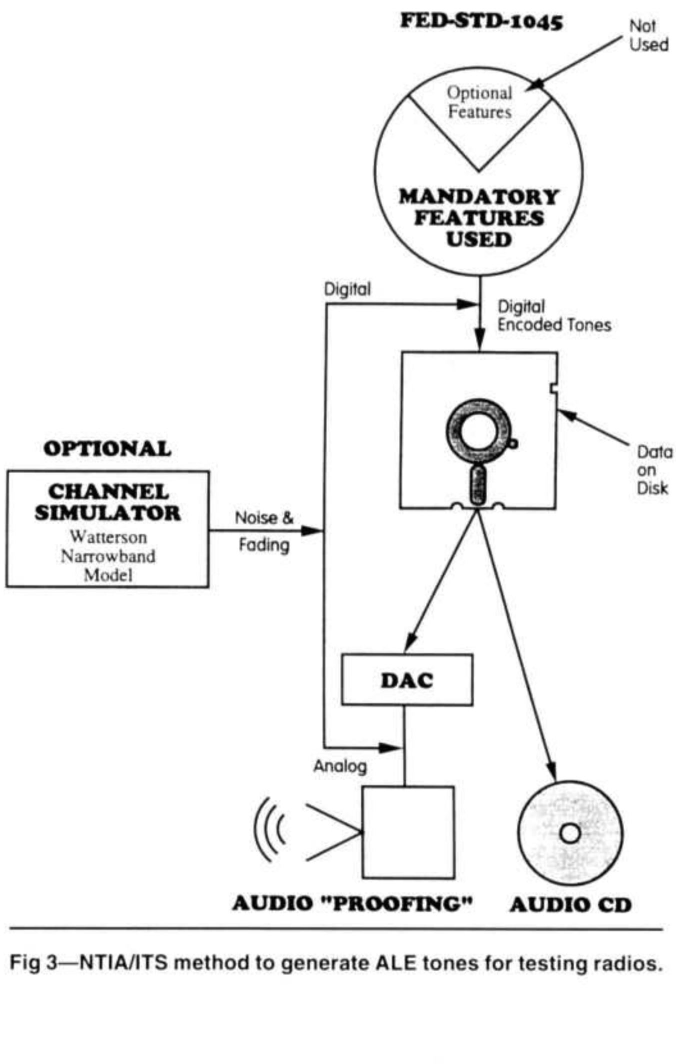

Fig 3—NTIA/ITS method to generate ALE tones for testing radios.

of the tested radio, while the disturbed tones will be used to test the performance of the radios in adverse conditions.

One of the questions that arises when using a one-direction or one-way test technique like this is: What happens to the value of the probability of linking? The subjective answer is easy: The probability of linking increases over the two-direction, three handshake linkup testing by some constant value. The theoretical value has not been solved at this time. Any theoretical values should obviously be verified by empirical data from several brands of radios. This computed/measured factor could be used as guideline to extrapolate the expected two-direction, three-way handshaking performance, but not in place of the actual performance required by the government standards for ALE radios.

Why an Audio CD?

There are many methods that could be used for storing and reproducing the ALE audio tones. These are:

1. audio cassette tape and player,
2. digital audio tape (DAT) and player,
3. audio compact disc (CD) and player,
4. computer sound file and sound card.

The first two are sequential techniques, while the latter two use random access for playback. For testing, especially repetitive testing, the random-access device has advantages over the tape. A good analog audio tape cassette deck may have sufficient accuracy to do some of the testing, but because of wow and flutter it should be looked at with suspicion. The DAT and CD, being digital devices, when used in a good quality home system can reproduce tones with an accuracy on the order of 5 to 30 parts per million (ppm). The analog cassette tapes may fall in the range of 1500 to 4000 ppm accuracy. The computer sound-card alternative is viable if the sound card provides sufficient accuracy. The first generation of sound cards that were available in 1992 were very dependent on the host processor chip and speed. Even with an 80386/486 at 33 MHz, the measured frequency accuracy was only slightly better than the better cassette tape decks. New-generation sound cards using digital signal processing (DSP) can achieve accuracies close to that of the CD player.[7] That type of accuracy may allow their use in the ALE radio tests.

The CD has a major feature that can prove advantageous in some test situations, and a detriment in others. The CD is the only medium that can not be altered. A great amount of security is provided to the entire HF industry and government agencies by using a common test source. The disadvantage is that if it is necessary to design a test on-the-fly with different addresses, etc, the other media can be altered; the CD cannot. The most flexible medium is the PC sound card, since the parameters may be modified and a new set of tones computed. This will be explained in more detail in the next section.

Several important factors remain, regardless of which medium is used, that provide motivation for producing a test tape or CD. It would:

1. meet the requirements of the HFIA test plan,
2. reduce the test equipment needed to verify interoperability,
3. eliminate the need for an expensive channel simulator for initial performance tests,
4. clarify, by example, several parts of the standard that have been misunderstood,
5. promote more widespread use of the government standards,
6. serve as a method to transfer technology from the government to smaller companies desiring to enter the market,
7. become the spring-board for future automatic testing of new features and next-generation government standards for ALE radios.

Table 1 HFIA Suggested Format of Standard Test Tape or CD.

Section	*Type of Call*	*Comments*
1	Sounds	Many addresses using all valid characters in the allowed set Noise added at three S/N levels for a selected address
2	100 Quick Sounds	Mix of long and short addresses
3	All Call	Simple and with AMD
4	Selective All Call	With known and foreign address
5	Individual Calling	With four levels of LQA in ackn.
6	Net Calls	Simple and with LQA and AMD
7	Group Calls	Same as above
8	Any Calls	Same as above
9	Selective Any Calls	With known and foreign address
10	Wild Card Calls	With length = 3 and 6 characters
11	Automatic Message Display (AMD)	90 Character message in Response of Call
12	Performance Tests	100 calls for each of the 12 sets with All Call & AMD of conditions in FED-STD-1045 matrix

Computer Generation of ALE Tones

Various methods were considered to generate clean tones for interoperability testing. All the modem/radio manufacturers use digital signal processing (DSP) boards with a development effort typically measured in man-years. Furthermore, all their code is proprietary and therefore not available. In spite of the industry's careful efforts to implement the standards, several operational differences were found in the initial tests between vendors (see note 5). We did not want to reinvent the wheel in our effort to make a standard audio test tape or CD, and we had a goal of doing the development in man-months, not man-years. It was apparent that a complete audit trail of input protocol, encoded tone values, and, of course, the output audio tones, would be mandatory for a standard test tape. Since we started the task in October, 1992, rapid progress has been made in both design and implementation, with about 3 man-months completed.

This paper focuses on the method used to generate the clean tones for the first 11 types of tests in Table 1. The subject of generating the tones through a disturbed set of channel conditions is beyond the scope of this paper and would not satisfy the intended purpose of this series of articles on ALE radios. In very general terms, as shown in Fig 3, the mandatory features of the ALE radio standards are used to generate a set of sound files which are stored on (magnetic) disk or tape. The actual format of the files depends on how the data is to be converted into audio. One format is the WAVE file (technically a RIFF file with sound data) defined by Microsoft and used in the *MS-DOS / Windows* 3.1 system. Any of the better-quality plug-in audio sound cards can play the files to external speakers. An extremely important point in this technique is that the files are made off-line, placed in a RAM drive on the computer, and then played from the RAM disk. It is *not* a real-time process; rather it is a "compute now, play later" philosophy. It is wise to use a 80386 or 80486 processor with a math coprocessor when doing the calculations. Depending on the number of bits and sampling speed, the computations can take 1 to 4 times the playback time.

