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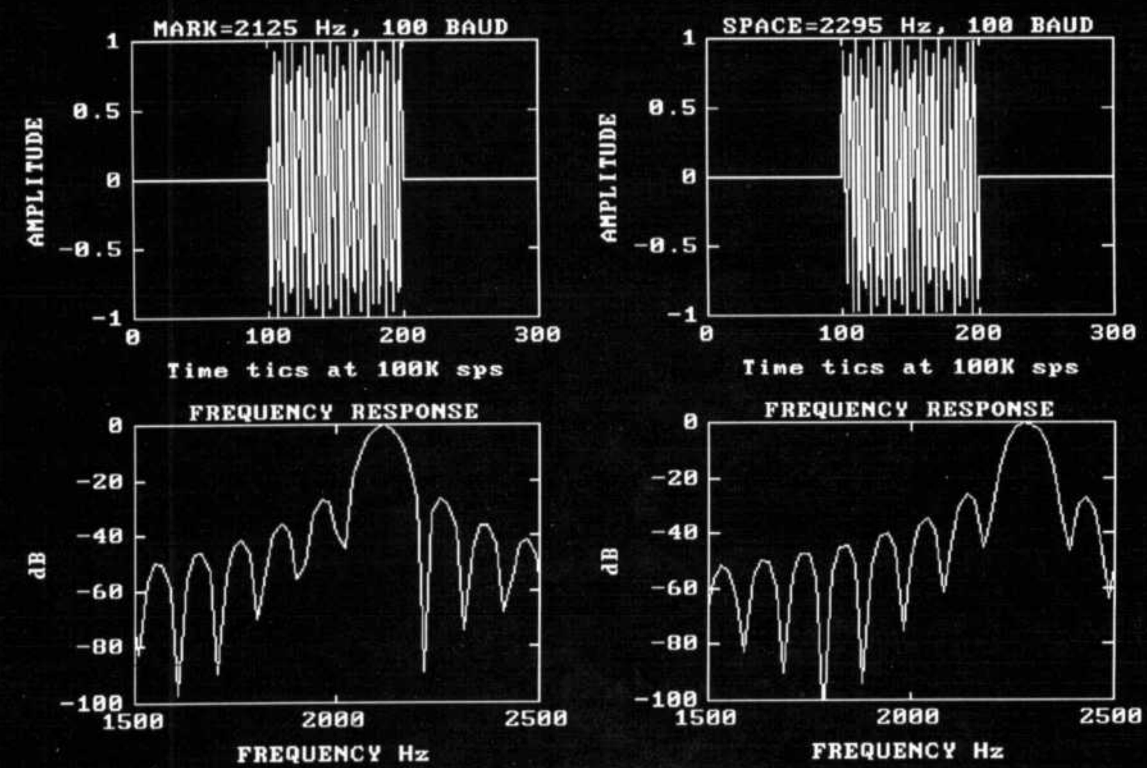
November 1994



High-Quality HF DSP Modem Design

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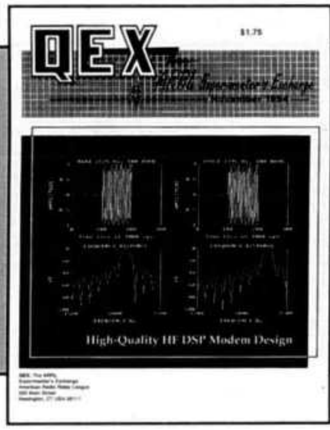

About the Cover

Designing a modem for HF is a challenge. DSP makes it easier, but you still have to do it right—and KC7WW has.

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Features

Columns

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THE AMERICAN RADIO RELAY LEAGUE

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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

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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*.

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Empirically Speaking

A Virtual Ionosphere

One of the challenges amateurs are grappling with these days is to create and promote efficient means of sending digital data at HF. Some useful work has taken place in this arena in the past few years, as demonstrated by the increasing presence of Pactor, Clover and G-TOR signals on the HF bands. Most of the effort to date has been in the area of protocol development. This is important, as protocols have a key role to play in overcoming the effects of the vagaries of HF propagation on digital data. But the other key player in the system is the modem, and, with the exception of Clover, little has been done to improve amateur HF modem technology.

Why is this? We feel that a large part of the reason is the incredible variation of the ionosphere and the consequent difficulty of taking an experimental approach to HF modem design. Sure, we can do on-the-air testing—that's what we have licenses for. But the ionosphere is a continually changing medium, and on-the-air testing can check modem performance over only a small range of possible channel characteristics at any time. To test against a wide range of ionospheric conditions requires a *lot* of on-the-air testing. Worse, it's difficult to say with any certainty just what the exact condition of the ionosphere is at any point in the test. The result is that modem design has relied on general principles, rather than on matching explicit ionospheric parameters.

It is, of course, quite possible to use these general principles to improve modem design—witness KC7WW's article in this issue. But a lot of people are thinking about more complex modulation schemes than the simple binary FSK that is commonly used for amateur digital communications. To implement more complex schemes calls for a more detailed consideration of the HF channel. And it calls for a better way of testing our modem solutions than random on-the-air use. It calls, in short, for an HF channel simulator.

Simulating the HF channel at the operating frequency would be difficult. Fortunately, we don't need to do that. We can perform our simulation at baseband. Since the baseband bandwidths of our HF signals are limited, the simulation problem lies well within the capabilities of modern general-purpose DSP systems. We can treat the propagation path as a time-varying linear system and simulate it. "All" we need to do is to define the channel characteristics and write the simulation code.

As this is being written, Tucson Amateur Packet Radio, Inc, (TAPR) is sponsoring an Internet mailing list to support discussion of simulator design and implementation—and related HF digital subjects. If the fruit of these discussions is effective HF channel simulation, we are poised to make a substantial improvement in our HF digital data systems. To participate in the discussion—even better, to participate in the work—join the hfsig@tapr.org mailing list by sending mail to listserv@tapr.org containing the single line:

subscribe hfsig <your name>.

This Month in *QEX*

DSP is the obvious choice for a modern HF modem design. This month Johan B. Forrer, KC7WW, presents "An Adaptive HF DSP Modem for 100 and 200 Baud" that provides state-of-the-art performance by using state-of-the-art technology.

"A Better A/D and Software for the DDC-Based Receiver," by Peter Traneus Anderson, KC1HR, updates the author's March, 1994, *QEX* article, "A Simple SSB Receiver Using a Digital Down Converter," with a 10-bit A/D, providing more than 60 dB of dynamic range.

Part 3, the final part of "Practical Microwave Antennas," by Paul Wade N1BWT, discusses lens antennas you can build and shows how to make effective microwave antenna measurements.

In this month's "RF" column, Zack Lau, KH6CP/1, shows how to improve the performance of 10-GHz wide-band FM systems that use Gunnplexers. *—KE3Z, email: jbloom@arrl.org (Internet)*

An Adaptive HF DSP Modem for 100 and 200 Baud

Applying DSP techniques results in low cost and superior performance in the harsh HF environment.

by Johan B. Forrer, KC7WW

Digital signal processing (DSP) hardware and software have become tools of the amateur digital experimenter. This article offers a novel and cost-effective implementation of a DSP modem for HF AMTOR and Pactor using a DSP-based personal computer (PC) sound card.

The object of HF modem design is to strike a balance between optimal performance, robustness and operational flexibility. The design described here has excellent performance that will meet the requirements of the discriminating user.

This project is based on a low-cost DSP sound card that uses the Analog Devices Personal Sound Architecture, also known as the "PSA chip set."[1] An introduction to programming this type of DSP sound card was presented in a previous *QEX* article.[2] The reader is encouraged to refer to that article.

The DSP modem component of this application forms only a small part of a moderately complex system consisting of various interrelated hardware and software components.

[1]Notes appear on page 10.

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Monroe, OR 97456

Demodulator Requirements

Probably the most critical part of the system design, as far as performance goes, is the demodulator. (See sidebar: "Modulation, Demodulation and Detection.") HF operation presents unique challenges to the modem designer (see sidebar: "The Challenge of HF"). To produce effective HF performance, the demodulator needs to have several key characteristics:

- Good dynamic range, the ability to deal with both weak and strong signals, is needed in any realistic HF environment.
- Filters of optimal bandwidth and reasonable group-delay characteristics are needed to allow passage of only the desired signals, without distortion. For an adaptive modem such as this one, switching between filters should be clean, without amplitude or phase anomalies.
- The demodulation algorithm should use robust noncoherent demodulation techniques to counter the effects of certain types of noise, phase error and intersymbol interference (ISI). A means for algorithm feedback control by the host-based protocol analyzer is also needed.
- The modem's operation should tolerate some degree of normal operator misadjustment and/or equipment operating tolerances.
- Modem status and tuning aids must be available to the operator. These include a signal-strength indication, overload indication and a means for the operator to judge the tuning error.

Most of these requirements translate into software specifications for implementation on either the DSP sound card or the PC platform.

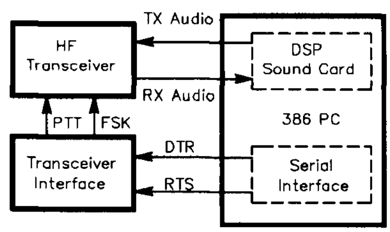


Fig 1—Components of the system. The DSP sound card processes received and transmitted audio signals, while the PC's serial port provides control lines for the transceiver.

Modulation, Demodulation and Detection

One way of learning about DSP is to experiment, and there are several affordable computer-based DSP learning tools available for this purpose. The design and testing of the filters used in this article are the result of such an experimental approach.

A better understanding of the characteristics of FSK modulation allows for optimal demodulation. It is fortunate that many applied DSP principles can be illustrated intuitively. One such concept is the duality of the time and frequency domains. Consider for example an FSK signal at an instant *t* in time—two possible signaling states exist:

$$S_1(t) = A\cos(\omega_1 t + \phi_1)$$

$$S_2(t) = A\cos(\omega_2 t + \phi_2)$$

$$A = \sqrt{\frac{2E}{T}}$$

where: A represents the energy per signaling element, angular rates ω_1 and ω_2 correspond to the mark and space tone frequencies, and ϕ_1 and ϕ_2 are arbitrary phases associated with S_1 and S_2 respectively. Fig A shows such waveform pulses. (*Matlab* and the Ryerson Polytech *Communications Toolbox* simulated the environment and produced the graphs in this article.)[1,2] Note that FSK tones are usually generated using a frequency-modulated oscillator, thus producing a continuous-phase signal. The examples shown here are for illustrative purposes only. These graphs attempt to illustrate several important properties of an FSK signal:

- A square pulse produces significant energy in its sidelobes. One object of modem filter design is to find a good compromise that will truncate such sidelobes while maintaining a good S/N ratio.
- The width of the pulse, that is, the duration of a signaling element, determines the bandwidth of the major lobe and also the pattern of the nulls in the frequency spectrum. Narrow pulses require additional bandwidth. For 100 baud, the null-to-null bandwidth of the major lobe is 200 Hz and the nulls in the spectrum are 100-Hz apart.
- The bandwidth of such a signal pulse at the –60-dB level may well be on the order of 800 Hz. This means that a high-power signal could occupy a significant amount of additional bandwidth, beyond that needed for communication. For narrow bandwidth requirements, special pulse shaping should be considered.

When the transmitted phase is of no concern to the detector we are using *noncoherent* detection. The minimum tone spacing for a noncoherent 100-baud FSK signal is 100 Hz, theoretically. Fig B shows that a 170-Hz shift FSK signal has adequate separation between its tones. Although not shown in this figure, the minimum spacing for a coherent FSK signal is only 50 Hz—half the bandwidth of noncoherent FSK. This modulation scheme is called minimum shift keying (MSK).

Fig B shows the overlapping frequency responses for a 100-baud, 170-Hz-shift noncoherent FSK signal. The bandwidth of this FSK signal is given by:

$BW = S + F \times B$

where S is the frequency shift, F is the shape factor, and B is the signaling speed, in bauds. A shape factor of 1.6 is commonly used. The theoretical bandwidth for the FSK signal is then 330 Hz (170 + 1.6 × 100), which roughly corresponds to the –16-dB level of Fig B.

Demodulation refers to the inverse process of modulation: recovery of the modulation signal. Detection, on the other hand, refers to the process of extracting symbols from the demodulated signal, as shown in Fig C. This article applies "passband" demodulation techniques because band-pass filters, tuned to the frequencies of interest, perform the demodulation. There also is a popular alternative method for demodulation that uses heterodyning and a low-pass filter to effectively shift the passband signal to a baseband form. That is called "baseband" demodulation. The performances of these two methods are comparable for this type of application. The passband method is chosen here for reasons of simplicity.

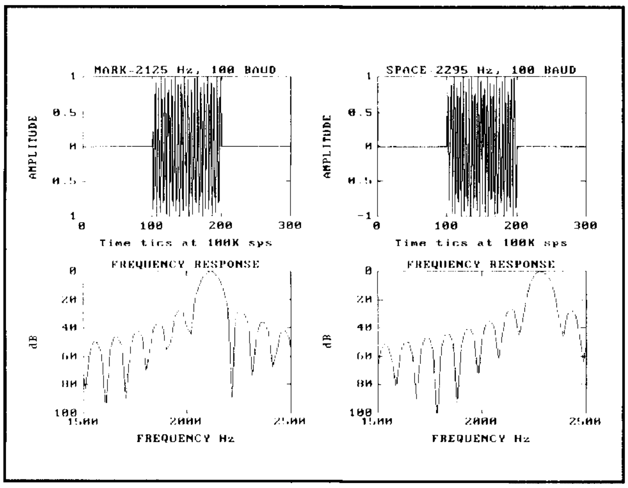


Fig A—Mark (2125 Hz) and space (2295 Hz) pulses comprising a 100-baud, two-state FSK signal are shown in the upper left and right parts of the figure, respectively. These tone pulses are each 10-ms long and have an effective repetition rate of 50 pulses per second (assuming a symmetric ...0101... modulation pattern). The lower part of the figure shows the respective frequency responses for these tone pulses. Note that the width of the main lobe, from null to null, is 2 × 100 = 200 Hz wide and that the nulls in the sidelobes are at 100-Hz intervals.

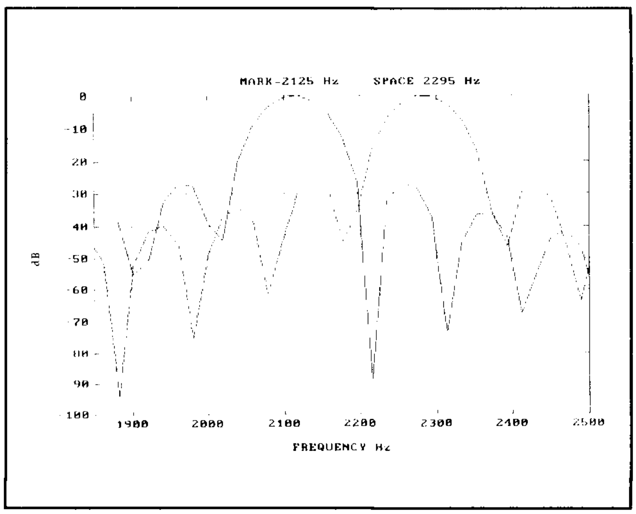


Fig B— The overlapping frequency responses of the mark and space tones of a 100-baud, 170-Hz-shift FSK signal with 2125-Hz mark and 2295-Hz space tones.

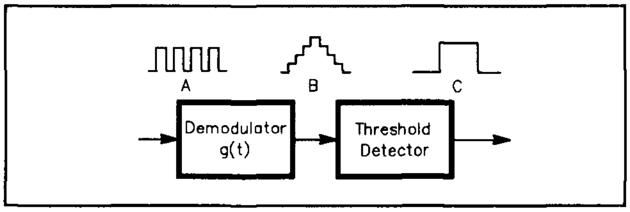


Fig C—Demodulation and detection of a tone pulse. The input signal (with additive noise), A, is applied to a demodulator to produce the signal at B. This signal, processed by the detector, produces the symbol signal, C.