Table 2 Government-required performance tests for ALE radios FED-STD-1045 Probability of Liking

Signal-to-Noise Ratio (dB) in 3 kHz BW			*Required Probability of Linking with 100 Attempts*
Gaussian Noise Channel	*CCIR Good Channel*	*CCIR Poor Channel*	
–2.5	+0.5	+1.0	≥ 25%
–1.5	+2.5	+3.0	≥ 50%
–0.5	+5.5	+6.0	≥ 85%
0.0	+8.5	+11.0	≥ 95%

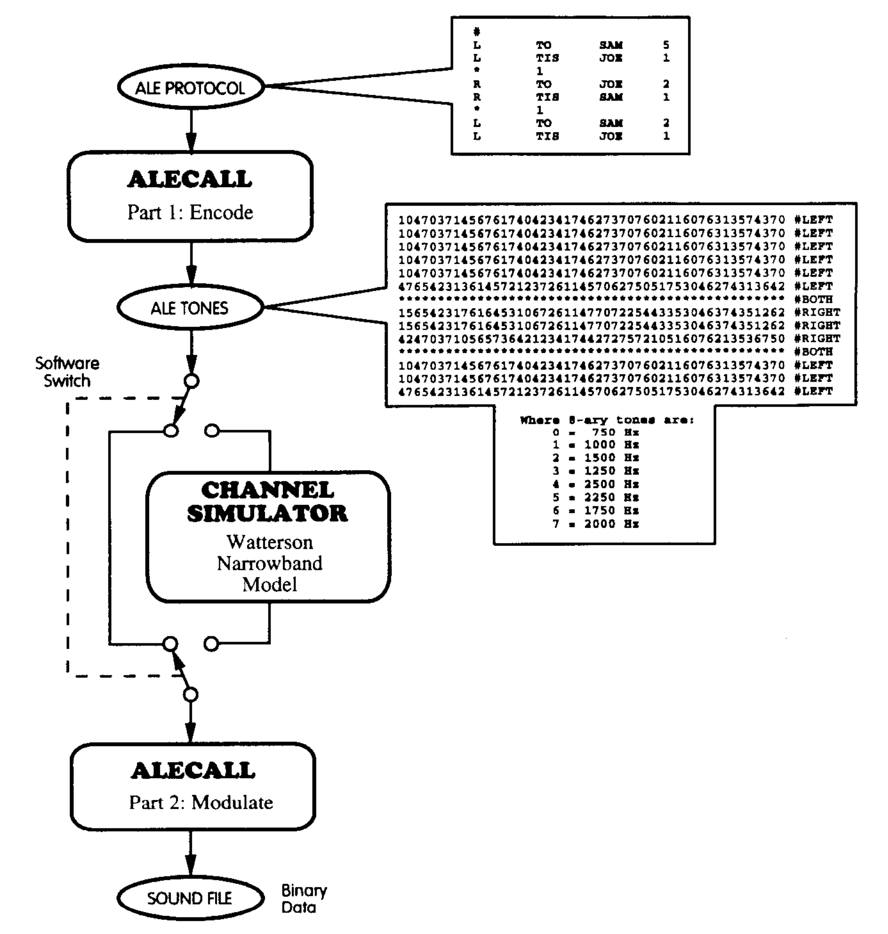

Fig 4—Overview of process to make a datafile of ALE modem tones.

The protocol defined in the Government ALE radio standards names 8 types of preambles to create an ALE call. Two types will be mentioned in this paper for the purpose of illustrating the technique of encoding the message. Fig 4 presents the overview of the method used to make the sound files. For simple ALE calls, a single program, ALECALL, reads an ASCII text protocol file of preambles and addresses and converts the information into the 8 tones used by the standards. These tones are then digitized in the modulator section of the ALECALL software program and written to the computer disk as a sound file. Although it would not be necessary to output the tones in numeric form, we chose to write them to a disk file as part of the audit trail for the user of the sound files. The two preambles in the example in Fig 4 are "TO" and "TIS" (TIS = THIS IS), which represent the addressee (TO) and the sender (TIS). Four single-character letters may be used in the first column of the input protocol file:

1. the "#" defines a comment line for informational purposes,
2. "L" or "R" directs the data on that line to the left or right speaker,
3. an "*" causes a delay of 49 bauds or 392 ms, since each symbol is 8 ms.

The preamble type for left and right audio channels is the second word of the line, followed by 3 characters of data or address. The last number in the line is the number of times to repeat the data for that line. The example protocol file begins on the left channel with 5 sets of "TO SAM" followed by one "THIS IS JOE" for the initial call. The response on the right channel from SAM is 2 sets of "TO JOE" followed by one "THIS IS SAM." The final acknowledgement by JOE is similar to the response. This type of file documents without any question the input to the encoder software. It also makes it possible to generate a stereo sound file with the originating radio on one channel (left in this example) and the radio being tested on the other channel. In actual testing, the right channel data would not be used, except as reference if desired.

The ALE standards define the modem tones as an 8-ary system with tones from 750 to 2500 Hz in steps of 250 Hz. The sequence is gray-scaled or mixed to help prevent selective fading destruction of a particular range of tones. Each tone may be represented by an octal number from 0 to 7 as shown in Fig 4. The ALE system uses 24 bits of data from each line in the protocol file and encodes these bits, along with a stuff bit, three times with interleaving into a 49-tone word. Each tone (baud) is on for 8 ms, so the entire data word is 392 ms. The sample tone file in Fig 4 shows 49 octal tone values per record, with the comment following them for the right or left channel. The time delays created by the "*" apply to both channels.

If the data is to be used for a DAT or CD, the choice of the sampling rate is the standard CD rate of 44,100 samples per second. Using a 16-bit digital-to-analog converter (DAC), and stereo for two-channel data, the example shown here is exactly 5.488 s in playing time, but took 9.6 s to generate and is slightly less than 1 Mbyte in size. We plan to use the computer model channel simulator shown in Fig 4 to create the disturbed tones that actual propagation conditions would cause. This portion of the software is still under development and is not discussed in this paper.

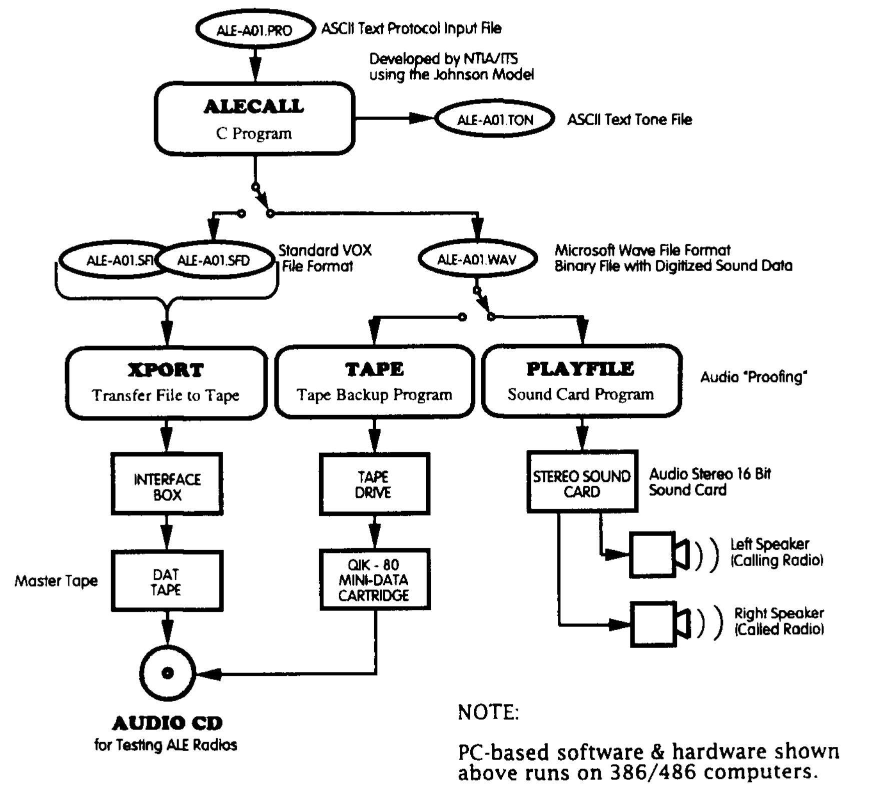

Fig 5—Emulation of ALE radio tone generation for laboratory testing.

Putting it Together

This section describes the complete technique used for making the audio data file for the DAT or CD, and use of the tones to actually place a call and link with a real ALE radio. The process consists of four steps:

1. create an ASCII text protocol file describing the ALE call (usually the 3-way handshake),
2. run the ALECALL program with the appropriate software switch to build a sound file and, if using the DAT, transfer the file to the tape,
3. set up a transceiver (with voice actuated keying) on an authorized frequency. The sound card (or CD or DAT) output should be plugged into the microphone jack of the transceiver, with proper impedance matching and levels to key the transmitter,
4. while the real ALE radio is operating (and scanning its pre-programmed frequencies), play the file of ALE tones into your transceiver. The ALE radio should stop scanning, respond, answer your call, and think it is linked with another ALE radio.

The diagram in Fig 5 shows steps 2 and 4. It will be assumed that the readers of this article will have equipment to perform step 3, or could improvise an equivalent with a balanced mixer and an RF amplifier. The simplest of calls, such as the type described in the previous section, use an ASCII text protocol which may be typed into the computer with any text editor. That satisfies step 1. More complex calls, such as a request for link quality analysis (LQA) score, require a preprocessor program to convert what we call a Level 1 protocol file (user-friendly English) into an ASCII Level 2 protocol file. The Level 2 file may include non-printable characters which are expressed by a printable combination of ASCII characters. An example is the SHIFT OUT character (decimal 14 = CONTROL N) which is written as the doublet "^N." This coding provides a powerful technique for creating the protocol files for a large number of the mandatory features in the Government ALE standards.

The ALECALL software used in step 2 will currently generate either a sound file of audio data (a format used by CCITT as a digitized voice file) with the ".SFD" extension or the WAVE file RIFF format with the ".WAV" file extension. The run-time choice depends on which method in step 4 will be used to reproduce the digital data into analog.

While the first generation of sound cards using DMA appear to be inadequate for precise audio generation of the ALE tones, the new generation of sound cards, with both a microcomputer and DSP chip on board, appear to work acceptably for the interoperability tests.

Summary

This paper describes the need for a simple standardized method of testing the ALE radios as they are developed and new features are added. The paper also presents a relatively simple and inexpensive technique for creating digitized data for a set of test tones using a personal computer and special software. The audio tones may be played by using a good quality audio sound card, or the digital data can be transferred to DAT (tape) and/or audio CD and played through a good quality DAT or audio CD player. Each file of digitized tones will exercise or test a different function of the ALE controller and radio. While the immediate goal of this effort is testing ALE radios, it would be a rather simple matter to use the same technique for testing many types of modems and related audio applications. The rapid turn-around time to modify an address and generate an actual call using the sound card is ideal for experimenters and for laboratory work. The reduction in test equipment and inventory of radios for preliminary testing of new features in the ALE controller is appreciable. Automation of repetitive tests controlled by a computer may be a simple matter when using the sound card which is already installed in the computer. On the other hand, using an inexpensive compact disc provides all vendors and customers with exactly the same unchangeable source of audio tones to test the operation (and performance) of new or existing systems.

A potential application for Amateur Radio operators could be in conjunction with passing messages into an operational ALE network. Assume that the MARS network had ALE repeaters, and an amateur had the ALE software and a sound card. The ham could call the repeater, form a link, and then pass traffic with whatever modems were normally used to exchange data. If you can think of other applications, the authors would appreciate comments sent to NTIA.

Acknowledgements

This work was supported by the National Communications System (NCS) and the National Telecommunications and Information Administration/Institute for Telecommunication Sciences (NTIA/ITS). Credit is given to Dr. Eric Johnson at New Mexico State University, Electrical Engineering Department, for the development of the encoding software while under contract to NTIA.

Notes

[1] FED-STD-1045, "Telecommunications: HF Radio Automatic Link Establishment," January 24, 1990.

[2] MIL-STD-188-141A, "Interoperability and Performance Standards for Medium and High Frequency Radio Equipment," September 15, 1988.

[3] Adair, Robert T. and Peach, David F., "A Federal Standard for HF Radio Automatic Link Establishment," *QEX*, January, 1990, pp 3-7.

[4] Prisutti, Matthew J. and Smullen, Helen, "Defense Information System HF Entry - A Look to the Future," *Proceedings, 1992 IEEE Military Communications Conference*, November, 1992.

[5] Smith, Paul C., Wortendyke, David R., Redding, Christopher, and Ingram, William, "Interoperability Testing of FED-STD-1045 HF Radios," *Proceedings: RF Expo West*, February 5, 1991, pp 119-126.

[6] Wortendyke, David R. and Riddle, Chris C., "Adaptive HF Radio Performance Evaluation Using Automated Instrumentation," *Proceedings, 1991 IEEE Military Communications Conference*, November 4, 1991, pp 49.2.1-49.2.7.

[7] Kendall, Robert, "Sound Boards, Coming of Age," *PC Magazine*, April 27, 1993, Volume 12, Number 8, pp 181-234. □□

Broadband Dipole and Unipole 3.5-to-4.0-MHz Antennas

Using only transmission lines, the author shows how 75/80-meter dipoles can be "stretched" to cover wider bandwidths.

Hugh Fallis, N4HF

This article discusses the design of a dipole antenna that will present a standing wave ratio (SWR) of no greater than 1.7:1 over a frequency range of 3.5 to 4.0 MHz. The NEC2 program is used to calculate two dipole configurations: a standard dipole and a folded dipole. NEC2 is an acronym which stands for Numerical Electromagnetic Code, second version. NEC is a powerful computer program developed for calculating the performance of antennas.

To achieve such a wide bandwidth, it is necessary to employ some means of improving the change in input impedance versus frequency for a dipole. My goal is to improve the dipole antenna bandwidth using only one quarter-wave shorted stub and one quarter-wave transmission line transformer. The stub and transformer are a quarter wave long only at the center frequency, eg, 3.750 MHz for the 80/75-meter band. If the line is 90 degrees long at 3.750 MHz, then the length will be 84 degrees at 3.500 MHz and 96 degrees at 4.000 MHz. At these lengths, the quarter wave transformer is still effective in transforming impedances.

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During my 40 years as an active ham operator, I have had a great interest in antennas. In my early years, I became an avid DXer and quickly learned the importance of a good antenna for successful DXing. I have been associated with Radio Free Europe and Radio Liberty for the past 28 years and have had the unique opportunity of being involved with the design and construction of some of the highest-gain shortwave antennas ever built. Working with shortwave antennas professionally has allowed me to gain insight into improving the frequency bandwidth performance of dipole radiators. I have made use of this knowledge in the design of a broadband dipole radiator for 80/75 meters.

The first step in my design is to calculate the dimensions of the antenna so that it resonates at midband. The second step is to calculate the input impedance of a dipole antenna in free space. Calculations using NEC2 showed that a dipole constructed with number 14 copper wire will be resonant in free space if its length is 38.91 meters. This dipole is shown in Fig 1. The definition of resonance for a dipole in this case is when the reactive component of the input impedance is zero, eg, $Z = R + j0$. Table 1 lists the input impedance of this dipole and the input SWR (relative to 75 Ω) versus frequency.