The demodulator is a special function, $g(t)$, with the property that, as it compares its own properties to that of an unknown signal that contains the wanted signal, it will show a high degree of correlation. For example, let $g(t)$ look like a mark pulse, as shown in the lower left graph of Fig A. When the unknown signal that contains a mark tone is correlated with $g(t)$, by sliding the unknown over $g(t)$ one sample at a time while computing the correlation, there will be near-perfect correlation at the center of the symbol. This is shown at B in Fig C. It follows that a noise pulse—or any pulse containing some other tone—would *not* show this high degree of correlation, regardless of how it is shifted in time. This principle is in essence what the theory of matched filters is all about: it describes a method by which an unknown signal, embedded in noise, can be detected in such a way as to optimize the signal-to-noise ratio (S/N). Unfortunately, the theory of matched filters also requires a good model of the noise, something that is very difficult to achieve for HF. But it is still a valid and useful technique.[3]

It should be evident that the required matched filters for 100-baud demodulation are those shown in the lower part of Fig A. The need for a different set of filters for 200-baud demodulation is also clear—thus the adaptive requirement for the modem.

The last piece of theory before we can show the DSP modem's architecture is the discriminator, shown in Fig C as a simple threshold detector. The choice of a threshold requires knowledge of the character of the wanted signal, as well as that of the noise. The theory of stochastic statistics, which deals with making decisions under uncertainty, may be applied in this situation. Maximum likelihood theory (MLT) in particular, is frequently used in this context.

To detect symbols in a two-state system, a minimal requirement is that when symbol $d = 0$ is sent:

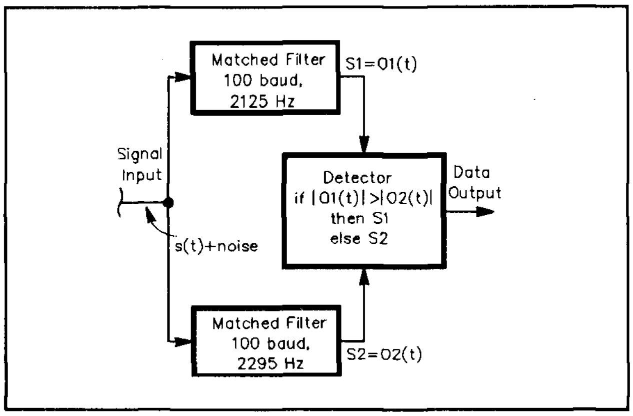


Fig D—A two-state discriminator based on maximum likelihood theory. A noisy signal is applied to two matched filters, one each for the binary states S_1 (mark) and S_2 (space). At time instant *t*, the matched filters produce outputs $O_1(t)$ and $O_2(t)$ respectively. The discriminator compares the magnitudes of O_1 and O_2—the output of the discriminator is set to the state corresponding to the largest matched filter output.

$$P\left[S_1 \middle| s(t)\right] > P\left[S_2 \middle| s(t)\right]$$

and similarly, when symbol $d = 1$ is sent:

$$P\left[S_1 \middle| s(t)\right] < P\left[S_2 \middle| s(t)\right]$$

This set of inequalities states a simple dichotomy: when symbol $d = 0$ is sent, the probability of recognizing state S_1, given signal $s(t)$, should be larger than the probability of recognizing state S_2. The opposite should hold true when symbol $d = 1$ is sent.

There are many functions that will satisfy this dichotomy, but there is one in particular that will maximize S/N as well. This function is a pair of matched filters.[3] Fig D shows a means of implementing this logic.

[1] *Matlab* is by The Math Works, Inc, Cochituate Place, 25 Prime Park Way, Natick, MA 01760.

[2] Zeytinoglu, M. and Ma, N. W., *Communications Toolbox for Matlab*, Department of Electrical and Computer Engineering, Ryerson Polytechnic University, Toronto, Ont, Canada. Obtainable via Internet anonymous ftp from: ftp.mathworks.com /pub/contrib/misc/comm_tbx.tar.

[3] Blahut, Richard E., *Digital Transmission of Information*, Addison Wesley, 1990, Chapter 2-3.

System Overview

This project integrates the signal processing capabilities of the sound card's DSP processor with the general-purpose PC platform. The DSP on the sound card functions as a coprocessor that performs real-time audio-frequency processing and operates independently—and concurrently—with the PC's processor. The PC's processor handles protocol and housekeeping functions. Fig 1 shows this relationship.

Four major components make up the body of the system software: the DSP bootstrap and real-time executive, DSP application code, low-level PC interface and host PC application. Fig 2 shows the functional relationship between these components.

DSP Software

The DSP software consists of two parts: the real-time executive (RTE), and the actual DSP modem code.

The DSP Real-Time Executive (RTE)

The DSP RTE software module provides:

- startup initialization;
- an asynchronous command interpreter for communication between the DSP and the host;
- interrupt service routines (ISRs) for DSP hardware such as the timer and the codec (the analog-to-digital and digital-to-analog converter); and
- periodic interrupts of the host for

The Challenge of HF

The influence of the ionosphere on HF propagation has been studied extensively.[1,2,3] The dynamic nature of the ionosphere affects HF digital communications in several ways.

In the most general case, multipath propagation—simultaneous reception of high-angle and low-angle rays—generally depends on both the operating frequency and the distance between stations.[1] Such multipath propagation is most noticeable at frequencies much lower than the maximum usable frequency (MUF). For example, it's quite common on the 40-m and 80-m amateur bands. The longest of such delays is typically at a frequency of 0.43 times the MUF and may introduce path differences on the order of 3 ms for a 2500-km path and 8 ms for a 1000-km path. Consider the case where both high-angle and low-angle signals are received with a 3.05-ms difference in path delay. Such a phase delay is equal to approximately 13 half-cycles of a 2125-Hz mark signal. This means that the two signals are effectively out of phase, which results in almost complete cancellation of the mark signal at the receiver (assuming equal amplitudes of the signals). But for a 2295-Hz signal, the 3.05-ms difference is about 14 half-cycles, so the signals combine in phase. Thus the mark signal is diminished, while the space signal is not. This effect is known as selective fading. Early RTTY used a wide 850-Hz shift, but this was changed to a narrower 170-Hz shift, in part due to problems with selective fading. Although it's not used much in Amateur Radio, diversity reception is one way to minimize the effects of fading. The higher frequency bands, 15 m, 10 m and 6 m, exhibit slow, deep, flat fading characteristics.

Rapid fading of either of the paths introduces some degree of phase jitter in the received signal that may result in intersymbol interference (ISI)—an overlap (in time) between adjacent signaling elements. In some cases, ISI can introduce problems in detecting when one signaling element ends and the next one starts. One advantage of noncoherent detection is that it is quite forgiving and will tolerate a fair amount of phase jitter, at a small cost in performance.

An additional cause of variation in HF propagation is electromagnetic polarity rotation as a radio signal passes through the Earth's magnetic field—also known as Faraday rotation. For example, a vertically-polarized signal may be received at the distant location with both horizontal and vertical components. Fortunately, this does not introduce a big problem, as most HF amateur stations use relatively low-gain antennas (compared to those being used at VHF and UHF, for instance). However, amateurs have used several innovative antenna designs to deal with such polarization diversity.

Noise at HF introduces perhaps the most challenging problems. Noise can come from several sources, some originating from distant locations, such as electrical storms. Other types of electrical noise may be caused by polarized rain drops, or broad-band noise as received from extraterrestrial sources such as the sun. These, along with man-made noise, make the noise content of a received HF signal quite different from the type of noise described in text books, that is, additive white Gaussian noise (AWGN), which has well-known statistical properties. These factors require special consideration in modem design for symbol detection in the presence of HF noise.

The performance of a demodulator depends greatly on use of the correct filter characteristic, one that optimizes signal-to-noise (S/N) ratio in particular. Both the bit error rates (BER) and the amount of noise energy contained in the recovered signal are greatly dependent on the filter bandwidth.[4] Filter bandwidth will also become increasingly important in the future as the competition for RF spectrum increases.

Transmission protocols also have to cope with data corruption due to noise, fading or adjacent-channel interference. This requires the use of data-integrity checksums, interleaving or burst-error detection and correction methods, tailored to the nature of the characteristic disturbance. These methods form the basis of block and convolutional coding theory.

[1]Freeman, Roger L., *Telecommunication Transmission Handbook*, 3rd Ed, Wiley, 1991, pp 549-577.

[2]Kennith Davies, *Ionospheric Propagation*, Peter Peregrinus, IEE, 1990, pp 191-193 (with acknowledgements to Van Elston, WA9FIO).

[3]McLarnon, Barry D., VE3JF. "New Directions in Amateur HF Digital Data Transmission Systems—Part 1," *QEX*, Dec 1987, pp 3-9.

[4]Sklar, Bernard, *Digital Communications: Fundamentals and Applications*, Prentice-Hall, 1988, Chapter 3.

transferring just-recovered data and for purposes of symbol synchronization.

Fig 3 shows the relationships between these DSP software components.

The RTE code first has to be loaded into the DSP's memory. We call this loading process *bootstrapping*, and it uses the DSP's factory-programmed on-chip ROM program. Executing this ROM code, typically after a hardware reset, the DSP starts from a known state and is ready to load code into the DSP's on-chip memory. Control passes to the instruction at the startup vector when loading is complete.

The RTE contains only a few essentials: the interrupt-vector table, command interpreter and other miscellaneous initialization code. "Hooks" to code outside the scope of the bootstrap allow separate loading of modules that handle the various DSP interrupts.

Loading of the DSP application code is performed via the RTE. Once the DSP is up and running with the bootstrapped RTE code, initially with only the asynchronous command interpreter executing, it awaits commands from the host. Subsequently, more extensive application code, such as the modem application, is loaded into DSP program and data memory by the host, which issues function calls 0x00D0 through 0x00D3 to the RTE (see Table 1).

Once the DSP application code is loaded, the host is ready to receive interrupts from the DSP. The host initiates processing by issuing function calls 0x00B0 (start timer) and 0x00B4 (start 1848 DMA, that is, start the A/D converter). The DSP modem is then fully functional, processing incoming audio samples and interrupting the host every millisecond with a sample of the demodulated data.

Additional RTE command-interpreter functions listed in Table 1 pro-

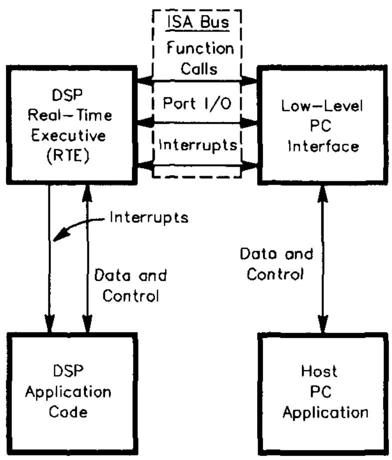


Fig 2—The software components of the system comprise both DSP and PC programs, which communicate via the ISA-bus interface.

Table 1—RTE Function Calls

Code	*Function*	*Parameters*	*Return Value*
0x00B0	Start timer	S	-
0x00B1	Terminate timer	S	-
0x00B4	Start 1848 DMA	S	-
0x00B5	End 1848 DMA	S	-
0x00BA	Update timing value	M	-
0x00BB	Update speaker level	M	-
0x00BC	Update line-in level	M	-
0xYYBD	Update databit state	S*	-
0xYYBE	Update modem select	S*	-
0x00D0	Read data memory	M	YES
0x00D1	Write data memory	M	-
0x00D2	Read pgm memory	M	YES
0x00D3	Write pgm memory	M	-
0x00D4	Send back ID string	S	YES

*Parameter sent in upper byte of code word.
Functions with an M in the parameter column require multiple calls to transfer the parameter. For functions with a return value, the value is returned as part of the software handshake mechanism.

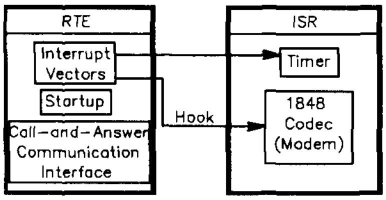


Fig 3—The DSP software comprises the real-time executive (RTE) and the application program. The RTE vectors interrupts to the application program codec ISR, which is loaded separately.

vide access to the timer's interrupt rate control (0x00BA), control over speaker and line-in levels (0x00BB and 0x00BC), 100/200-baud modem select (0x00BE) and modulator tone control (0x00BD).

DSP Modem Code

A fair amount of design effort is required to determine the needed architecture and filters for the demodulator. (See sidebar: "DSP Modem Design"). This demodulator contains an input band-pass filter, discriminator filters and data low-pass filter for each speed. These components are grouped together for either 100- or 200-baud operation, with the selection of speed determined by the state of the "modem_select" (MS) word. The host protocol analyzer decides which mode is appropriate and controls the status of this word with a 0x00BE function call. Note that MS can be changed on the fly, but its effect will only be take place on the next 1848 codec interrupt.

The DSP software interrupts the PC host at one-millisecond intervals using the DSP timer. This allows for 10× oversampling of 100-baud signals and 5× oversampling at 200 baud. Protocols such as Pactor require bit clocks with accuracy better than 30 parts per million (ppm). Experience has shown that differences exist among the crystal clocks of different sound cards—even those from the same manufacturer—demanding the use of different timer constants for different boards. Sometimes it even becomes necessary to use dithering between two clock values to obtain the required accuracy. Although such drastic measures may appear extreme, keep in mind that the host protocol program can track only small amounts of phase drift. During periods when the received signal is weak or corrupted by interference, it is critical to maintain good clock accuracy, as software tracking of timing is not feasible then. This has become even more critical with newer protocols with long frames.

The complimentary component of the demodulator is the modulator; it generates an audio signal for application to a transceiver's balanced modulator. For FSK, the modulator is much simpler than the demodulator. The modulator uses a table-look-up algorithm. The look-up table consists of sine values over one quadrant: zero to $\pi/2$ radians, spaced at $\pi/64$-radian increments. The value of the "databit" word (0=muted, 1=mark tone, 2=space tone) controls audio tone generation. A "cumulative phase angle" (CPA) word is updated at every A/D tick with the appropriate phase increment. CPA is then used as an index into the sine look-up table. It is of course necessary afterwards to adjust the sign for the appropriate quadrant.

The modem code is a straightforward linear progression initiated by the 1848 codec interrupt. The DSP's alternate register set is used to provide fast context switching during interrupt handling. The DSP's hardware pipeline further allows for multiple instruction execution during execution of FIR filter code.[3]

Host PC Software

The PC software provides user control and low-level hardware interfacing. Fig 4 shows the major PC software components.

Note that the application contains a real-time component, the ISR that handles interrupts from the DSP. This piece of low-level code, once started, has to process the steady stream of interrupts from the DSP—one each millisecond—without ever missing a beat. Besides storing data and communicating with the DSP using function calls, this low-level software also may contain the main protocol handler, the part of the software that decodes bit streams into data frames, extracts and converts codes, does checksum calculations and handles the data stream to and from the user interface. This part of the code easily accounts for 90% of

DSP Modem

Fig A shows how a modem design is derived by applying the principles described in the "Modulation, Demodulation and Detection" sidebar. Each of the functions shown will have to be implemented by a DSP algorithm.

Figs B through D show the input band-pass and discriminator filters. These filters were designed using the Parks-McClellan algorithm.[1]

DSP designers have their favorite filter-design tools. Two popular filter-design tools were used in this design, both of which produce highly efficient DSP filters. The Parks-McClellan program is based on min-max optimization using the Remez-exchange algorithm. There are, however, several other popular methods for generating coefficients for FIR filters. The data low-pass filters for the DSP modem were designed using the window method.[2] Doblinger's *FIRMINI* program offers a choice of window types for different kinds of filter type.[3] A Kaiser-Bessel response was selected for the low-pass filters, with cut-off frequencies set at 100 Hz and 200 Hz. The results are shown in Fig E. These filters are not critical, however, and separate filters are not really necessary.

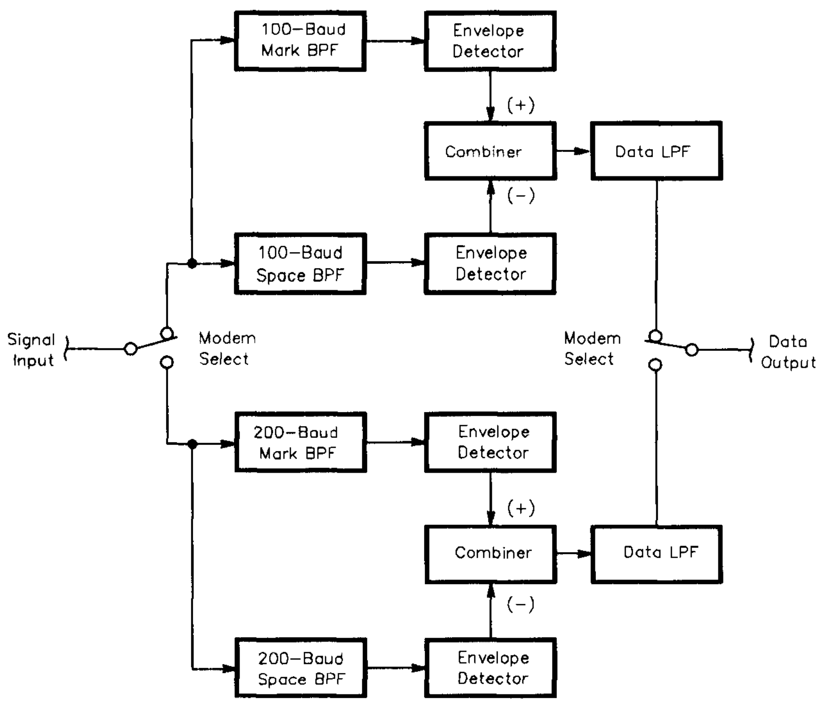
It is evident that the discriminator filters do not exactly duplicate the matched filter specifications—they have much wider bandwidth. The specifications shown have been arrived at by much experimentation and on-the-air testing and strike a balance between optimal performance and ease of use. The 100-baud modem will work at either 170- or 200-Hz shift and will pass adequate quality 200-baud traffic such that the protocol analyzer can recognize this state and be able to switch to the 200-baud modem. The 200-baud modem is optimized for 200-Hz shift, although it uses the same passband center as does the 100-baud modem.

[1]McClellan, J. H., Parks, T. W. and Rabiner, L. R., "A computer program for designing optimum FIR linear phase digital filters," *IEEE Transactions on Audio Electroacoustics*, Vol AU-21(6), 1973, pp 506-526. A program for PCs is obtainable via Internet anonymous FTP from: ftp.ucsd.edu as /hamradio/dsp/fir.arc.

[2]Taylor, F. J., *Digital Filter Design Handbook*, Dekker, 1983.

[3]Doblinger, G., *FIRMINI*, TU-Wien, Institut E389. 1994. Obtainable via anonymous FTP from: ftp.analog.com, 21xx family tools.

Fig A—The DSP modem architecture. The input signal is routed via one of two modems according to the state of the modem-select switch. Each modem consists of a band-pass filter, a pair of matched filters for mark and space, an envelope detector, a signal combiner and a data low-pass filter. The output is a signed 16-bit word that corresponds to the magnitude of the detected state. Input signal sampling is at 5512.5 samples per second (sps) while the output rate is 1000 sps.



the total programming effort. The remainder of the development effort is split equally between the DSP code and the PC user interface.

The software on the PC must:

- set up the DSP hardware, do the DSP bootstrap and load the DSP modem code to the DSP memory;
- prepare the PC for receipt of interrupts from the DSP;
- start the DSP application program;
- initialize the protocol analyzer and user interface;
- service commands from the user while maintaining the real-time protocol analyzer; and
- upon a user-initiated shutdown, terminate the DSP task in an orderly manner, close all files and restore the PC interrupt system.

Putting It Together

Source code for the DSP modem, DSP RTE and a simple user interface that includes the needed PC low-level I/O is provided for demonstration pur-

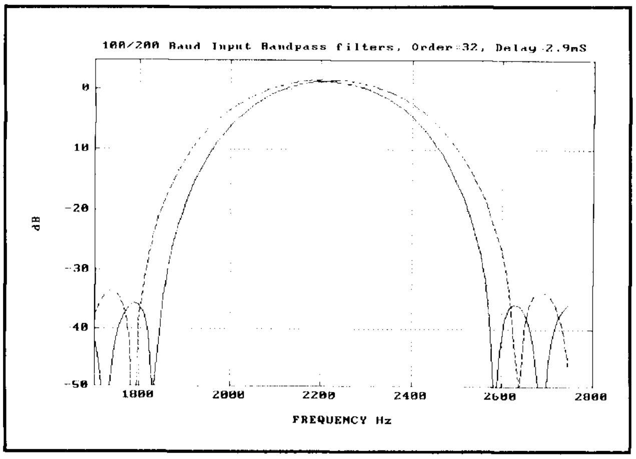


Fig B—Input band-pass filter responses for 100 and 200 baud. The solid line is the 100-baud filter, the dashed line is the 200-baud filter. The mark and space tones are 2125 and 2295 Hz respectively. The sample rate is 5512.5 sps.

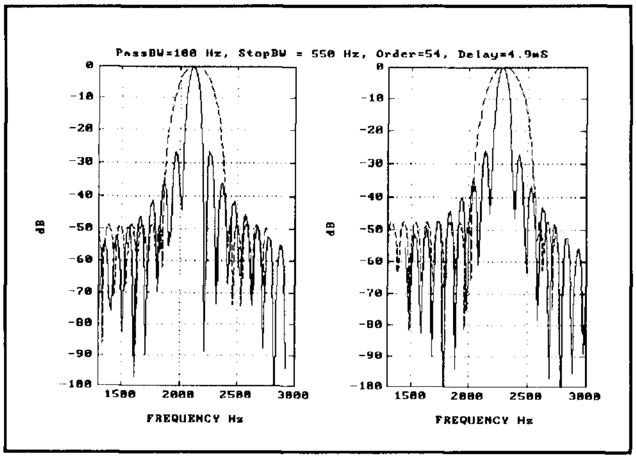


Fig C—The detector band-pass filter responses for 100 baud (dashed line). The mark filter, at the left, is centered at 2125 Hz, while the space filter is at 2295 Hz. The sample rate is 5512.5 sps. The solid line shows the response of a matched filter.

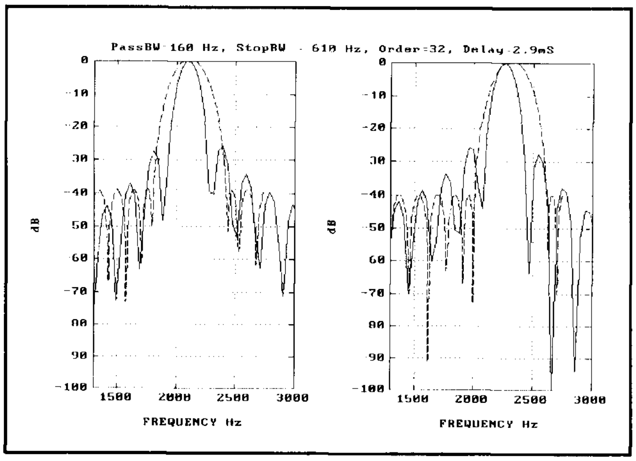


Fig D—The detector band-pass filter responses for 200 baud (dashed line). The mark filter, at the left, is centered at 2125 Hz, while the space filter is at 2295 Hz. The sample rate is 5512.5 sps. The solid line shows the response of a matched filter.

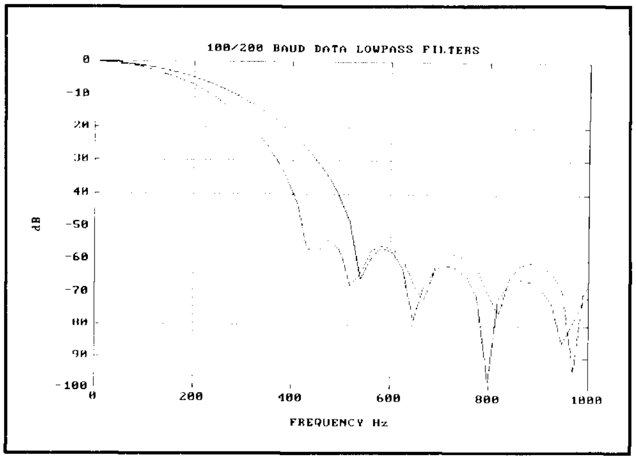


Fig E—The data low-pass filter responses for 100 baud (dashed line) and 200 baud (solid line), as designed by the *FIRMINI* program. The cutoff frequencies are 100 and 200 Hz, respectively, and the sample rate is 5512.5 sps. 60 dB of rejection was chosen for the stopband.

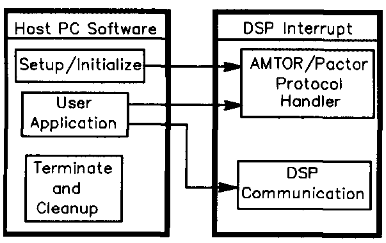


Fig 4—The PC software components handle user interface and interface to the DSP modem.

poses.[4] A full-featured user-interface program for AMTOR and Pactor is also available.

The demonstration software is a fully functional program that loads the DSP modem code and provides user control over the DSP's audio and timing. This code provides a basis for further development. An interesting possibility, for example, is to extend the code to become a pop-up terminate-and-stay-resident (TSR) program activated by a "hot-key." This would allow the user to control the DSP modem at will. The PC's serial interface provides a convenient means of integrating the modem into existing TNCs.

Here's a very brief overview of the necessary steps to build the working application. The supplied software consists of three modules: the DSP RTE, DSP modem and the PC user-interface demo. In addition, the software tools from the introductory article on programming the DSP sound card are required to generate the loadble DSP files and compile and link the C++ modules. (Use the latest support library code, supplied with the code example, during compilation and linking.) An updated version of the SpAsm 21 assembler is also provided.

The RTE.DSP file contains the DSP RTE source code. The code includes interrupt vectors, DSP start-up code and the call-and-answer command interface. Assemble the source code and convert it to a bootable format using the commands:

```
spasm21 rte.dsp
cload rte
```

The SpAsm21 assembler takes RTE.DSP as input and produces RTE.CDE (an intermediate object). CLOAD produces RTE.LD, the loadable DSP object file, from RTE.CDE.

MODEM.DSP contains the DSP modem source code. This code includes the two modems, one each for 100- and 200-baud operation. It also contains code for the modulator and its lookup table. Assemble this module with the command:

```
spasm modem.dsp
```

DEMO.CPP contains C++ source code for the demonstration user-interface program, which includes several functions to do bootstrapping, loading of the modem code to the DSP and other low-level I/O functions. This module must be compiled using the large model. Also, note that the audio-source and input-gain parameters may require adjustment for different PSA cards. The demo code contains example values for Cardinal and Orchid cards. Compile and link the demo program as follows, using Borland C++:

```
bcc -c -ml -O2 -Z -G demo.cpp
tlink /x c:demo PSA1 PSA2,
  emu mathl
```

When the demo program is run, there will be a brief pause as the DSP boots and the user interface initializes itself. If a signal source is connected to the sound card's microphone jack, the sound, processed by the input bandpass filter, will play through the sound card's speaker. The demo program's menu allows for the selection of several commands that you can experiment with.

I've made available two shareware programs, PSATOR and PSA-PACTOR, which offer full transmit and receive capabilities. These provide good examples of the integration of the DSP modem with a protocol analyzer application.[5]

Conclusion

This design offers a flexible and cost-effective approach with extremely good performance. It also provides sufficient background material to encourage both experts and newcomers to experiment and extend the technology.

I'm grateful to all those who have participated in many stimulating discussions on this subject and who have provided assistance during beta testing. This work was possible only with the generous contributions of many gifted individuals.

Notes

[1]Analog Devices, One Technology Way, P O Box 9106, Norwood, MA 02062-9106, USA. For literature, contact Analog Devices Literature Center at 617-461-3392. For further information on the Analog Devices PSA chip set, see the frequently-asked-questions (FAQ) obtainable from Analog Devices Signal Computing BBS at 617-461-4258 or via Internet anonymous FTP from: ftp.analog.com in /pub/tools/psa-sdk/.
Sound cards and manufacturers that use the PSA chip set include:
Echo Personal Sound System-Echo Speech Corp;
Cardinal Pro 16 (plus)-Cardinal Technologies;
Orchid SoundWave 32-Orchid Technology;
Wearnes Beethoven ADSP16 - Wearnes;
Western Digital Paradise 16-DSP-Western Digital; and
Adaptec AME-1570 - Adaptec.

[2]Forrer, Johan B., KC7WW, "Programming a DSP Sound Card for Amateur Radio," *QEX*, Aug 1994, pp 9-19.

[3]For an example to code an efficient FIR filter, see *ADSP-2100 Family User's Manual*, Prentice-Hall, 1993, Chapter 14.

[4]The demonstration software is available in the file QEXHFM.ZIP from the ARRL BBS (203-666-0578) or via Internet by anonymous FTP to ftp.cs.buffalo.edu in the /pub/ham-radio directory.

[5]The author's PSATOR and PSA-PACTOR programs are available for downloading. See note 4.

A Better A/D and Software for the DDC-Based Receiver

Better dynamic range results from adding a higher-resolution A/D converter to KC1HR's digital receiver

by Peter Traneus Anderson, KC1HR

This article describes two upgrades to my simple SSB receiver using a digital down-converter (DDC): an improved analog front end and A/D converter and a simple control program.[1] The improved front end increases the nominal dynamic range from 36 dB to 60 dB. The control program permits tuning the receiver from the keyboard of a PC.

As before, this is intended as a starting point for further improvements and to encourage experimentation with new technology.

Fig 1 shows the improved band-pass filter and A/D converter. This circuitry replaces Fig 4 in Note 1. The input signal passes through a variable band-pass filter and a two-stage amplifier to the A/D converter. The A/D output is applied to the DDC input.

The signal strengths at my antenna are low enough so that the increased dynamic range of the new A/D eliminates, for me, the need for an input attenuator. In strong-signal areas, an attenuator will still be needed.

The band-pass filter includes coils L1 and L2 and a dual 400-pF variable capacitor. Plug-in coils L1 and L2 are inductively coupled. The frequency range covered is changed by changing the values of L1 and L2.

Fig 2 shows L1 and L2 for three bands: 0.55 to 1.6 MHz, 3.4 to 11 MHz and 14 to 22 MHz.

The first band provides coverage of the AM broadcast band. This is useful for testing and demonstrations, as stable signals are always available.

The second band provides coverage of the 80, 40 and 30-meter amateur bands, as well as the 49, 41 and 31-meter shortwave broadcast bands.

The third band provides coverage of the 20, 17 and 15-meter amateur bands. The signal frequency is between one-half and one times the A/D clock frequency. This results in the signals being aliased down to frequencies between zero and one-half of the clock frequency. The effect is identical to what would happen if a mixer were placed between the band-pass filter and the A/D input, with the mixer's LO port being driven by the A/D clock frequency. A signal appears to the DDC to be at a frequency equal to the clock frequency minus the signal frequency, and the sidebands are swapped. Thus, an upper sideband (USB) signal at 21.3 MHz appears to the DDC as a lower sideband (LSB) signal at (25 MHz–21.3 MHz) = 3.7 MHz. A USB signal at 14 MHz appears to the DDC as an LSB signal at (25 MHz–14 MHz) = 11 MHz.

For the 0.55 to 1.6-MHz band, L1 is omitted. A loop antenna from an old AM radio is used for L2. Strong local stations will overload the A/D. Detuning the variable capacitor or orienting the loop to put a null in the direction of the strong station will eliminate the overload.

I added a two-turn link, connected between the antenna-input terminal of L1 and the ground terminal of L2, to permit using an external antenna for weak stations.

For the 3.4 to 11-MHz band, L1 and L2 are each a pair of 3.3 μH, axial-lead miniature molded RF chokes in series. The two chokes in L1 are aligned axially, as are the two chokes in L2. The two pairs of chokes are placed close together with the choke bodies touching to provide coupling from L1 to L2.

[1]Notes appear on page 15.