In Table 1, the useful frequency range of this dipole falls somewhere between 3.650 and 3.850 MHz without exceeding 1.7:1. The frequency range of

Table 1—Input Impedance Vs Frequency of a Dipole

Frequency (MHz)	*Dipole Input Impedance (Ω)*	*SWR*
3.500	61.02 −j111.29	4.5
3.525	62.11 −j100.04	3.9
3.550	63.21 −j88.84	3.4
3.575	64.33 −j77.70	2.9
3.600	65.46 −j66.60	2.5
3.625	66.61 −j55.54	2.2
3.650	67.77 −j44.54	1.9
3.675	68.96 −j33.57	1.6
3.700	70.16 −j22.64	1.4
3.725	71.38 −j11.75	1.2
3.750	72.61 −j9.06	1.04
3.775	73.87 +j9.91	1.1
3.800	75.13 +j20.69	1.3
3.825	76.43 +j31.44	1.5
3.850	77.73 +j42.16	1.7
3.875	79.06 +j52.85	2.0
3.900	80.41 +j63.53	2.2
3.925	81.77 +j74.16	2.5
3.950	83.16 +j84.78	2.8
3.975	84.56 +j95.38	3.1
4.000	85.99 +j105.95	3.8

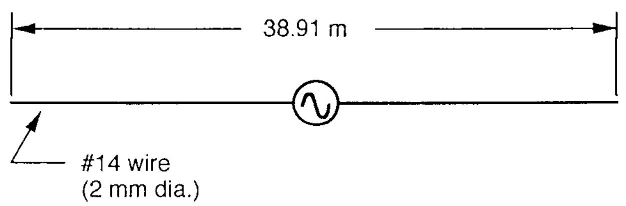

Fig 1—80/75-meter dipole antenna.

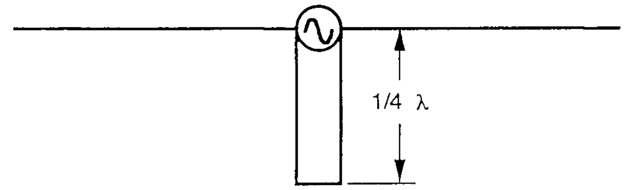

Fig 2—Dipole with a quarter-wave shorted stub.

the dipole antenna can be increased by paralleling a shorted stub at the input of the dipole. The input admittance of this dipole at 3.500 MHz is 0.003788 + j0.006909 S, and the input admittance at 4.0 MHz is 0.004618 – j0.005690 S. The formula for the admittance of a shorted-stub transmission line is:

$$Y = -j\, G_0 \cot \theta \qquad (1)$$

Theta, θ, equals the electrical length of the stub in degrees. G_0 is the stub's characteristic admittance ($1/Z_0$). In this example, I assume that at center, 3.750 MHz, the shorted stub has an electrical length of 90°, eg, the stub is 20 meters in length. At 4.0 MHz the stub will have an electrical length of 96°, eg, (4.0/3.75) × 90° = 96°. At 4.0 MHz, I would like Y in Eq 1 to equal +j 0.00569 S. This means that the stub would cancel out the dipole's reactance at 4.0 MHz. With this information, it is now possible to calculate the required value of G_0, the characteristic admittance of the transmission line used for the stub. Using algebra, Eq 1 can be rearranged to solve for G_0, the stub characteristic admittance.

$$G_0 = j\, Y \tan \theta \qquad (1a)$$

Solving Eq 1a shows that G_0 = 0.054 S, with its reciprocal in ohms (impedance) being 18.5. Of course, it is not practical to construct a 20-meter length of 18.5-Ω balanced, open-wire transmission line. Nevertheless, we shall still analyze the effect on SWR by adding this stub as shown in Fig 2.

I have developed computer programs for converting impedances to admittances and admittances to impedances, as well as computer programs for calculating the input impedance or admittance of a transmission terminated with a load. I wrote these programs in FORTRAN, which is equipped to work with complex numbers. A Smith chart can also be used to make such calculations, however, this chart can become very cumbersome if many calculations have to be made. Table 2 lists the results of adding a 20-meter stub with the characteristic impedance of 18.5 Ω.

By comparing Table 1 with Table 2, it is obvious that adding a shorted quarter-wave stub at the input of the dipole did not significantly increase the bandwidth of this antenna. It is also interesting to note that at the band edges, the input resistance increased and the reactance decreased.

A variation of a dipole that will improve its input impedance over a range of frequencies is the folded dipole. Fig 3 shows the dimensions of the folded dipole antenna being considered to provide a SWR of less than 1.7 from 3.5 to 4.0 MHz. Table 3 lists the free space performance of the folded dipole as calculated by the NEC2 code. The SWR is relative to 300 Ω.

A comparison between Tables 1 and 3 shows that the folded dipole provides at least 50 kHz more bandwidth than a standard dipole. Using the same procedure for calculating a stub, placed in parallel with the antenna input, a value of 68.26 Ω is achieved. However, since 68 Ω is close to the impedance of 75-Ω twin lead, I have decided to use 75 Ω for the characteristic impedance of the quarter-wave shorted stub.

Table 4 shows the resulting input impedance with the stub and the input SWR relative to 300 Ω for the folded dipole.

Table 2—Input Impedance Vs Frequency, Dipole with Matching Stub

Frequency (MHz)	*Dipole Input Impedance (Ω)*		*SWR*
3.500	239.24	–j76.96	3.6
3.525	183.10	–j85.75	3.1
3.550	144.34	–j79.46	2.7
3.575	118.46	–j68.59	2.3
3.600	101.10	–j56.98	2.0
3.625	89.37	–j45.88	1.8
3.650	81.49	–j35.60	1.6
3.675	76.39	–j26.10	1.4
3.700	73.42	–j17.25	1.3
3.725	72.22	–j8.89	1.1
3.750	72.61	–j0.91	1.03
3.775	74.57	+j6.82	1.1
3.800	78.20	+j14.37	1.2
3.825	83.72	+j21.72	1.3
3.850	91.57	+j28.76	1.5
3.875	102.29	+j35.19	1.7
3.900	116.69	+j40.29	1.8
3.925	135.62	+j42.64	2.1
3.950	159.72	+j39.70	2.3
3.975	188.20	+j27.21	2.6
4.000	216.54	+j0.025	2.9

Table 3—Input Impedance Vs Frequency of a Folded Dipole

Frequency (MHz)	*Folded Dipole Input Impedance (Ω)*		*SWR*
3.500	279.89	–j324.73	2.9
3.525	279.27	–j288.67	2.6
3.550	279.36	–j253.57	2.3
3.575	279.00	–j219.46	2.1
3.600	279.92	–j186.16	1.9
3.625	279.93	–j153.61	1.7
3.650	281.10	–j121.49	1.5
3.675	282.01	–j90.52	1.4
3.700	283.45	–j59.85	1.2
3.725	285.17	–j28.70	1.1
3.750	287.16	–j0.00	1.04
3.775	289.41	+j29.28	1.1
3.800	291.93	+j58.19	1.2
3.825	294.72	+j86.79	1.3
3.850	297.79	+j115.11	1.5
3.875	301.12	+j143.20	1.6
3.900	304.71	+j171.04	1.8
3.925	308.63	+j198.80	1.9
3.950	312.84	+j226.40	2.1
3.975	317.34	+j253.91	2.2
4.000	322.15	+j281.37	2.4

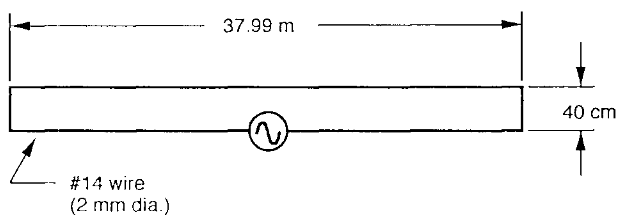

Fig 3—Folded dipole antenna.