990 Pine Street
Burlington, VT 05401
email: traneus@emba.uvm.edu(Internet)

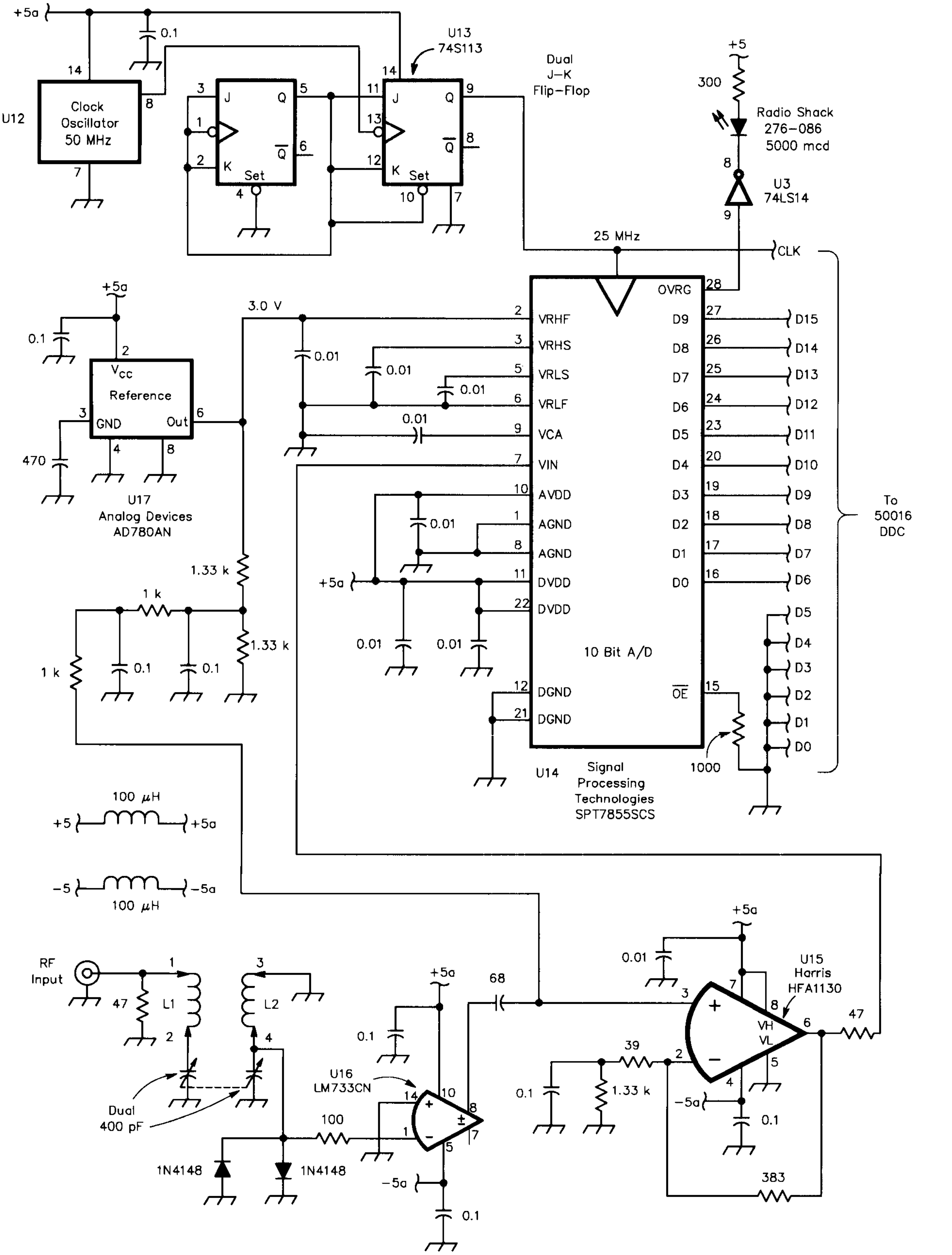


Fig 1—Improved band-pass filter and A/D converter. This circuitry replaces Fig 4 in Note 1. The filter is sharper due to looser coupling. See Fig 2 for values of plug-in coils L1 and L2. The A/D converter is upgraded to 10 bits.

For the 14 to 22-MHz band, L1 and L2 are each a pair of 3.3 µH axial-lead miniature molded RF chokes in parallel. The four chokes are placed close together with choke bodies touching to provide coupling from L1 to L2.

The two-stage preamplifier provides a gain of 55. The gain is reduced from the previous design to keep the input voltage corresponding to one count in the A/D roughly the same as before.

The first stage, U16, is an LM733 IC RF amplifier run with all gain pins open, to give a gain of 5. The 733 is an old, cheap, readily available 90-MHz bandwidth differential preamplifier.

The second stage, U15, an HFA1130P from Harris Semiconductor, is a high-speed current-mode operational amplifier (op amp) configured for an ac gain of 11 and a dc gain of unity.[2]

In a conventional (voltage-mode) op amp, the output voltage is proportional to the voltage difference between the noninverting input and the inverting input. Both inputs have high impedance and draw very little current. The output drives an external feedback network to maintain the voltage difference between the two inputs at zero.

In a current-mode op amp, the noninverting input is a high-impedance input that drives an ideal emitter follower. This follower is made using NPN and PNP transistors so it can provide output currents of either polarity, with zero base-to-emitter voltage drop.

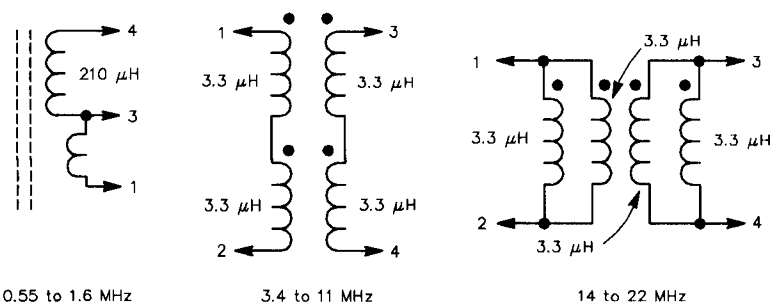


Fig 2—L1 and L2 for three bands: 0.55 to 1.6 MHz, 3.4 to 11 MHz and 14 to 22 MHz. The highest band drives one of the alias responses of the A/D to provide USB reception instead of LSB reception.

ARRL Lab Tests of the DDC-Based Receiver

ARRL Laboratory Engineer Mike Gruber, WA1SVF, tested the improved receiver in the ARRL Lab, while I assisted by operating the receiver.

All of the tests were performed with the receiver set for 400-Hz bandwidth, using the 10-bit A/D converter. The results follow:

Minimum Detectable Signal (MDS):

At 3.520 MHz, MDS = –129 dBm
At 14.020 MHz, MDS = –105 dBm
At 21.020 MHz, MDS = –129 dBm

Because the MDS was unexpectedly higher at 14 MHz, we did not perform the other tests at 14 MHz.

Third-Order Intermodulation Distortion Dynamic Range (IMDDR):

At 3.520 MHz, IMDDR = 57 dB
At 21.020 MHz, IMDDR = 70 dB

Because the A/D is only 10 bits linear, I expected about 60-dB IMDDR. We measured IMDDR a second time at 21 MHz. The band-pass filter tuning sometimes affected 21-MHz IMDDR and sometimes didn't. The second measurement at 21 MHz gave 73 dB.

Blocking Dynamic Range (BDR):

At 3.520 MHz, BDR = 91 dB
At 21.020 MHz, BDR = 91 dB

The BDR was limited by overflow in the A/D.

The output of the emitter follower is actually the inverting input. Thus, the emitter follower forces the inverting-input voltage to be equal to the voltage on the noninverting input. The current needed to do this forcing comes from the collector of the emitter follower. The collector current is forced to flow through a capacitor internal to the op amp. The voltage across this capacitor is buffered by another ideal emitter follower to provide the op-amp output. The output drives an external feedback network to maintain the inverting input current at zero.

The current-mode op amp used here provides two advantages over a voltage-mode op amp: more gain at high frequencies and output voltage clamping.

Voltage-mode op amps exhibit constant gain-bandwidth. The product of the closed-loop gain and the closed-loop bandwidth is a constant. Thus, if you want more gain, you must accept less bandwidth.

Current-mode op amps exhibit constant bandwidth. The closed-loop bandwidth is roughly constant, independent of the closed-loop gain. Thus, the gain-bandwidth increases as the closed-loop gain increases, within limits. The bandwidth does vary with the resistance of the feedback resistor. (See the HFA1130 spec sheet for details.)

The voltage-clamping feature is set by the wiring of pins 8 (+limit) and 5 (–limit) to keep U15's output voltage between ground and +5 V. This is necessary, as the CMOS A/D can be damaged by input voltages below ground.

U17, an AD780AN from Analog Devices, provides a stable 3-V reference for the A/D converter.[3] The full-scale range of the A/D analog input extends from zero to the reference voltage. A pair of resistors sets the dc operating point of the preamplifier output to be half-scale of the A/D input.

Oscillator U12 and flip-flop U13 provide a stable 25-MHz square-wave clock to the A/D and the DDC. Both the A/D and the DDC respond to the rising edge of the clock.

The 10-bit A/D converter, U14, is a SPT7855SCS from Signal Processing Technologies.[4] This is a 25-mega-sample-per-second (MSPS) CMOS A/D. Internally, it is 16 matched 10-bit succesive-approximation A/Ds running in time-staggered fashion. Each internal A/D converts one input sample every 16 clocks. Overall, one sample is converted during each clock,

```
#include <stdio.h>
#include <stdlib.h>
#include <string.h>
/* PC control program for 50016 receiver @ 25 MHz */
/* 20aug94 PTAnderson KC1HR */

main()
{
   int tempint = 0;
   int gain_state = 0;
   int j = 0;
   int c = 'n';
   int exit_char = '!';
   char preamble[] = "20011";
   char postamble[] = "00001";
   char ph_inc_string[] = "20011011001100101110101010001011110010001";
   long maxfreq = 12000000;
   long freq = 5000000;
   float tempfloat;
   double dfreq;
   double dsfreq;
   double dafreq;
   double dph_inc;
   long ph_inc;
   double fclock = 25000000.0;
   double two_up32_over_fclock;
   two_up32_over_fclock = 4294967296.0 / fclock;

/* initialize receiver to minimum gain */
   fprintf(stdprn,"@A\n");

/* initialize control registers 2 and 3 (same data for SSB or CW) */
   fprintf(stdprn,"20100000000000000000000000000000000000000\n");
   fprintf(stdprn,"20110000000000000000000000000000000000000\n");

/* print help table */
   printf("\nSimple SSB receiver set for lower sideband reception");
   printf("\n\ncenter of RF passband = frequency - 1.75 kHz");
   printf("\n\nexit to DOS:    set DDC clock kHz:     set frequency kHz:");
   printf("\n\n!                     #                          =");
   printf("\n\nup frequency:");
  printf("\n\n q=1MHz w=100kHz e=10kHz r=5kHz t=1kHz y=100Hz u=10Hz i=1Hz");
   printf("\n\ndown frequency:");
  printf("\n\n a=1MHz s=100kHz d=10kHz f=5kHz g=1kHz h=100Hz j=10Hz k=1Hz");
   printf("\n\ngain:                                      bandwidth Hz:");
   printf("\n\n   z=up   x=dn     c=min                     n=2000 m=400");
   printf("\n\nlast command, gain, alias, frequency in kHz now are:\n\n");

/* while not exit char */
   while(c-exit_char){

     switch (c)
     {
       case 'z':            /* up gain */
   if( gain_state > 14 )
     gain_state = 15;
   else
   {
     gain_state = gain_state + 1;
     fprintf(stdprn,"C");
   }
   break;
       case 'x':            /* dn gain */
   if ( gain_state < 2 )
   {
     gain_state = 0;
     fprintf(stdprn,"A");
   }
   else
   {
     gain_state = gain_state - 1;
     fprintf(stdprn,"B");
   }
   break;
       case 'c':            /* min gain */
   gain_state = 0;
   fprintf(stdprn,"A");
   break;
       case '=':            /* set freq in kHz*/
   printf("\rfreq kHz ");
   cscanf( "%f", &tempfloat );
   freq = 1000.0 * tempfloat;
   tempint = getch();
   break;
       case '#':            /* set fclock in kHz */
   printf("\rclock kHz ");
   cscanf( "%f", &tempfloat );
   fclock = 1000.0 * tempfloat;
   two_up32_over_fclock = 4294967296.0 / fclock;
   tempint = getch();
   break;
       case 'n':                  /* 2000 Hz bandwidth */
  fprintf(stdprn,"21000000010000000000000000000000000101001\n");
  fprintf(stdprn,"21010000011011111001111111011100011100101\n");
  fprintf(stdprn,"21100000000011101010100100100000001011010\n");
break;
       case 'm':                  /* 400 Hz bandwidth */
  fprintf(stdprn,"21000000010000000000000000000000000010011\n");
  fprintf(stdprn,"21010010001011100000101001100110100100101\n");
fprintf(stdprn,"21100000000011101010100100100000111001000\n");
break;
       case 'i':                  /* up 1Hz */
freq = freq + 1 ;
break;
       case 'k':                  /* dn 1Hz */
freq = freq - 1 ;
break;
       case 'u':                  /* up 10Hz */
freq = freq + 10 ;
break;
       case 'j':                  /* dn 10Hz */
freq = freq - 10 ;
break;
       case 'y':                  /* up 100Hz */
freq = freq + 100 ;
break;
       case 'h':                  /* dn 100Hz */
freq = freq - 100 ;
break;
       case 't':                  /* up 1kHz */
freq = freq + 1000 ;
break;
       case 'g':                  /* dn 1kHz */
freq = freq - 1000 ;
break;
       case 'r':                  /* up 5kHz */
freq = freq + 5000 ;
break;
       case 'f':                  /* dn 5kHz */
freq = freq - 5000 ;
break;
       case 'e':                  /* up 10kHz */
freq = freq + 10000 ;
break;
       case 'd':                  /* dn 10kHz */
freq = freq - 10000 ;
break;
       case 'w':                  /* up 100kHz */
freq = freq + 100000 ;
break;
       case 's':                  /* dn 100kHz */
freq = freq - 100000 ;
break;
       case 'q':                  /* up 1MHz */
freq = freq + 1000000 ;
break;
       case 'a':                  /* dn 1MHz */
freq = freq - 1000000 ;
break;
       default:
         break;
     }
     if ( freq > maxfreq )
       freq = maxfreq;
     if ( freq < 10000 )
       freq = 10000;
     dfreq = freq;
     dsfreq = 0.001*dfreq;
     dafreq = 0.001*(fclock-dfreq);
     dph_inc = (dfreq-1750) * two_up32_over_fclock;
     ph_inc = dph_inc;
     dph_inc = ph_inc;
    printf("\r%c %2.2u %9.3f %9.3f        ", c, gain_state, dafreq,
dsfreq);
     strcpy( ph_inc_string, preamble );
     for( j = 0; j < 31; j = j + 1 )
     {
       ph_inc = ph_inc << 1;
       if( ph_inc < 0 )
strcat( ph_inc_string, "1" );
       else
strcat( ph_inc_string, "0" );
     }
     strcat( ph_inc_string, postamble );
     for( j = 0; j < 41; j = j + 1 )
     {
       fprintf(stdprn,"%c", ph_inc_string[j]);
     }
     fprintf(stdprn,"\n");
     c = getch();
   }
 }
```

←**Fig 3—Software listing of the control program. This program was compiled in *Microsoft Quick C*. Other C compilers should work with few or no changes.**

but the data appears 16 clocks after the input is sampled.

The 10 data bits from the A/D go to the high 10 bits of the DDC data input. The low 6 bits of the DDC data input remain grounded.

The overrange output of the A/D is buffered by a spare inverter to drive a high-brightness LED. The LED is on when the A/D overflows. The A/D overflows a lot less than the old 6-bit A/D did.

Analog Devices has just introduced a 12-bit, 30 MSPS A/D, the AD9026, costing $238 in thousand quantity.[3,5] Unlike some older 12-bit high-speed A/Ds, this A/D is specified to show good linearity at full speed, giving 72 dB of dynamic range. This A/D has a full-scale input range of −1 V to +1 V, so level shifting and input clamping are not needed. A gain-of-nine amplifier, ac-coupled to the A/D input, should work fine with the band-pass filter and square-wave clock of Fig 1.

Fig 3 is a listing of a simple program to control the receiver. It is written in C. I used *Microsoft Quick C*, version 2.5. Other C compilers should work with few or no changes. Any differences will most likely be found in the library input and output functions.

The program provides commands to increment and decrement the operating frequency by various amounts, to directly enter the frequency, to raise and lower the gain and to change the bandwidth to 2000 or 400 Hz.

The program starts by declaring and initializing variables, then initializes the receiver, prints out a command list and enters the main loop.

The commands are all single characters. The current command character is initialized to 'n,' which means "set bandwidth to 2000 Hz." In the main loop, the current command character is compared to each of the valid commands and the appropriate command, if any, is executed. Thus, the first command executed is to set the bandwidth to 2000 Hz.

After executing a command, the program recalculates the current frequency in kilohertz, the new alias in kilohertz and the new phase increment. The phase increment is calculated as a 31-bit integer, as a long integer is 31 bits plus sign.

The program then prints out the current command character, gain setting, alias and frequency.

Next, the phase increment is converted to the 41-character string format needed by the receiver: a 5-character preamble, 31-character phase increment (most-significant bit first) and a five-character postamble.

The phase increment is converted to characters by left-shifting each bit into the sign bit. The sign bit is then tested to determine if the character should be 1 or 0. Another 0 is included in the postamble to provide the 32nd (least-significant) bit of the phase increment needed by the DDC.