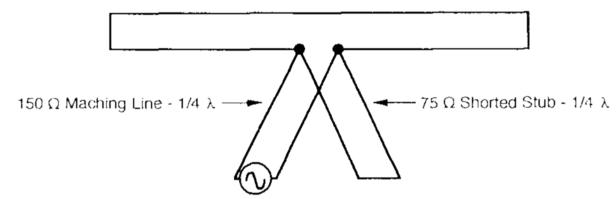

Fig 4—Matching arrangement for the 80/75-meter folded dipole.

A comparison of Tables 1, 2, 3 and 4 shows the following:

1. The bandwidth of the uncompensated dipole is approximately 185 kHz. The bandwidth of the uncompensated *folded* dipole is approximately 240 kHz.
2. The use of a quarter-wave shorted stub connected in parallel with the antenna input results in a band width of approximately 260 kHz for a dipole and 400 kHz for the folded dipole.

This comparison demonstrates the inherent wider bandwidth of the folded dipole.

My experience in matching broad-band antennas has taught me that in many cases, the use of a quarter-wave transmission line to transform a higher impedance to a lower impedance will result in a lower SWR across the desired bandwidth of the antenna. In most cases, the greater the ratio between the antenna impedance and the desired load impedance, the more effective the reduction in SWR. Most amateurs are familiar with the equation for determining the characteristic impedance (characteristic resistance since we assume the transmission line has no loss). The equation is as follows:

$$Z_O = \sqrt{Z_A \times Z_I} \qquad (2)$$

Where:

Z_A equals the antenna impedance.
Z_I equals the desired input impedance.
Z_O equals the characteristic impedance of the quarter wave transmission line transformer.

A comparison of Tables 2 and 4 show that the folded dipole is a better candidate for the use of a quarter-wave matching line than a standard dipole.

In applying formula 2 to calculate the required characteristic impedance of the quarter wave transformer, I selected the desired Z_I to equal 50.0 + j0.0 Ω and used 300.0 +j0.0 Ω for Z_A. Therefore, Z_O will equal 122.47 + j0 Ω. For practical considerations, I will use 150 Ω for the characteristic impedance of the quarter-wave matching section. This line is connected in parallel with the folded dipole input which still has the matching stub connected. Fig 4 shows this configuration. Table 5 shows the resulting input impedance and SWR relative to 50 Ω at the transformer input.

The use of the stub and the transformer in conjunction with a folded dipole antenna provides an antenna that exceeds our design goal. The maximum SWR is 1.51:1 between 3.5 and 4.0 MHz. Only low-loss matching elements have been used to achieve this wide bandwidth.

At this point, it should be emphasized that this is only a theoretical design. I have not built such an antenna because I don't have the room to do so, but what is important is the design approach.

In the next part of this article I shall use this broadbanding technique on a folded dipole antenna over real ground. I shall also demonstrate the effect antenna height above ground has on the antenna input impedance and its resonant frequency. I calculated, using NEC2, the input impedance of a folded dipole at ⅛, ¼, ½, ¾ and 1 wavelength above a perfect reflecting surface, and above average ground. (Average ground is defined as having a relative dielectric constant of 10 and a conduc-

Table 4—Input Impedance Vs Frequency, Folded Dipole with Matching Stub

Frequency (MHz)	*Folded Dipole Input Impedance (Ω)*		*SWR*
3.500	620.66	–j148.99	2.2
3.525	528.31	–j161.44	2.0
3.550	457.51	–j154.26	1.8
3.575	403.74	–j139.05	1.6
3.600	364.19	–j119.48	1.5
3.625	335.20	–j98.62	1.4
3.650	314.61	–j77.32	1.3
3.675	300.15	–j57.22	1.2
3.700	291.29	–j37.42	1.1
3.725	287.09	–j18.34	1.08
3.750	287.16	–j0.00	1.04
3.775	291.31	+j17.58	1.07
3.800	299.60	+j34.32	1.1
3.825	312.26	+j50.05	1.2
3.850	329.71	+j64.39	1.3
3.875	352.53	+j76.70	1.3
3.900	381.38	+j85.88	1.4
3.925	417.14	+j90.33	1.5
3.950	460.08	+j87.41	1.6
3.975	509.91	+j73.77	1.8
4.000	564.50	+j43.77	1.9

Table 5—Input Impedance Vs Frequency, Folded Dipole of Fig 4

Frequency (MHz)	*Folded Dipole Input Impedance (Ω)*		*SWR*
3.500	34.23	–j6.68	1.51
3.525	38.67	–j1.32	1.29
3.550	43.71	+j3.35	1.16
3.575	49.23	+j7.28	1.16
3.600	55.12	+j10.11	1.24
3.625	61.14	+j11.56	1.34
3.650	66.92	+j11.50	1.42
3.675	71.97	+j10.14	1.49
3.700	75.81	+j7.41	1.54
3.725	78.00	+j3.84	1.57
3.750	78.35	+j0.00	1.57
3.775	76.91	–j3.48	1.54
3.800	73.98	–j6.11	1.50
3.825	69.97	–j7.56	1.43
3.850	65.36	–j7.72	1.35
3.875	60.52	–j6.65	1.25
3.900	55.74	–j4.49	1.15
3.925	51.21	–j1.42	1.04
3.950	47.03	+j2.37	1.08
3.975	43.24	+j6.75	1.23
4.000	39.83	+j11.56	1.41

Table 6—Lengths of Resonant Folded Dipole Vs Height

λ	Dipole Lengths in Meters and (Feet)	
	Perfect Reflector	*Avg Earth*
1/8	37.078 (121.6)	37.071 (121.6)
1/4	37.010 (121.4)	37.271 (122.3)
1/2	38.618 (126.7)	38.458 (126.2)
3/4	37.578 (123.4)	37.676 (123.6)
1	38.312 (125.7)	38.240 (125.5)

tivity of 0.01 S /meter.) At each height above the reflecting plane (ground) it is necessary to readjust the length of the folded dipole to make the antenna resonant at 3.7500 MHz.

Table 6 lists the required lengths for resonant folded dipoles at each of the above mentioned heights above a perfect reflecting surface and average ground.

A review of Table 6 indicates that the properties of the reflecting plane and the antenna height above the reflecting plane do affect the resonant frequency of an antenna. This information is useful in the construction and testing of antennas. For example, assume that the dimension of the antenna length was adjusted for resonance while ⅛ wavelength above ground. If the antenna were raised to its operating height of ½ wavelength, its resonant frequency would be 140 kHz higher in frequency in the case of an 80-meter antenna. This would be rather troublesome since it would mean that you would have to lower your antenna and make it longer! Lawson mentions this effect in his article on Yagi antennas (Lawson, J.L. "Yagi Antenna Design," *Ham Radio*, Jan, Feb, May, June and July, 1980).

By placing an antenna at a height h above a perfect reflector, the effect on the input impedance of the antenna would be the same in free space if a second antenna were placed at a distance $2h$ from the first. The second antenna would have a current distribution 180 degrees out of phase with the first. The second antenna is called the image antenna. In fact, one can draw the analogy of placing a pen a distance h above a flat mirror and that of placing an antenna a distance h above ground. The image will appear to be a distance h below the surface of the mirror or ground.

In the case of placing the antenna ⅛ wavelength above ground, the coupling between the antenna and its image is such that it raises the effective Q of the folded dipole so that it can no longer be broadbanded using one stub and one quarter-wave transformer. Since that was my design goal, no broadbanding data will be given for the ⅛-wave case. However, an alternative arrangement of using a broadband ¼-wave vertical folded dipole antenna will be discussed.

Table 7 lists the calculated input SWR, relative to 50 Ω, for the folded dipoles listed in Table 6. The identical matching system for matching the folded dipole in free space is used for matching these dipoles. The broadbanding results are quite good.