The 41-character string is then sent to the receiver, followed by a newline character string (carriage return and line feed).

Lastly, the program waits for a new command character to be typed before returning to the start of the loop.

As with the hardware, this program is intended as a starting point to be improved upon.

I wish to thank Gil Gianetti, N1FEB, for his support and assistance in the development of this receiver.

Notes

[1]Anderson, P. T., "A Simple SSB Receiver Using a Digital Down Converter," *QEX*, Mar 1994, pp 3-7.

[2]Harris Semiconductor, 1301 Woody Burke Road, Melbourne, FL 32902, tel: 407-724-3000.

[3]Analog Devices, One Technology Way, PO Box 9106, Norwood, MA 02062-9106, tel: 617-329-4700.

[4]Signal Processing Technologies, 4755 Forge Road, Colorado Springs, CO 80907, tel: 719-528-2300.

[5]Wirbel, L., "TI rolls DSP for cellular; ADI offers converters," *Electronic Engineering Times*, July 18, 1994, p 83.

Practical Microwave Antennas

Part 3—Lens antennas and microwave antenna measurements.

by Paul Wade, N1BWT

In the previous parts of this series we discussed horn antennas and parabolic dish antennas. We now turn our attention to the third type of practical microwave antennas: lenses. I'll also describe microwave antenna measurement techniques and conclude with a discussion of actual measurement results and a comparison of the three types of antennas.

Lens Antennas

For portable microwave operation, particularly if backpacking is necessary, dishes or large horns may be heavy and bulky to carry. A metal-plate lens antenna is an attractive alternative. Placed in front of a modest-sized horn, the lens provides some additional gain, much like eyeglasses on a near-sighted person. The lens antennas I have built and tested are cheap and easy to construct, light in weight and noncritical to adjust. The *HDL_ANT* computer program makes designing them easy, as well.

There are other forms of microwave lenses—for instance, dielectric lenses and Fresnel lenses—but the metal-plate lens is probably the easiest to build and lightest to carry, so it is the only type I'll describe here.

The metal-lens antenna is constructed of a series of thin metal plates with air between them. The curvature of the edges of the plates forms the lens, and the space between the plates forms a series of waveguides. Fortunately, we can get "air" in a solid form to make construction easier: Styrofoam looks just like air to RF, and it keeps the metal plates accurately spaced. We use aluminum foil for the plates, attaching it to the Styrofoam with spray adhesive and shaping the curvature with a hobby knife on a compass. Designs are limited to those using circles, to ease construction.

Background

These metal-plate lenses were originally described for 10 GHz by KB1VC and me at the 1992 Eastern VHF/UHF Conference, but there is no good reason to limit them to that band.[1] The need for more gain became apparent to us during the 1991 10-GHz Contest. We were atop Burke Mountain in Vermont, on a day as clear as the tourist brochures promise. We could see Mt. Greylock in Massachusetts, where KH6CP was located, but it was too far to work with horn antennas on our Gunnplexers. After K1LPS humped his two-foot dish up the fire tower, we knew that wasn't the best answer for portable work.

Later, we found an article in *VHF Communications* on lens antennas by Angel Vilaseca, HB9SLV, which intrigued us.[2] It described how to design a metal-lens antenna but did not present expected gain or measured results.

We then searched through the references to try to understand how these antennas work, finally discovering that the best work was done before we were born, by Kock.[3] Kock's paper makes it clear how the metal lens antenna works, and, more importantly, that it *does* work!

Lens Basics

The metal plate lens works, in principle, like any other lens. A similar optical lens would take a broad beam of light and shape it, by refraction, into a narrower beam.[4] Refraction occurs at the interface of two materials in which light travels at different

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Shirley, MA 01464

[1]Notes appear on page 24.

speeds and changes the direction of travel of the beam of light. If the beam is formed of many rays of light, each one may be bent; the ones at the edge of the beam bend more so they end up parallel to the center rays, which are hardly bent at all. For this to work, each ray must take exactly the same time to travel from its source, at the focal point of the lens, to its destination. Since light travels more slowly in glass, a lens is thicker at the middle, to slow down the rays with a shorter path, and thinner at the edges, to allow the rays with longer paths to catch up, as shown in Fig 1. The needed curvature of the lens to form the beam exactly is an ellipse, but for small bending angles a circle is almost identical to an ellipse, and nearly all optical lenses are ground with spherical curves.

Since light and RF are both electromagnetic waves, we could use glass—or any other dielectric—to make a lens for 10 GHz. For example, a recent article described a dielectric lens made of epoxy resin.[5] But for larger sizes this quickly becomes less attractive, and most dielectrics are rather lossy at 10 GHz. Low-loss materials are available but are costly and relatively heavy and difficult to shape.

The Metal-Plate Lens

Since electromagnetic waves travel at different speeds in waveguide and in free space, why not use waveguides of different lengths to form a lens? This has been done and is known as an "eggcrate" lens.[6] However, it is easier to make a group of parallel plates that form wide parallel waveguides, simply shaping the input and output edges of these waveguides to change the path lengths and form the lens surface. This differs from an optical lens in that the phase of the electromagnetic wave travels *faster* in a waveguide than in free space.[7] Thus, the required curvature of a metal lens antenna is the *opposite* of an equivalent optical or dielectric lens—in this case, concave instead of convex. We can still get away with using circular curvatures instead of ellipses as long as we aren't trying to bend the rays too sharply. For that reason, we feed the lens with a small horn, which does part of the beam forming, as shown in Fig 2. Of course, if we want both horizontal and vertical beam shaping, we need a spherical shape, so we must shape the surface described by the edges of the metal plates into a sphere like that of Fig 3.

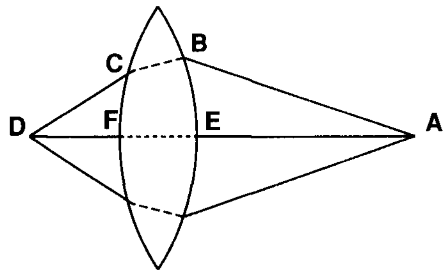

Fig 1—A simple lens. The travel time for each of the rays must be the same, so the time along the line ABCD is the same as that along the line AEFD.

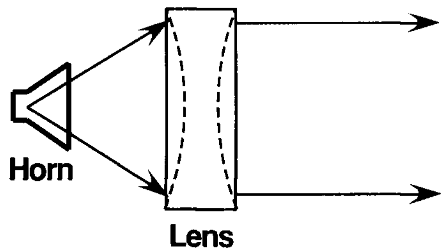

Fig 2—Feeding a lens with a horn lets the horn provide part of the beam shaping.

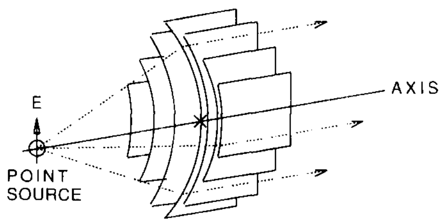

Fig 3—A spherical lens can be formed by a series of spaced plates.

Lens Design

While the *HDL_ANT* program removes the drudgery from lens design and makes it available to amateurs, a general description of lens design might aid in understanding what is happening and what the computer is telling you.

First, some design objectives are needed: how big a lens is desired, and what are the dimensions of the horn feeding it? Gain is determined by aperture (roughly the diameter for dishes, horns and lenses). A good rule of thumb is that doubling the aperture will increase the gain by 6 dB. For instance, an 8-inch lens in front of a 4-inch horn would add 6 dB to the gain of the horn, and a 16-inch lens would add 12 dB. So, modest gain improvements take modest sizes, but really large gains require huge antennas no matter what kind. However, a 6-dB increase in gain will double the range of a system over a line-of-sight path.

The horn dimensions may be determined by availability, or you may have the design freedom to build the horn as well. The beam width of the horn (which is usually smaller than the physical flare angle of the horn) is used to determine the focal length of the lens. Kraus gives the following approximations for beam width in degrees and dB gain over a dipole:[8]

$$W_{\mathrm{Eplane}} = \frac{56}{A_{E\lambda}} \qquad \text{Eq 1}$$

$$W_{\mathrm{Hplane}} = \frac{67}{A_{H\lambda}} \qquad \text{Eq 2}$$

$$Gain \approx 10\log_{10}(4.5A_{E\lambda}A_{H\lambda}) \qquad \text{Eq 3}$$

where $A_{E\lambda}$ is the aperture dimension in wavelengths in the E-plane, and $A_{H\lambda}$ is the aperture in wavelengths in the H-plane. These approximations are accurate enough to begin designing. From the beam width and desired lens aperture, finding the focal length f is a matter of geometry:

$$f = \frac{Lens\ diameter}{2\tan\left(\frac{W_{\mathrm{Eplane}}}{2}\right)} \qquad \text{Eq 4}$$

The final and most critical dimension is the spacing of the metal plates. The blue Styrofoam sheets sold as insulation have excellent thickness uniformity, and ¾ inch is pretty near optimum for 10 GHz, but the actual dimension should be measured carefully. The thickness determines the index of refraction:

$$index = \sqrt{1-\left(\frac{\lambda_0}{2\times spacing}\right)^2} \qquad \text{Eq 5}$$

which is the ratio of the wavelength in the lens to the wavelength in free space.

Next comes calculation of the lens curvature. The optimum curve is an ellipse, but we know that spherical lenses have been used for optics since Galileo, so a circle is a usable approximation. We can show that the circle is an excellent fit if the focal length is more than twice the lens diameter; photographers will recognize this as an f/2 lens. This suggests that the feed horn have a beam width of no more

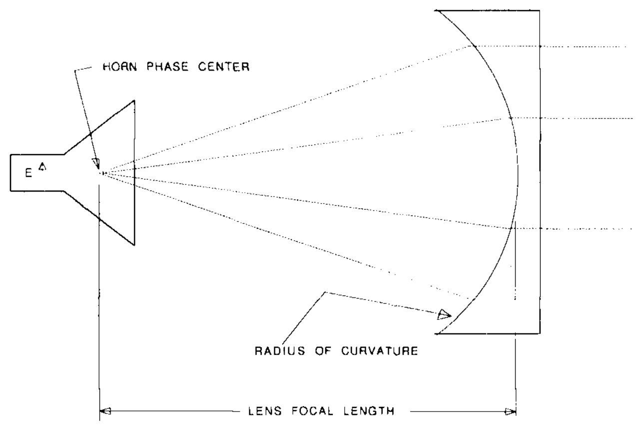

Fig 4—A single-curved lens. The radius of curvature is found using Eq 6, with the radius of the flat side set to infinity.

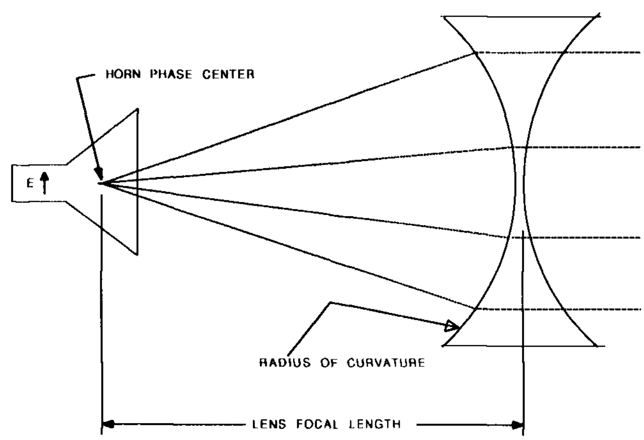

Fig 6—A double-curved lens. *HDL_ANT* provides both single-curved and double-curved lens designs.

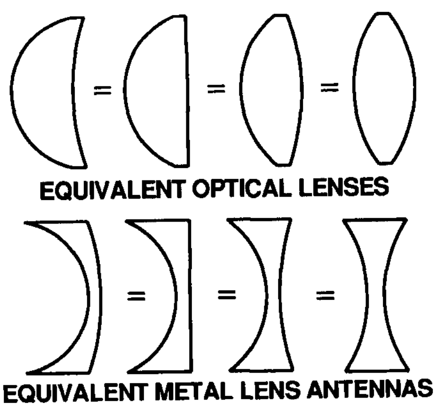

Fig 5—Each of the lenses shown has the same focal length, per Eq 6.

than 28°, or a horn aperture of at least 2 wavelengths.

The radius of curvature of the two lens surfaces is calculated from the lensmaker's formula (see Note 4):

$$\frac{1}{f} = (index - 1)\left(\frac{1}{R_1} - \frac{1}{R_2}\right) \qquad \text{Eq 6}$$

where a negative radius is a concave surface. For the single-curved surface of Fig 4, one radius is set to infinity. All combinations of R1 and R2 that satisfy the formula are equivalent, as shown in Fig 5; the computer program calculates the single-curved and symmetrical double-curved solutions (Fig 6). The radius of curvature calculated above is for the surface, and thus the central plate, which has the full curvature. The rest of the plates must be successively wider and have smaller radii so that the edges of all the plates form a spherical lens surface. This is more geometry, and the program does the calculations for each plate.

The final calculation involves the phase-centers of the horn, so that the lens-to-horn distance matches the focal length. This is a difficult calculation involving calculation of Fresnel sines and cosines; KB1VC deserves credit for the programming.[9,10] Without a computer, you would use trial-and-error looking for best gain. What the calculations will show is that many horns, particularly the "optimum" designs, have much different phase centers in the E- and H-planes. The program offers to make a crude compensation for this, but, if possible, the H-plane aperture of the horn should be adjusted slightly to match the phase centers. A few trial runs of the program should enable you to find a good combination. If you already have a horn, either try the compensation or just use the E-plane phase center.

For very large lenses, the size may be reduced by stepping the width of the plates into zones which keep transmission in phase, as shown in Fig 7. The program will suggest a step dimension if it is useful. At 10 GHz, a step is useful only for lenses larger than 2 feet in diameter.

Construction

Construction is straightforward, using metal plates of aluminum foil spaced by Styrofoam, as suggested by HB9SLV (see Note 2). A 2-foot by 8-foot sheet of blue Styrofoam, ¾-inch thick, is less than five dollars at the local lumberyard and will make several antennas. The aluminum foil is attached to the foam using artist's spray adhesive, available at art supply stores. Spray both surfaces lightly, let them dry for a minute or two, then spread the foil smoothly on the foam. If the adhesive melts the foam, you are using too much.

Next mark the outline of a rectangle for each metal plate on the foil. These will be used later to cut the foam and line up the plates, so they should all be the same size. Then mark the center of each curve and measure off the radius to the center of the circle. Using a compass with a hobby knife attached, place the point at the center of the circle and cut the curve through the foil into the foam. When all the curves are cut, peel off the unwanted foil, leaving the lens plates. Then cut up the rectangles with a razor blade and stack the blocks into a lens. (You did number them, didn't you?) Each rectangle should have foil on one side. If it looks good, glue them up two at a time. The final antenna will be a block of foam—there is no need to shape the foam to the lens curve. Shrink-wrapping the lens with thin plastic makes nice weatherproofing.

A few helpful hints are in order. Sharp knife blades really help in this process, and permanent markers don't smear. Also, if the foam is cut halfway through, it will snap cleanly on the line.

Adjustment

A metal-lens antenna only works in the E-plane. This is parallel to the elements of a dipole or Yagi but perpendicular to the wide dimension of a waveguide. The plates *must* be

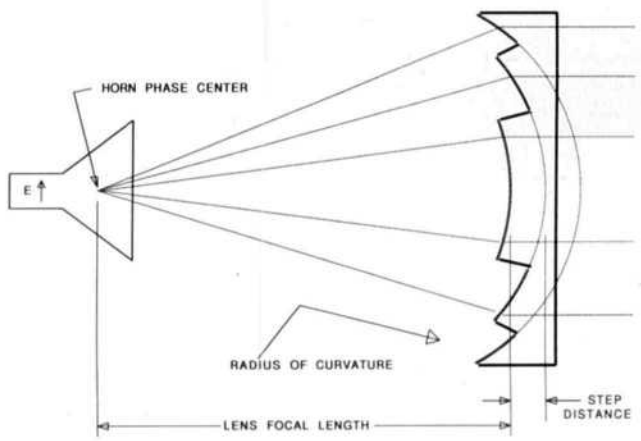

Fig 7—A zoned lens can be used to implement large lenses, reducing the needed thickness.

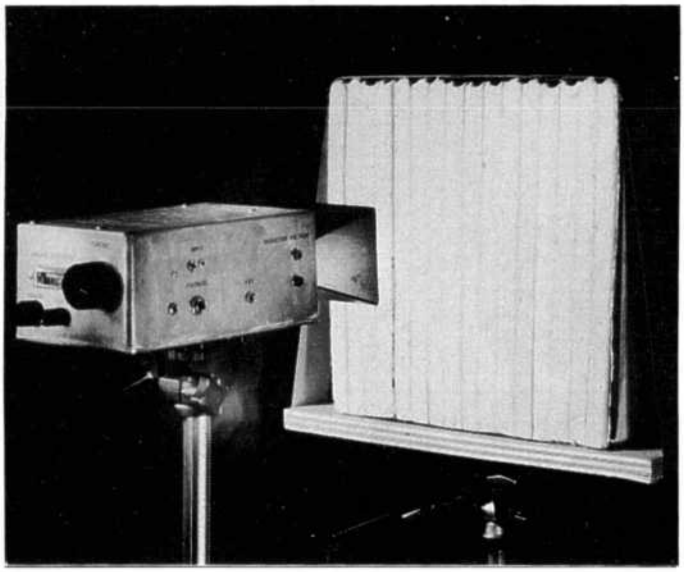

Fig 8—A 300-mm lens placed in front of a Gunnplexer transceiver provides about 10 dB of additional gain over that of the horn alone.

perpendicular to the wide dimension to provide gain.

The horn should point through the center of the lens, but the focus distance is not as critical as with a dish. Aiming is done by pointing the feed horn; the lens focuses the beam more tightly but does not change the beam direction. Tilting the lens will *not* steer the beam—if you don't believe this, take an optical lens, like a magnifying glass, focus it on something, and tilt it.

We found that the best gain was with the horn slightly closer to the lens than calculated, probably because of edge effects. Making the size of the plates slightly larger than calculated would probably eliminate this effect and make the gain a bit higher; since a wavelength at 10 GHz is about an inch, an inch or two oversize is plenty.

One other interesting effect was found with Gunnplexers: since the transmitter is also the receiver local oscillator, reflected power from the lens adds to the LO power, or subtracts when out-of-phase. This makes the received signal strength vary with every half-inch change in lens-to-horn distance, with very little change in signal strength observed at the other station. So, adjust the spacing for the best received signal. Of course, this effect does not exist on a system with low LO radiation.

Using the HDL_ANT *Program*

The lens section of *HDL_ANT* calculates the dimensions for the plates of a lens. Since all curves are circles that are easily drawn with a compass, templates are not generated. The basic input data is entered interactively, then results are presented in tabular form. If you like the results, they may be saved to a file for printing or further processing; if not, try another run with new data.

All dimensions are in millimeters. There are two reasons for this: the first is that odd fractions lead to errors in measurement, and the second is that one millimeter is a good tolerance for 10-GHz lens dimensions. If all measurements are made to the nearest whole millimeter, good results can be expected. The only exception is in the plate spacing, and that is accurately controlled by the foam thickness.

Results

We have constructed and tested three metal-plate lens antennas to date: a 150-mm single-curved version, and 150-mm and 300-mm double-curved versions. Fig 8 shows the 300-mm lens fed by a WBFM Gunnplexer system, and Listing 1 shows the *HDL_ANT* design of this antenna. All the lenses are designed to be fed with the standard Gunnplexer horn, which has well-matched phase centers, whether by design or by accident. Gain measurements are shown in Table 1. The lenses perform with about 50% efficiency if we consider them as having a round aperture; the corners do not contribute significantly, but we made them square for convenient fabrication and mounting.

We also used the lenses during several contests during 1992, 1993 and 1994. The 300-mm lens increased the range of our WBFM Gunnplexer transceivers by approximately 50%, to over 200 km, enabling contacts over new paths. The equipment was still highly portable due to the light weight of the lens, and they have survived mishaps with only a few harmless dents in the foam.

Further Uses for Metal Lenses

The metal-lens antenna could be useful at other frequencies: for 5.76 GHz a foam thickness of around 35 mm would be good, and at 24 GHz approximately 8-mm-thick foam might work, though it could be lossy at that frequency.

A lens can also be part of a more complex antenna system. For instance, a divergent lens can be used to provide better illumination for some of the very deep dishes that are sometimes available as surplus. A book on optics (such as Note 4) will show how to change the focal points appropriately.

Lens Summary

We have demonstrated that metal-lens antennas may be easily designed and constructed using the *HDL_ANT* computer program and that a book-sized lens, light and rugged enough for backpacking, provides gain enhancement adequate to double the range of a Gunnplexer system.

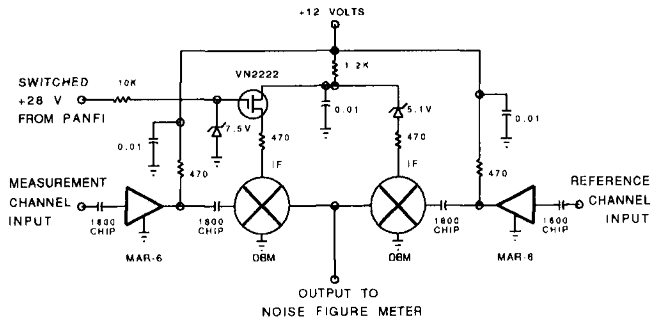

Fig 9—This switch can be used to automatically switch between the reference and measurement paths of the antenna gain measurement system.

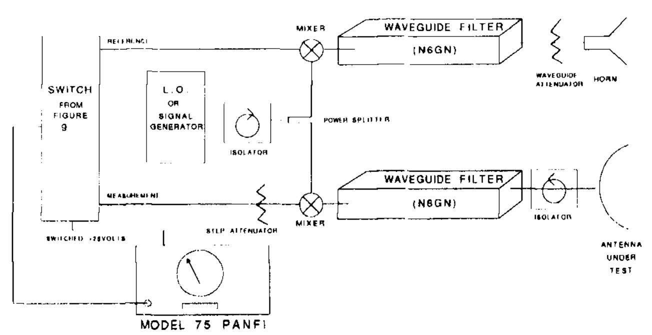

Fig 10—Test set-up for 10-GHz antenna measurements.

Antenna Gain Measurement

Hams have been measuring antennas for many years at VHF and UHF frequencies, and we have seen marked improvement in antenna designs and performance as a result. Very few serious antenna measurements have been made at 10 GHz; the additional difficulties at these frequencies are not trivial. I'll describe a few new twists that make it more feasible.

Overview

Antenna measurement techniques have been well described by K2RIW and W2IMU;[11,12] the latter also appears in the *ARRL Antenna Book* and is required reading for anyone considering making antenna measurements.[13]

The antenna range is set up for antenna ratiometry so that two paths are constantly being compared, both originating from a common transmitting antenna. (See Note 11.) These two paths are called the "reference" and "measurement" paths. The reference path uses a fixed antenna that receives what should be a constant level. In reality, there are continuous small fluctuations in the received signal at microwave frequencies, even over a short distance like an antenna range. Using ratiometry, the reference path allows these random variations in the source power or the path loss to be corrected by an instrument that constantly compares the signal from the measurement path with that of this reference path. First, a standard antenna with known gain is measured and the reading is recorded. Then, when an unknown antenna is measured, the difference between it and the standard antenna determines the gain of the unknown antenna.

Instrumentation

One measuring instrument commonly used is the venerable HP 416 ratiometer, with crystal detectors used to sense the RF. Basically, this technique compares the outputs of two crystal (diode) detectors. The crystal detectors present a problem: a matched pair is needed, and these are hard to find for 10 GHz. Also, diode detectors have poor sensitivity and dynamic range, so it is necessary to provide adequate power to keep the detectors operating in the square-law region where they are accurate. Another problem is drift in the old vacuum-tube HP 416.

It seemed to me that a superheterodyne technique was needed. If the signal could be converted to some lower frequency, it could be received on a better receiver. If the two channels had separate converters, the comparison could be made at the lower frequency. Finally, if we simply switched between the two channels at the lower frequency, the output would be an AM signal. If the switching rate were at an audio frequency, an AM receiver would thus have an audio output amplitude proportional to the difference in signal level between the two channels. Once the signals are combined by the switch, they may be easily amplified as needed at the lower frequency.

At 10-GHz, frequency stability is always a problem, so a normal communications receiver might be too sharp. From work with 10-GHz WBFM, I know that most signals are stable to within a few hundred kHz after warm-up, so a receiver with a 1-MHz bandwidth should be acceptable. While I was wondering if there might be something usable in a surplus catalog, it occurred to me that I already owned a perfectly usable solid-state wideband AM receiver—an AILTECH Model 75 Precision Automatic Noise Figure Meter (PANFI), which I found at a surplus auction. Not only that, it also has a synchronous detector and an output to synchronously switch the input signal (normally the noise source). The meter reads the difference in signal level as the input is switched; in this application, instead of the difference with a noise source switched on and off, it reads the difference as it switches between the two channels, with excellent resolution. If a signal much stronger than the noise is applied, the meter responds only to the signal rather than noise. (While checking the references for this paper, I discovered that K2RIW—see Note 11—had suggested use of the AILTECH Model 75 PANFI in 1976, but no one had remembered so I had to rediscover it!)

The only problem with the PANFI is that it is calibrated to solve the noise equation:

$$F_{dB} = T_{ex(dB)} - 10\log_{10}(Y-1) \qquad \text{Eq 7}$$

This requires a bit of arithmetic on a calculator or using the *HDL_ANT* computer program to undo the results and find the difference in dB:

$$Y_{dB} = 10\log_{10}\left(1 + 10^{(T_{ex(dB)} - F_{dB})/10}\right) \qquad \text{Eq 8}$$

Otherwise, the indicated gains are very optimistic.

The input to most noise figure meters is at 30 MHz, so I used a surplus signal generator to generate an LO 30 MHz away from 10368 MHz and used my 10-GHz transverter as the source transmitter. The signal generator provides the LO for two surplus waveguide diode mixers, but a pair of

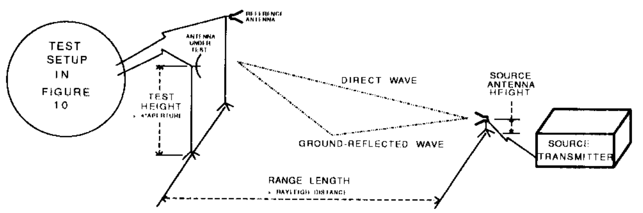

Fig 11—A ground-reflection antenna range with ratiometry. The required range length and antenna heights depend on the frequency and the characteristics of the antennas.

mixers like the ones in the transverter would also be fine.[14] Matched mixers aren't needed since they are linear mixers with wide dynamic range. I preceded each mixer with a band-pass filter, but there probably aren't many stray 10-GHz signals around. An isolator in the antenna line is useful when the antennas may have high VSWR.

Everything after the mixers is at 30 MHz, so ordinary cables and components complete the setup. I included a step attenuator in the measurement path to double-check the meter readings.

The switch, shown schematically in Fig 9, uses a common double-balanced mixer (DBM) as an attenuator in each path. Applying a dc current though the diodes in a DBM varies the attenuation; the DBM has high loss with no dc current, and low loss with dc current applied; I measured 54 dB of loss at 0 mA and 2.8 dB at 20 mA. An FET and some zener diodes provide a crude switching circuit to switch the current in response to the 28-V output from the PANFI.

Fig 10 shows the antenna measuring setup for 10 GHz. The reference path uses a small horn as the receiving antenna, and the source antenna is another horn. After completing all connections, the signal generator is adjusted for maximum received signal, as indicated with the PANFI switched to the noise OFF position. Then the PANFI is switched to AUTO to display the difference between the paths, which is converted to relative gain using the above equation.

Antenna Range

The length of the antenna range is important: if it is too short, there will be significant phase difference over the aperture of the antenna being tested, resulting in low measured gain. The minimum range length to avoid this error is the Rayleigh distance:

$$\textit{Rayleigh distance} = \frac{2D^2}{\lambda} \qquad \text{Eq 9}$$

A few trial calculations will show that this requires miles of range for large dishes. Fortunately, the Rayleigh distance for the 25-inch dish that I wanted to measure is only 91 feet at 10 GHz.

The antenna range is a ground-reflection range, as shown in Fig 11, where the range is designed to account for ground reflection and control it. One alternative would be to place the antennas high enough that ground reflection would be insignificant; however, in order to keep the reflected signal contribution from the ground to less than 0.5 dB, both ends of the range would have to be 122 feet high for a range length of 91 feet. Another type of range requires the signal path to be at a 45° angle to the ground, so the antenna height would only be 91 feet. For most amateur work, antenna heights like these are impractical, so the ground-

```
                ****************
                *              *
                *   HDL_ANT    *
                *              *
                ****************

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         Version 1.0E with option for the metric-impaired.

         Any licensed radio amateur may use this program without
         charge; all other persons must send $73 to the
         ARRL Foundation, 225 Main St., Newington, CT 06111

         Metal plate lens antenna for microwaves

           ALL dimensions are in millimeters!

              ------------
             |     E      |
             |     ^ -->H |    waveguide
             |     |      |
              ------------

      At a center frequency of 10.265 GHz

       For a lens with a diameter of 301.55 mm. and a plate spacing of
    18.847 mm.

       Fed by a horn of axial length = 76.2 mm,
           H-plane aperture = 90.5 mm
           E-plane aperture = 73 mm
            and a Gain of 17.41 dB over isotropic

       E-Plane phase center is 1.57 wavelengths inside horn mouth
       H-Plane phase center is 1.59 wavelengths inside horn mouth
      Calculations for an f/2.52 lens with a focal length of 761.33 mm.
           providing an estimated gain of 12.32 db over the horn

      Distance from horn mouth to center of lens curve is 715.38 mm.
```

```
Radius of Curvature of lens plates starting from center plate

 Plate   Single radius    double radius    plate width
 -----   -------------    -------------    -----------

 ctr      280.01 mm.       560.03 mm.      min
  1       279.38 mm.       559.71 mm.      min + 0.63
  2       277.46 mm.       558.76 mm.      min + 2.55
  3       274.25 mm.       557.16 mm.      min + 5.77
  4       269.67 mm.       554.93 mm.      min + 10.34
  5       263.68 mm.       552.04 mm.      min + 16.33
  6       256.16 mm.       548.49 mm.      min + 23.85
  7       246.99 mm.       544.26 mm.      min + 33.03
  8       235.95 mm.       539.35 mm.      min + 44.06


     ****** UNBELIEVABLY CRUDE GRAPHICS ******

             Single Curve          Double Curve

                     -----------          -------------
                      ----------           -----------
                       ---------     ^      ---------
      --                --------     |       -------
     |  |        <      --------   H |       ---+---
     |  |        horn   --------     |       -------
      --                ---------           ---------
     waveguide         ----------          -----------
                      -----------         -------------
                                               +
                                  centerline   +
                                               +
                     -----------          -------------
                      \  LENS  |          \   LENS    /
     waveguide         \ PLATE |     ^     \  PLATE  /
                        \      |     |      \       /
     -------    <        )<--->|   E |       )<-+->(
     |     |    horn    / min. |     |      /  min. \
     -------        -->/ thick.|        -->/  thick. \
                     /  /      |        /  /          \
                    /   -----------    /   -------------
                   /                  /
                    Radius of curvature
```

Listing 1

reflection range is used.

In order to have the phase error as low in the vertical plane as in the horizontal plane, the height of the antenna being measured must be at least *four* times its aperture diameter, which is 100 inches for the 25-inch dish.[15] I suspect that most amateur antenna ranges have had insufficient antenna height and consequently have had trouble measuring higher-gain antennas accurately. My first measurements, at a height of about 4 feet, showed lower than expected efficiency for the larger dishes. Raising the height made the measured efficiency greatest for the larger dish, as you would expect, since the feedhorn blocks a smaller percentage of the aperture.

The received energy should be at a maximum at the height of the antenna being measured. For a ground-reflection antenna range, this is controlled by the height of the source antenna:

$$h_{\text{source}} = \frac{\lambda}{4}\left(\frac{\textit{Range length}}{h_{\text{receiving}}}\right) \qquad \text{Eq 10}$$

which works out to about 3 inches for the 91-foot-long range with the receiving antenna 100 inches high. Therefore, by adjusting the reference and measurement antennas to over 8 feet high (easily done on the back of a pickup truck or a porch) and placing the height of the source and antenna around 3 inches, good, reliable and accurate results may be obtained (see Table 1).