In the case of the horizontal folded dipole, the antenna must be at least a quarter wave above ground (66 feet for 80 m and 132 for 160 m). This requirement significantly reduces the number

Table 7—SWR Vs Frequency for Folded Dipoles

Frequency (MHz)	*Ant Height Above Ground in Wavelengths*			
	1.0	*3/4*	*1/2*	*1/4*
3.500	1.49	1.52	1.51	1.55
3.525	1.28	1.31	1.30	1.33
3.550	1.14	1.18	1.17	1.16
3.575	1.12	1.18	1.18	1.05
3.600	1.21	1.25	1.27	1.08
3.625	1.31	1.34	1.38	1.17
3.650	1.41	1.41	1.48	1.24
3.675	1.49	1.47	1.57	1.30
3.700	1.55	1.51	1.63	1.33
3.725	1.59	1.52	1.67	1.35
3.750	1.61	1.51	1.68	1.34
3.775	1.59	1.48	1.67	1.32
3.800	1.57	1.43	1.63	1.28
3.825	1.50	1.36	1.56	1.23
3.850	1.42	1.28	1.48	1.16
3.875	1.50	1.19	1.38	1.10
3.900	1.60	1.10	1,27	1.10
3.925	1.11	1.05	1.16	1.15
3.950	1.09	1.15	1.10	1.26
3.975	1.22	1.29	1.20	1.41
4.000	1.41	1.47	1.37	1.60

of amateurs that can take advantage of the broadbanding scheme described in this article. However, one configuration of a quarter-wave vertical antenna that can be broadbanded using a similar approach is a folded unipole.

A vertical radiator configured as a folded unipole is depicted in Fig 5. The folded unipole model I have chosen to use consists of an aluminum tower insulated from ground with two AWG 14

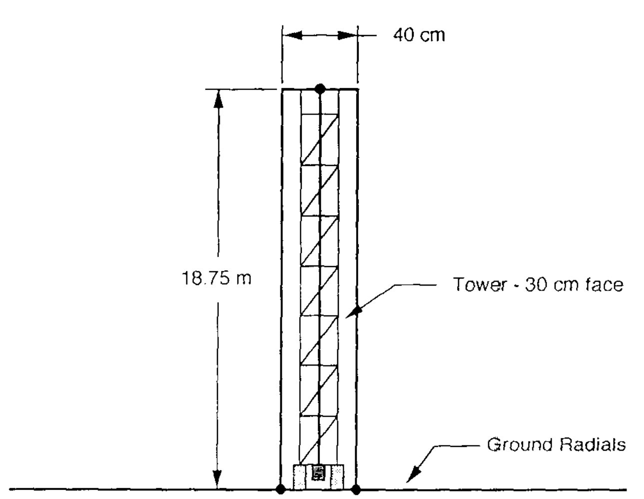

Fig 5—Folded vertical antenna.

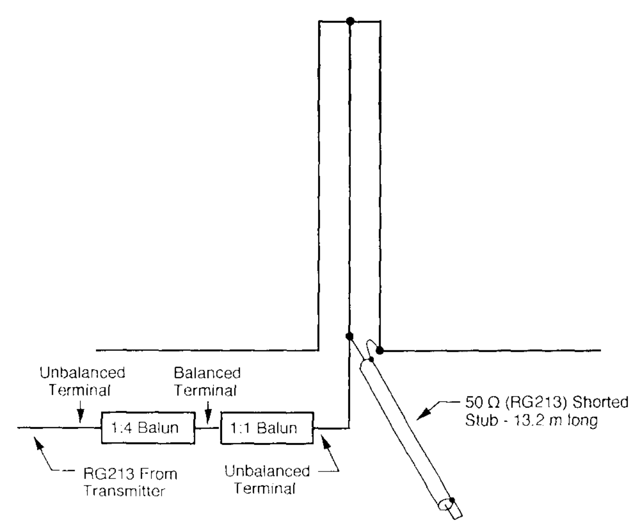

Fig 6—Schematic diagram of the folded vertical antenna.

wires connected to an aluminum outrigger at the top of the tower. These two wires go parallel down the tower to ground and then are connected to the radial ground system. In other words, this configuration is half of a three-wire folded dipole. Three wires are necessary to arrive at an average input impedance of 200 Ω so that a reasonable broadband matching arrangement can be realized. Since the input impedance of the three wire vertical unipole is relatively high, a ground system consisting of 20 half-wave radials would be ideal, but fewer and shorter radials should still give adequate performance.

Fig 6 is a schematic of the matching system. Table 8 lists the input impedance and SWR, relative to 50 Ω, of the folded vertical unipole with the broadband matching system in place.

The theoretical broadband matching is outstanding for this folded unipole. The maximum input SWR is 1.63 and the average SWR is below 1.25. This same unipole dimensioned for 160 meters will provide an even better match since the bandwidth of the top band is less than that of the 80/75-meter bands. I chose the 80/75-meter band for illustrating this broadbanding scheme because in the US this amateur band has the widest bandwidth, 14.3%.

One might ask why go to all this trouble when one can feed a dipole with an open-wire line and use an antenna tuner to match the transmitter to the antenna. After all, an antenna need not be resonant to radiate. My response is that it is not always practical to bring an open-wire line from the antenna to the shack. If changes in the direction of an open-wire line from the antenna to the shack are necessary, these changes in direction must be carefully done, for if they are not, the line will radiate, which can increase local interference. In some cases, the SWR on the open-wire line can become very high, and if the line is long, losses may be high. Normally, when the SWR is very high, a small change in operating frequency requires readjusting the antenna tuner.

I would appreciate hearing from those of you who apply this design approach to broadbanding their 80/75- or 160-meter antennas. I believe this approach is best suited for the folded unipole vertical. The matching system is much simpler, and all the components are readily available.

Finally, I would like to thank W4SG and W6MZ for their review of and suggested revisions to this article. □□

Table 8—Input Impedance Vs Frequency of Folded Unipole

Frequency (MHz)	*Matched Input Impedance (Ω)*	*SWR*
3.500	81.38 −j3.28	1.63
3.525	72.78 −j7.45	1.48
3.550	65.42 −j9.22	1.37
3.575	59.42 −j9.51	1.28
3.600	54.67 −j8.95	1.21
3.625	50.96 −j7.91	1.17
3.650	48.12 −j6.62	1.15
3.675	46.00 −j5.22	1.14
3.700	44.51 −j3.78	1.15
3.725	43.48 −j2.35	1.16
3.750	43.01 −j0.99	1.16
3.775	42.76 −j0.30	1.17
3.800	43.20 +j1.49	1.16
3.825	43.85 +j2.55	1.15
3.850	44.87 +j3.46	1.14
3.875	46.24 +j4.17	1.12
3.900	47.97 +j4.63	1.11
3.925	50.05 +j4.78	1.10
3.950	52.46 +j4.53	1.11
3.975	55.20 +j3.78	1.13
4.000	58.17 +j2.43	1.17

Digital Communications

Harold E. Price, NK6K

Embedded Processors Made Easy

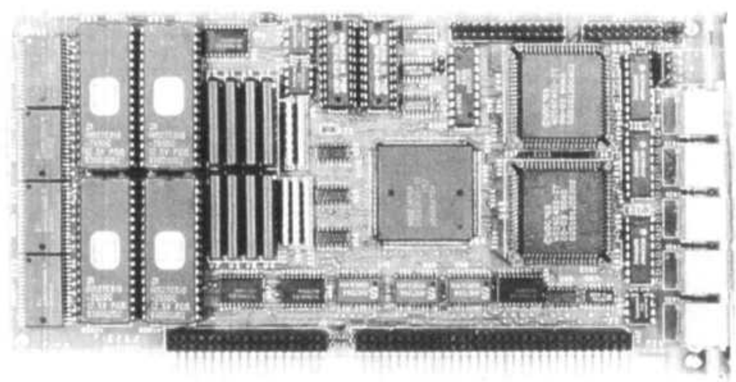

Fig 1

Last month, we tackled various esoterica like forward error correction and *Windows* Dynamic Data Exchange. This month, we go back to basics with: "How to Impress the Neighbors and Distract Them From the Unsightly Antennas on the Roof," or "Embedded Processors for People Who Are Afraid of Embedded Processors."