The Standard-Gain Antenna

In order to measure meaningful antenna gains, an antenna with known gain is required. Recall that all measurements are relative to a known standard. A dipole is useless as a standard—its broad pattern receives so many stray reflections that repeatable readings are nearly impossible, and its gain is so much lower than a 30+ dB dish that equipment accuracy is a problem; few instruments are accurate over a 30-dB (1000:1 power ratio) range.

What is required is an antenna with a known gain, preferably a gain of the same order of magnitude as the antennas to be measured. At microwave frequencies, the gain of a horn antenna can be calculated quite accurately from the physical dimensions. The algorithm used in the *HDL_ANT* program will be accurate within about 0.2 dB if good construction techniques are used.

Table 1—Summary of 10.368-GHz Antenna Measurements (N1BWT 12/18/93, 5/14/94, 9/15/94)

Antenna	*Focal distance*	*Gain (dBi)*	*Efficiency*
Standard Gain Horn, (22.5 dBi calculated) Scientific-Atlanta Model 12-8.2, courtesy KM3T, gain thanks to John Berry, Scientific-Atlanta.		22.45	43%
Gunnplexer Horn (17.45 dBi calculated)		17.5	57%
+ 6" lens	~8"	20.9	45%
+ 12" lens	~21"	27.4	50%
Surplus horn (19.4 dBi calculated)		19.6	67%
W1RIL loop Yagi		16.0	
22" dish, *f/D* = 0.39, surplus, feed = 11 GHz Superfeed:*			
unmodified feed	8.25"	33.1	55%
with feedline to reflector		32.2	45%
modified feed		32.9	53%
25" dish, *f/D* = 0.45, Satellite City, with the following feeds:			
11 GHz Superfeed,*	10.875"	34.4	58%
Clavin feed	11.125"	34.1	54%
Rectangular Horn,			
E=0.9", H=1.38"	10.625"	33.7	49%
E=1.14", H=0.9"	10.625"	32.9	41%
WR-90 to coax Transition	11.0"	32.7	39%
WA1MBA log periodic	10.94"	32.4	37%
18" dish, *f/D* = 0.42, Satellite City, with the following feeds:			
11 GHz Superfeed,*	7.75"	31.7	60%
Clavin feed	7.875"	31.2	53%
Rectangular Horn E=0.9", H=1.38"		31.5	57%
WR-90 to coax Transition, rect flange		30.2	42%
WR-90 to coax Transition round flange, od = 2.15"		30.2	42%
Cylindrical horn with slotted choke ring	7.875" to choke ring	~28**	~26%**
WA3RMX Triband feed		~17**	
24" Commercial (Prodelin) dish antenna: feed is rectangular horn fed by WR-90 waveguide "shepherd's crook"		33.6	52%

RANGE LENGTH = 102 feet. $2D^2/\lambda$ = 91 feet. Test height ~ 8 feet.

FOCAL DISTANCE SENSITIVITY: each feed was adjusted for max gain. Gain was down 1 dB about ¼" either way from peak.

Notes: *11 GHz Superfeed is a Chaparral feedhorn for 11-GHz TVRO.
**These feeds were not positioned accurately—more gain is possible.

For even better accuracy, several companies make standard-gain horns with good calibration data. For 10 GHz, a standard-gain horn was lent to me by KM3T—he was lucky enough to find one surplus. Mr. John Berry of Scientific-Atlanta was kind enough to provide the gain calibration curve.

Measurements

Once the antenna range is designed and set up, it must be checked out before making actual measurements. This is best done with an antenna with a fairly broad pattern, like a medium-sized horn, as the test antenna. First, the attenuators are adjusted for a convenient meter reading. Then the field uniformity is probed by moving the test antenna horizontally and vertically around the intended measurement point. The indicated gain should peak at the center and should not vary significantly over an area larger than any antenna to be tested; the variation should be less than one dB. At this point, the height of the source antenna usually needs to be adjusted to get the vertical peak at the intended receiving height. Finally, the test antenna is held stationary, and calibrated attenuation steps are added in the test path to make sure the indicated gain (after correction if using a PANFI) changes by the amount of attenuation added. With a ratiometer, the attenuation must be added at the microwave frequency, but a PANFI system like Fig 10, with a linear mixer, allows the use of an IF attenuator; step attenuators are much easier to find (or build) for 30 MHz than for 10 GHz.

Now the range is ready to make measurements. The standard-gain antenna is inserted as the test antenna, aimed for maximum indication, and the attenuators adjusted for a meter reading that will keep expected gains within the range of the meter. All gain measurements will be the difference from this standard reading added to the gain of the standard-gain antenna. The standard-gain antenna is replaced by an antenna to be tested, the new antenna is aimed for maximum gain, and its indicated gain is recorded. The difference between this indicated gain and the standard reading, after correction, added to the known gain of the standard-gain antenna, is the gain of the test antenna. The reading with the standard-gain antenna should be checked frequently to correct for instrumentation drift; ratiometry with the reference antenna corrects for other sources of drift.

Measurement Results

I set up a 10-GHz antenna range in my yard that was 102 feet long, more than the Rayleigh distance, with the equipment described above. The received field was probed for uniformity, and the height of the source antenna was adjusted for a flat field at the required height. Then I was able to start measurements, using a standard-gain horn for comparison. The results are shown in Table 1.

The first thing that became apparent is that all adjustments on a dish are critical. In the field, looking at a tiny S-meter, it doesn't seem so difficult to point a dish with a beamwidth of only about 3 degrees. The PANFI, however, has a large meter with one dB expanded out to nearly an inch. On this meter, even tiny adjustments have obvious effects, demonstrating how touchy aiming a dish is.

The most critical dimension of a dish is the focal length—the axial distance from the feed to the center of the dish. A change of ¼-inch, or about a quarter-wavelength, changed the gain by a dB or more.

The critical focal length suggests that it is crucial to have the phase center of the feed exactly at the focus of the reflector. Since the phase center is rarely specified for a feedhorn, we must determine it empirically by finding the maximum gain on a reflector with known focal length, which we can estimate from templates for various f/D. Thus we can estimate the phase centers for all the feeds in Table 1.

For the Chaparral style feed horns, we can deduce some further information. Several different feeds were measured, with two different dimensions, and with adjustable choke rings. Regardless of where the choke ring was set, maximum gain occurred with the choke ring the same distance from the reflector. This implies that the phase center is controlled by the position of the choke ring, not the central waveguide. The version designed for 11-GHz TVRO use, with the gain shown in Table 1, has an apparent phase center in front of the choke ring, while a larger one, dimensioned for 10 GHz, has the apparent phase center behind the choke ring (inside the ring) and provides gain similar to the smaller one.

As for efficiency, none of the dish measurements in Table 1 exceeds 60%, and it is obviously easy to get efficiencies less than 50%. This suggests to me that the 55% quoted in the books is far from typical, and careful design and measurement is needed to reach or exceed it. As illustrated in Part 2 of this series, dishes with small f/D (less than 0.3) may be very difficult to feed efficiently.

On the other hand, several of our amateur feeds have higher efficiencies than the commercial dish antenna shown in Table 1. If you find a surplus dish with a feed, don't assume it is the best possible one—different applications may require optimization of other parameters. For instance, WA1MBA has been working on a broadband log-periodic feed. The efficiency at 10 GHz shown in Table 1 is rather poor by comparison, but having a single dish feed that offers reasonable performance at several amateur microwave bands is an exciting possibility.

Several of the dish measurements in Table 1 were made with a coax-to-waveguide transition as a feed—the open-ended waveguide flange acts as a small horn. This is not an optimum feed, as shown by the low efficiency, but it is one that is readily available for comparisons. If the feed for your dish does not perform significantly better than a plain waveguide flange, it can certainly stand improvement.

Measured horn and lens efficiencies are comparable to dish efficiencies, so we can conclude that all three types of antennas can provide the same gain for the same aperture area. This leaves us free to choose the type of antenna best suited to the application.

Conclusion

Horns, dishes and lenses are all high-performance microwave antennas well-suited for amateur communications. Horns are small, rugged, and reliable, good for rover operation; they may be supplemented with a lens acting as an "amplifier" for increased gain. Dishes offer the ultimate in gain, at the expense of size and narrow beamwidth.

Horn and lens construction is easily within amateur capability, but parabolic reflectors at microwave frequencies require construction accuracy that is difficult to achieve. A dish antenna using a manufactured reflector still requires careful attention to detail to realize high efficiency.

Amateur antenna gain measurement at 10 GHz with good results has been demonstrated using ratiometry, and a noise figure meter is a good solid-state replacement for a vacuum-tube ratiometer. Antenna gain measurements are valuable for making critical adjustments and for verifying that an antenna is providing the performance expected. Better antenna gain measurements should bring the same improvement to amateur microwave antennas that years of antenna measuring contests have brought to VHF and UHF antennas.

Acknowledgements

I'd like to thank Bob Egan, N1BAQ; Larry Filby, K1LPS; Dick Knadle, K2RIW; Barry Malowanchuk, VE4MA; Dave Pascoe, KM3T; Matt Reilly, KB1VC; Ken Schofield, W1RIL; Dan Thompson, N1IOL; and Tom Williams, WA1MBA, for their help, and Beth Wade, N1SAI, and Filomena Didiano for locating obscure references. I'd also like to remember the late Dick Turrin, W2IMU, who generously shared his vast knowledge of antennas and antenna measurement with many hams—he taught me more than I'll ever remember.

Notes

[1]Wade, P., N1BWT, and Reilly, M., KB1VC, "Metal Lens Antennas for 10 GHz," *Proceedings of the 18th Eastern VHF/UHF Conference*, ARRL, May 1992, pp 71-78.

[2]Vilaseca, Angel, HB9SLV, "Microwave Lens Antennas," *VHF Communications*, 3/90, pp 179-189.

[3]Kock, Winston E., "Metal-Lens Antennas," *Proc. I.R.E.*, Nov 1946, pp 828-836.

[4]Hecht, Eugene, *Optics*, Addison-Wesley, 1987.

[5]Franke, John M., WA4WDL, "Pour an Antenna for X-band," *73*, Feb 1991.

[6]Sletten, Carlyle J. (W1YLV), *Reflector and Lens Antennas*, Aertech House, 1988.

[7]We will not attempt to explain this seeming magic here. Refer to "phase velocity" in any good book on microwaves. Rest assured that no laws of physics are being violated.

[8]Kraus, John (W8JK), *Antennas*, McGraw Hill, 1956.

[9]Muehldorf, Eugen I., "The Phase Center of Horn Antennas," reprinted in Love, A. W., *Electromagnetic Horn Antennas*, IEEE Press, 1976.

[10]Abramowitz, Milton and Stegun, Irene A., *Handbook of Mathematical Functions*, Dover, 1972.

[11]Knadle, R. T., K2RIW, "UHF Antenna Ratiometry," *QST*, Feb 1976, pp 22-25.

[12]Turrin, D., W2IMU, "Antenna Performance Measurements," *QST*, Nov 1974, pp 35-41.

[13]Hall, G., K1TD, ed, *The ARRL Antenna Book*, ARRL, 1991, pp 27-43 to 27-48 (Also in the 1994 Edition.)

[14]Wade, P., N1BWT, "Building Blocks for a 10 GHz Transverter," *Proceedings of the 19th Eastern VHF/UHF Conference*, ARRL, May 1993, pp 75-85.

[15]Gillespie, E. S., "Measurement of Antenna Radiation Characteristics," in *Antenna Handbook: theory, applications, and design*, Y.T. Lo and S.W. Lee, editors, Van Nostrand Reinhold, 1988, page 32-22.

RF

by Zack Lau, KH6CP/1

10-GHz WBFM—Improved Designs

In trying to make long-distance wide-band FM (WBFM) contacts on 10 GHz, I've discovered the shortcomings of some of the existing designs. These designs are simple to implement using devices such as the M/A-COM Gunnplexer, but simplicity sometimes means performance compromises as well. For instance, the varactor diodes used in most Gunnplexers are much more sensitive at low tuning voltages than at higher voltages. This means you will get much more deviation at lower operating frequencies than at higher frequencies unless the modulating amplitude is changed. On the lab bench, it isn't too difficult to adjust a deviation control, but such simple tasks become more difficult out in the field, where you often have to both manipulate the radio's controls and hold the antenna in exactly the right spot to dodge the trees, which are excellent attenuators of X-band signals.

Many of the existing WBFM circuits are sensitive to supply voltage and don't really provide adequate idiot protection. Ideally, a portable setup should work with nominal 13.8-V supply voltages that drop down to 10.5 V or so and should survive reverse polarity. Proper operation over a wide range of supply voltage allows the use of sealed lead-acid batteries, which are often more cost effective than NiCads. Reverse polarity protection is particularly useful out in the field, where mistakes tend to occur more frequently. On the other hand, designing the circuits for low current drain isn't all that useful, since Gunn diodes typically draw 100 to 200 mA anyway. Saving a few mA in the supporting circuitry might get you a few more percent of battery life, but you can easily lose more than that in struggling to make a difficult contact because your gear isn't up to par.

Two other important factors to consider are interference resistance and operation in extremely high or low humidity. Not surprisingly, broadcast transmitters are often located on the very same mountains used for amateur 10-GHz work. So, unless you want to listen to TV sync buzz, it's a good idea to make your equipment as resistant to high power VHF/UHF signals as possible. Here in New England, rain is expected during 10-GHz contests, so circuitry that can operate despite a bit of stray moisture is preferred. Low-impedance circuits handle stray, moisture-induced current paths better than do high-impedance circuits. On the other hand, you can also get extremely dry weather at high elevations, and it would be nice if the gear doesn't fail when zapped with a little static electricity.

Taking all these factors into consideration, I decided that you really want to use low-impedance bipolar/diode circuitry, rather than MOSFET circuitry. Plus, microwave bipolar/diode parts are easier to find these days than MOSFETs.

For the ultimate in EMI resistance, circuits should be built over a copper ground plane rather than on a printed circuit board. (No circuit board patterns are available for the HF/LF/audio circuitry described here for that reason—build them "dead bug" style.) Shielding is obtained by soldering together pieces of unetched double-sided fiberglass circuit board to form an open box, then adding a cover made out of 24-mil aluminum sheet. You can pack more into the box by building circuitry on both the top and bottom of a piece of copper-clad fiberglass and forming a box with top and bottom openings to gain access to the circuitry.

A Better Receive Mixer

The weak point of many Gunn transceivers is the receive mixer. The M/A-COM Gunnplexer with a circulator shown in Fig 1 works reasonably well, especially if the mixer injection

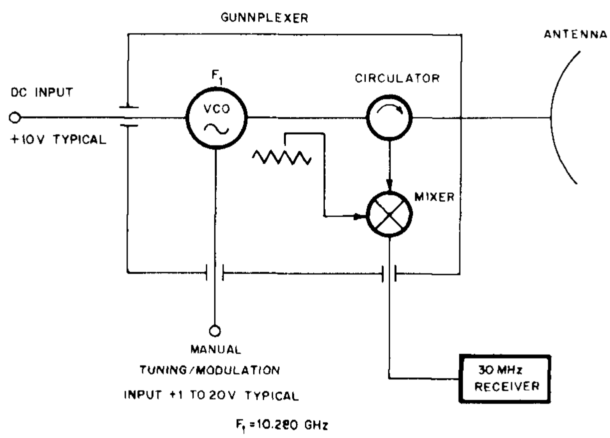

Fig 1—Schematic diagram of an M/A-COM Gunnplexer. This unit is optimized for communication work.

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is properly adjusted with a weak signal or a noise-figure (NF) meter for best performance. Weak-signal testing isn't too difficult with a WBFM radio since quieting improves considerably over a small range of input signal level. Circulator operation is load-dependent, so optimizing the injection level when you change antennas is a good idea. If you run high power, it may be possible to damage the receiver by upsetting the antenna impedance with a nearby object. The importance of a dc path for the mixer diodes cannot be stressed too highly. You can lose many dB of sensitivity if there isn't a good dc path through the diodes. For instance, a new Alpha source I used had a 10-kΩ mixer termination supplied. This is probably adequate over short paths, but adding a 680-Ω resistor in parallel dropped the noise figure from a lousy 22 dB to a reasonable 14 dB. 14 dB isn't too bad considering there is no circulator in this particular unit. (With no circulator, the designers of this unit had to place the mixer diode in the waveguide so that it wouldn't get too much LO signal. But this also causes it to receive less than the maximum received signal. With a circulator, the diode can be placed for optimum receipt of the input signal while the circulator ensures the proper LO level.)