I was in the middle of a bull session with a group of neighbors the other night. We were watching Sean Taylor and his son Brandon testing their soapbox derby car. They were talking about making some changes to make it go faster. I opened my big mouth and said wouldn't it be neat if a computer would keep track of the speed during a run so they could see if their changes made any difference. Worse, I said I could have such a thing ready by the next night. Since it was already after midnight, "the next night" was only 17 hours away.

The point of all this is to have an excuse to discuss embedded processors. I suppose there is a formal definition of "embedded processor" somewhere, but to me it has been the difference between, "Here is our computer, on which we do such-and-such," and "Here we do such-and-such, and somewhere inside, a computer helps us do it." Usually, anything an embedded processor can do, a properly equipped IBM PC clone can do too; though usually at greater cost, more power, a larger footprint, and with a keyboard, video card, and monitor trailing behind.

For many projects, a single board with a CPU, some RAM, some ROM, and some I/O will do nicely. There are many such products on the market. Some are coded in assembler, some in FORTH, some even in BASIC. Cross assemblers and cross compilers are available. Some have floating point libraries, little file systems, fancy real-time operating systems, little UNIX, bigger UNIX; some don't. Embedded processors many times require you to learn a new programming language, a new operating system, and a new hardware architecture. To be sure, if you want to manufacture two million VCRs, using a CPU that is ten dollars cheaper will save you twenty million, so new learning in this environment is cost effective.

On the other hand, say you have 17 hours to do a one-time project. You have no time for new learning, you'd like to use development tools and a hardware architecture that you are very familiar with. For many hams, that is the IBM PC programming model. The basics have remained unchanged since 1984. We know how to program the serial and parallel ports. We know how the timer works. We have a test and development system, compilers with floating point libraries, and operating systems that write to disk. I recently bought two CD-ROMs for $50 with 1.2 Gigabytes (compressed!) of sample programs and source code.

While it would certainly have been feasible to hook a speed sensor to a notebook PC and throw it in the soapbox derby car, trusting $4000 of consumer electronics to an 11-year-old careening down the street in $50 worth of pine is not my first choice. It would be nice to have an IBM PC on a single board that requires only a simple serial connection, no keyboard and no screen. It would also be nice if it ran off a single voltage.

Fortunately, there are such things. The one I've used for several recent projects is the KS-6 from KILA Systems.[1] (I have no relation to them other than as a user.) A board configured with a 16-MHz V40 CPU, two serial ports and one parallel port, 512 kbytes of memory and a 128-kbyte SRAM pseudo-floppy drive costs a bit less than $500.00. This is a CMOS, but not strictly low-power, device, consuming

5949 Pudding Stone Lane
Bethel Park, PA 15102
email: nk6k@amsat.org (Internet)
71635,1174 (CompuServe)

[1]KILA Systems, 2300 Central Ave, Suite C, Boulder CO 80301, tel 303 444-7737.

Fig 2

around 700 mA at 5 V at full speed. The clock speed can be divided by 2, 4, and 8 to provide low-power modes. The V40 is the NEC version of an Intel 80186, code compatible with the original IBM PC 8088. The KS-6 is shown in Fig 1.

A feature of the KILA system is that it can provide up to five serial ports on the card; four standard PC ports plus the V40's on-chip port configured as the console. It also has standard IBM PC (ISA) 16-bit bus connectors. This allows it to be used in a passive backplane with other PC adapter cards.

The card is useful all by itself, however. It requires a single five-volt supply and provides its own RS-232 levels. It auto-boots to an A> prompt. It includes DOS in ROM and an XMODEM program that can upload files to the "hard" drive, implemented in battery-backed SRAM. This will allow you to place a program in the C: drive, and to read/write and save data on the C: drive. The on-board SRAM can be expanded to 512 kbytes. Two byte-wide sockets can provide a ROM-based D: drive for a 1-Mbyte pseudo floppy drive.

As shown in Fig 2, the card was mounted in a box and powered with a six-volt lantern battery. The computer and battery were stowed behind the driver. I used the standard PC parallel (LPT1) port to provide a couple of status LEDs and the speed sensing device. Fig 3 shows how to get up to eight status lights and a make/break input from a PC printer port.

To light an LED with the circuit in Fig 3, set the matching bit to zero and write to I/O address 0x378. A one bit turns the LED off. The switch can be read in bit six of I/O address 0x379.

Life Lessons

In this little throw-away project, I was able to make use of many hard-won lessons I have learned over the years. Like the poster titled "Everything I Needed to Know I Learned in Kindergarten," I could write a book called "Everything I Needed to Know I Learned on Field Day."

The first lesson, especially important with embedded processors, is: flashing lights are great. You need to be able to tell at a glance if the otherwise-silent black box is working. I always use one LED to let me know when power is applied. Another is set to blink once a second to let me know the program is running. Another blinks once each time the program does whatever it is supposed to do, in this case, sense wheel revolutions. These lights always save a lot of time in test and setup, and can diagnose some problems just as well as a $20,000 logic analyzer.

Second lesson: have a backup plan. My method for measuring the speed of the derby car was to count the number of wheel revolutions per unit time. Knowing the distance travelled per revolution gives distance over time. Speed, total distance, acceleration, a veritable orgy of physics and input for *MathCad* and *Excel* can be produced from this simple measurement. The first method that came to mind was to attach an IR transmitter and receiver to the body of the car and a small mirror to the wheel. The mirror would reflect the IR to the receiver which would ground the -ACK line on the parallel port, allowing me to count. That was an easy program to write, and I had it running on the KS-6 by 2:00 a.m. I went to the local Radio Shack later that morning to check out the selection of IR components.

Radio Shack had a receive module that would detect IR and only trigger if the IR was on/off modulated at 40 kHz. This would protect my simple project from false triggering on background sources. Radio Shack, of course, did not have a matching transmit module that generated a 40-kHz signal. One IR LED, receive module, 555 timer chip, handful of RC, and two hours later I had a transmitter that, according to a scope, was driving the LED at 40 kHz. There was no sign that the receive module was triggering, though.

It was noon, six hours to show time. My backup plan was to use a simple IR detector and software to counter any falsing that might occur. Back to Radio Shack for a IR emitter/detector pair. Back home. No dice. The current consumption was what I thought it should be for the transmit LED, but the receiver wasn't seeing it. More time goes by. Two hours to show time.

Lesson three: don't overlook the low-tech solution. As my final backup plan, I had purchased a lever-driven micro switch; if all else failed, I could rig the switch to trigger once per revolution. My first choice wasn't a mechanical switch, of course. The switch wouldn't

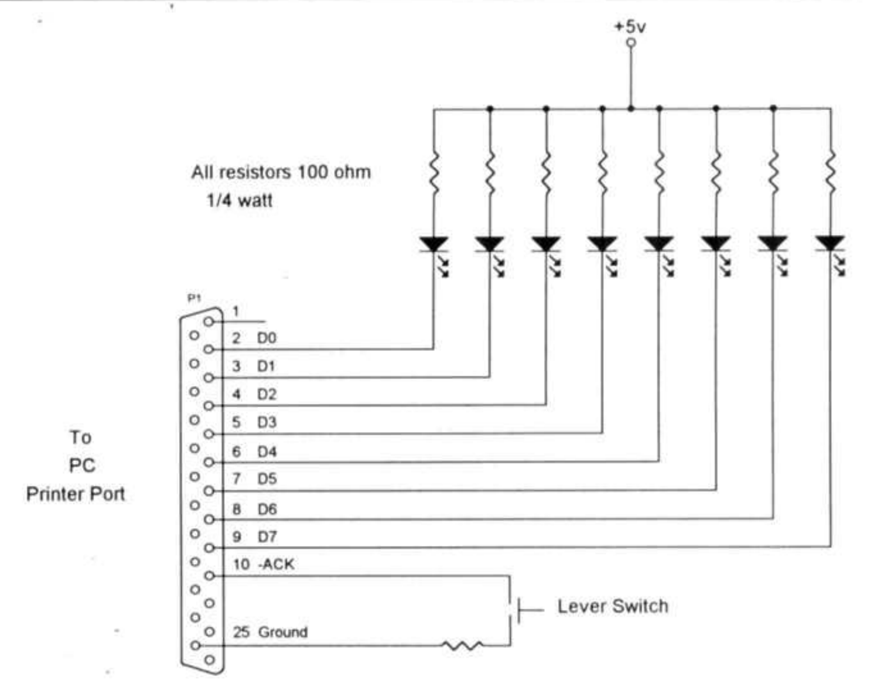

Fig 3

last long, as I expected 18 revolutions per second at top speed. I also didn't know what the switch's recovery time would be. Finally, a mechanical switch isn't nearly as impressive as the IR sensor. The mechanical switch had the advantage of being more nearly foolproof, however, and Fig 4 shows the speed sensor in its final configuration. The switch was taped to the axle support, and an expired credit card was taped to the wheel to push the lever once per revolution.

One of the status LEDs blinked each time the switch closed, allowing us to replace the switch and credit card as they wore out. The KS-6 counted switch closings (through a debounce routine), and wrote the number of closings per second into a file. At the end of each test run, a serial cable was run between the car and a laptop PC. XMODEM was used to download the file for analysis. Next year, we'll add an HT and a TNC for easier access to the data.

This project achieved my goal of exposing the neighborhood kids to a computer that did something other than aid the rescue of the Mushroom Princess. The computer showed that the Taylor's car setup was already optimized, and the car went on to beat all others of its body type in the Fourth of July race. It was beaten by several cars of a newer streamlined design, though. Next year, I'll tell you about the KS-6 based wind tunnel...

The point, though, is that there are rugged, low-power, single-card IBM PCs on the market. You don't need any fancy support gear to program them. I tested the program on my PC, then uploaded the .EXE file to the SRAM drive using XMODEM. These boards provide an excellent tool for quick development, low-volume instrumentation or process-control applications. I used the KILA KS-6 because it provides five serial ports and can be used in a passive backplane. Other smaller, lower-power cards are available. Several are based on the PC chip from Chips and Technologies. This single chip provides most of an IBM-PC, including 80186-compatible CPU, DMA, timers, serial port, speaker port, keyboard interface, and CGA video interface. One such system, with RAM and ROM, is 3.5" × 2.5", and uses 100 mA at 5 V when running at 7.2 MHz, 46 mA when in low-power sleep mode (clocks running), and 1 mA when suspended. This card, from DSP Systems in England, is $800.

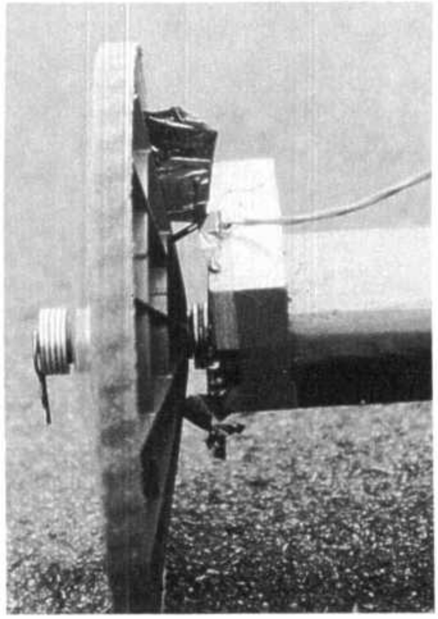

Fig 4

Mailbag

One of the problems for mostly digital guys like me is that modular computer building blocks, as in the project described above, are easier to come by than modular radio building blocks. Over the past few columns, we've been iterating toward the desire for a virtual radio, as was pointed out in email from Howie Cahn, WB2CPU: "I was pleased to see two important topics discussed in your June *QEX* column: computer control of radios and standards for data transfer among *Windows* applications. I agree with most of what you, VE3PAZ/WA3ZIA, and WB4SBE/8 had to say. I think one of the reasons for the reluctance of transceiver manufacturers to provide better support for external computer control is that it might make clearer that computer control, using a 'virtual radio' (derived from the term 'virtual instrument') graphical user-interface is better than having all those expensive and inflexible knobs, readouts, and switches on the radio front panel. I would suggest taking even one more step—giving the panelless radio box an open, modular hardware architecture so that the hardware and the hardware/software combination would benefit from the same advantages that having an open software framework provides. I wrote an article in the Winter '93 issue of *Communications Quarterly*, 'Bringing Amateur Radio into the Computer Age,' presenting some ideas on the subject."

Howie also sent me information on an Open Radio Architecture working group organized by Eric Scace, K3NA. Contact Howie at howi@world.std.com, Eric at ERIC.L.SCACE@sprintintl.sprint.com. □□

Upcoming Technical Conferences

Eastern States VHF/UHF/SHF Conference

• August 13-15, 1993, The Quality Inn, Vernon, Connecticut

• Contact: Byron Blanchard, N1EKV, 16 Round Hill Road, Lexington, MA 02173 or tel 617-862-1380. (Registration will be $25 at door.)

• Events: Friday, informal gathering; Saturday, technical talks, "rap sessions" on VHF/UHF bands, and noise-figure clinic and contest; Saturday evening, banquet, "VHF Trivia" quiz, and informal social activities; Sunday morning, antenna gain measurement (220-2304 MHz) and flea market, weather permitting.

• Hotel reservations: call Lori Tozier (at the Quality Inn) at 203-646-5700/ 800-228-5151 for the special rate of $49.95 single, $55 double.

1993 ARRL Conference on Digital Communications

• September 10-11, 1993, Holiday Inn Airport, Tampa, Florida

• Contact: Brian Lantz, KO4KS, TPALAN, 6403 N Paddock Avenue, Tampa, FL 33614, Tel 813-877-1469, Internet brianlantz@delphi.com.

• Events: Friday, registration and round-table discussions; Saturday, registration, technical talks, discussions on "standards."

• Misc: several "distractions" are available on site and within a short drive.

• Hotel reservations: call the hotel direct at 813-879-4800 and mention the conference. A special rate of $50 per night is available from 3 days before through 3 days after.

• **Call for papers:** Camera-ready papers should be sent to Maty Weinberg at ARRL HQ by **August 6, 1993**. Electronic submissions will be accepted, please call Maty at 203-666-1541 or Internet at lweinber@arrl.org.

Microwave Update '93

• September 23-26, 1993, Northwest Atlanta Hilton, Atlanta, Georgia

• Contact: Jim Davey, WA8NLC, 4664 Jefferson Township Place, Marietta, GA 30066, tel [W] 404 333-2136 [H] 404-998-6971.

• Events: Thursday evening, registration and socializing; Friday, technical sessions, flea market and noise figure contest; Saturday, technical sessions and banquet; Sunday, a short meeting is planned.

• Hotel reservations: call Kim Gilliam (at the Hilton) at 800-234-9304 by September 1 for the special rate of $62 per night. An airport shuttle bus is available at a round-trip cost of $25. Call 800-237-0709 for reservations.

(*Have an upcoming technical event? Drop us a note with the all the details and we'll include it in* Upcoming Technical Conferences.) □□