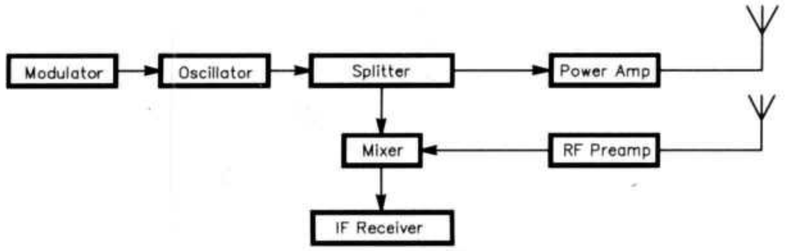

Fig 2—Block diagram of a high-performance Gunn transceiver.

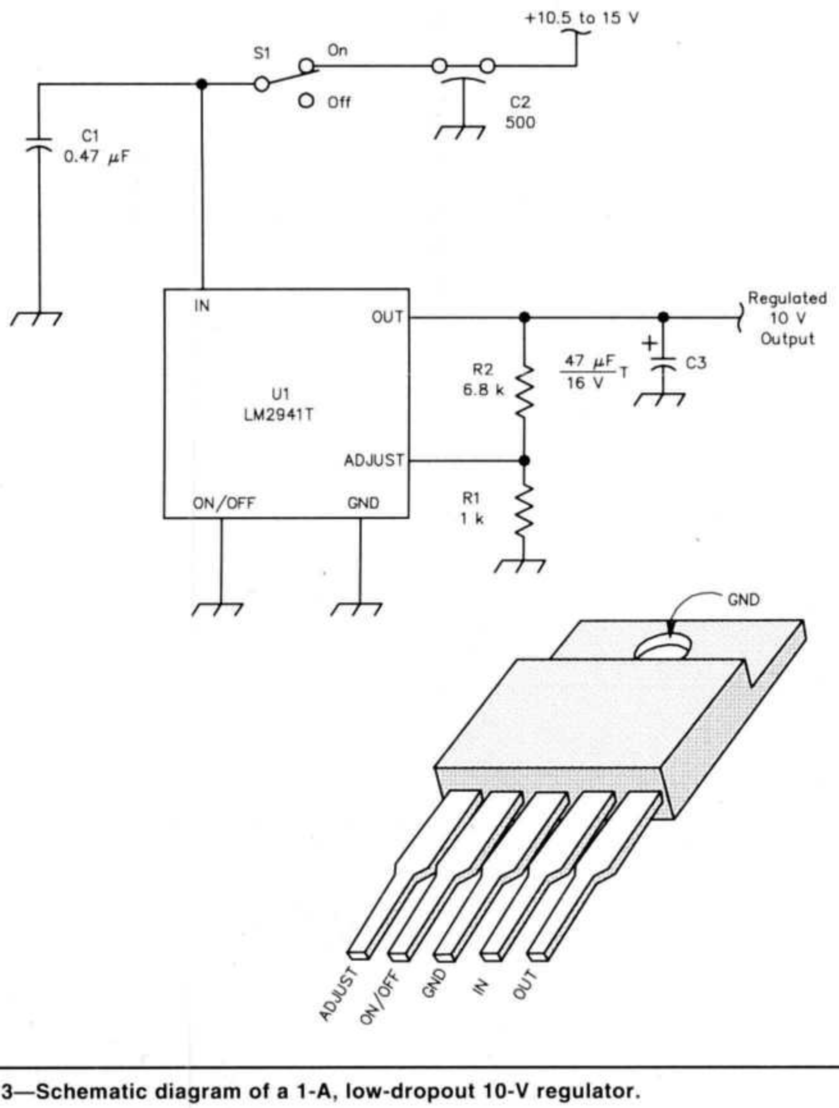

Fig 3—Schematic diagram of a 1-A, low-dropout 10-V regulator.

C1—0.47-μF non-polarized capacitor.
C3—47-μF, 16-V tantalum capacitor. This is absolutely essential for the regulator to work properly. The LM2941T data sheet says that 22-μF is the minimum value, but a much larger value can be used.
R1 and R2 determine the output voltage, which is approximately 1.28 (R1+R2) /R1.

If you have a source with no mixer, I suggest making an etched 3/2-λ hybrid ring mixer like the one used in my 10-GHz narrowband transceiver, published in *QST*.[1] While the existing splitter design will cost you 3 dB in transmitter power, the increased mixer sensitivity will make up for it if both ends of the link are identical.

For better than M/A-COM performance, you can add the amplifiers described in the article, as shown in Fig 2. This will improve performance by 10 to 13 dB over typical Gunn transceivers designed for communication work, and even more over those designed for automatic door openers. For really high performance, you might add a surplus high-power traveling-wave-tube amplifier (TWTA) or solid-state amplifier.

An interesting possibility is to design a high-performance setup for both narrowband and wideband operation. A minor mechanical complication results from the fact that narrowband operation is done using horizontal polarization while WBFM is vertically polarized. But, if you can do it, having the ability to work horizontally polarized WBFM is useful, since you often run across beginners setting up their WBFM systems with nonstandard polarization.

Antennas for Full Duplex

I decided it was best to use two

[1]Notes appear on page 31.

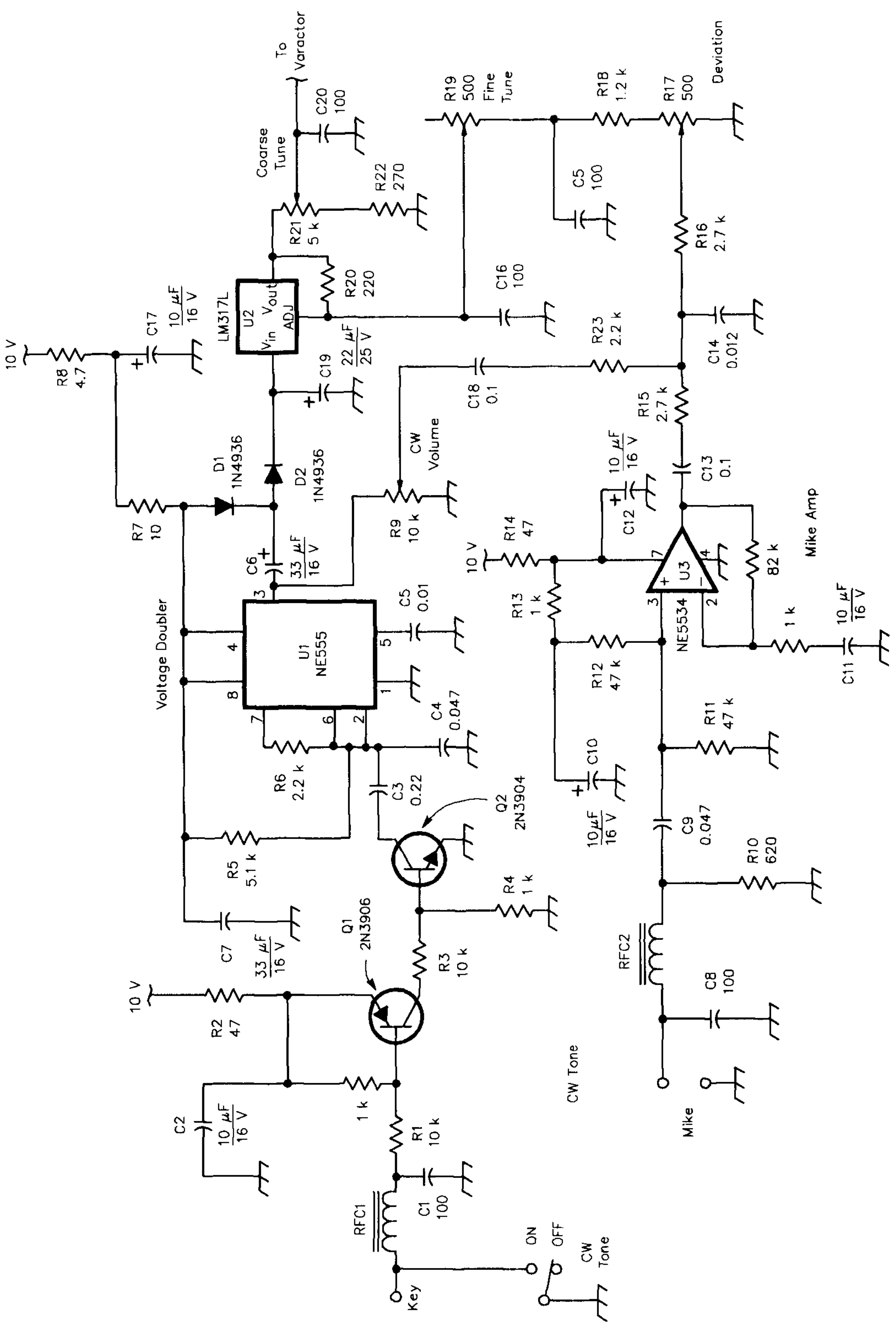

Fig 4—Schematic diagram of the varactor voltage generator circuit.

C1, C8, C15, C16, C20—100-pF bypass capacitors. These make the circuit less sensitive to VHF electromagnetic fields.
D1, D2—1N4936 1-A fast-recovery rectifiers.
Q1—2N3906 general purpose PNP switching transistor.
Q2—2N3904 general purpose NPN switching transistor.
R9—10-kΩ, linear-taper trimmer potentiometer.
R17, R19—500-Ω, linear-taper panel-mount potentiometers.
R21—5-kΩ, linear-taper panel-mount potentiometer.
RFC1, RFC2—5 turns # 28 enam. wire wound through an FB-101-43 ferrite bead.
U1—NE555 timer IC.
U2—LM317L adjustable regulator IC. The LM317T is easier to find but larger.
U3—NE5534 op amp IC.

under-illuminated dish antennas to preserve the full-duplex nature of the setup. It's tough enough setting up contacts without having to explain to the operator on the other end how to work your half-duplex station with his full-duplex equipment! I considered using a circulator to feed a single antenna but decided that adjusting the circulator terminations for proper isolation between the transmitter and receiver wasn't worth the trouble. Under-illuminating a pair of 16-inch, 0.25-f/D dishes with scalar feeds made a lot of sense—not only is it very difficult to fully illuminate such dishes, but too much gain can make it difficult to find stations. A good discussion of dish antenna feeds has been presented by Paul Wade, N1BWT.[2]

The use of separate dish antennas makes it easy to add a preamplifier. I used the 2.4-dB NF circuit I published in the 1993 *QST* article. (See Note 1.) You want to use a fairly rugged device with a reasonable dynamic range since coupling between antennas can result in rather large signals. You aren't going to hear much if your preamplifier saturates. Thus, it is also important to keep the gain of the preamplifier at reasonable levels unless you have a lot of attenuation between antennas. A different setup I tried, using a pair of 19-dBi horn antennas so it could be carried up stairs, didn't have enough isolation to allow preamps to be cascaded. But the under-illuminated dishes allow use of a pair of cascaded 17-dB-gain preamps.

Generating the Varactor Voltage

To get the most tuning range from a varactor diode, you want the tuning voltage to vary between 1 and 20 V. Sometimes you can cheat a bit by running it down to 0.7 or 0.6 V, though you really don't want to do this, since the Q of the diode degrades as the voltage drops. Too low a Q will result in excessive phase noise, which degrades the receiver performance, particularly if you are using a low IF; the noise will end up in the receiver passband. On the other hand, if you are using a voltage regulator you may have only 10 V to begin with. My solution is to use an LM2941T adjustable voltage regulator set to 10 V, as shown in Fig 3. This regulator can delivér 1 A, even with the input voltage down to 10.5 V. This device also offers reverse polarity protection, so extra protective diodes aren't needed. (Of course, C1, the input bypass capacitor, has to be nonpolarized or it may blow up if you hook the power up backwards.) To get a bit more voltage, I used an NE555 oscillator driving a diode doubler, shown in Fig 4. With this circuit, you can use a standard three-terminal voltage regulator chip, such as the LM317L, and still have around 14 V for the varactor, even at a battery voltage of 10.5 V. If you need the full 20 V, you might consider wiring up a diode tripler. I had plenty of tuning range so I didn't bother.

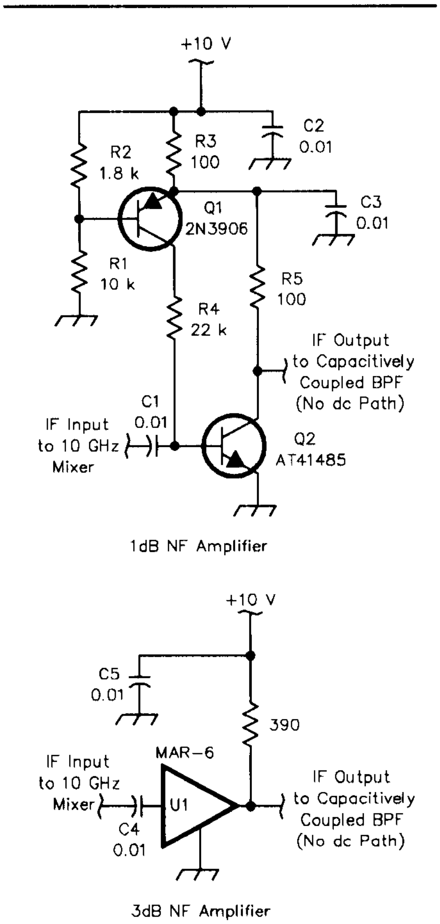

Fig 5—Schematic diagrams of two suggested microwave post-mixer amplifier circuits.
C1, C4—These capacitors determine the bandwidth of the amplifier. See text.
C7, C11—9 to 50-pF ceramic trimmer capacitors (Mouser 24AA084 or equivalent).
Q2—AT41485 low-noise bipolar transistor. Do not substitute.
U1—MAR-6, MSA-0685 MMIC.

The voltage from the voltage doubler is applied to an LM317L voltage regulator, which is used to produce an adjustable varactor voltage. The modulating audio signal is summed with the tuning voltage by applying the audio to vary the regulator's reference voltage. A low-dropout regulator such as the LM2941T isn't suitable here because the large capacitance needed at the output of a low-dropout regulator makes it difficult to modulate at high frequencies. And you can't eliminate this large capacitor because these regulators aren't stable without the proper load.

A possible problem with an oscillator/voltage doubler circuit is that the oscillator signal may appear on the tuning voltage and modulate the transmitter. I decided to use this as a feature. In setting up WBFM links, you typically send a CW tone that you listen for on your receiver. Normally, the NE555 oscillator runs at about 50 kHz, which is easily filtered out from the tuning voltage, and any small amount that isn't filtered out is too high in frequency to hear anyway. To send CW, the oscillator is slowed to an audio frequency. (I normally bring a little memory keyer, so I provided a ⅛-inch jack for keying. I also installed a little push-button on the side of the box, in case I forget the keyer.)

Using an LM317L voltage regulator for tuning isn't entirely necessary. What it allows is the practical use of two potentiometers instead of a single multiturn potentiometer. When potentiometers are cascaded, the fine-tuning adjustment is often useless over part of the range. But with the regulator circuit, the fine-tuning potentiometer has a useful tuning rate no matter where the coarse-tuning potentiometer is set. You could just use a 10-turn potentiometer instead, since the LM2941T provides a regulated voltage, but I find it more useful to have both fine- and coarse-tuning controls. You do have to be a little careful in using cascaded voltage regulators; the LM317L won't work properly if you don't provide enough overhead voltage. It needs an input voltage that is 1.3 to 1.9 V greater than its output voltage. Failure to provide sufficient overhead can seriously distort the modulation of the varactor voltage.

Note that the amplitude of the modulating voltage decreases as the coarse-tuning control is set for lower tuning voltages. This solves the nonlinear modulation sensitivity problem I mentioned at the beginning.

The audio circuit of Fig 4 works, although I haven't gotten around to optimizing it. A difficulty is that its response should match those of existing receivers. These vary from cheap converted broadcast FM receivers to expensive ARR units, so it isn't clear what response is needed for best weak-signal performance. I do know that my voice isn't exactly typical, so perhaps someone else is better suited to optimize the circuit. A dual op amp, such as the NE5532, would allow more flexibility than the single NE5534 used here. A better circuit would probably incorporate amplitude limiting to prevent over-deviation—I've run into cases where the other operator got so excited that I couldn't copy anything!

Post-Mixer Amplifiers

Two possible post-mixer amplifiers are shown in Fig 5. In the first of these, an AT41485 bipolar transistor offers about a 1-dB NF over a very wide bandwidth. The gain of this circuit is a bit excessive: over 25 dB at HF and still above 20 dB at UHF. Al Ward published a version of this (without active biasing) in the 1989 *Microwave Update*.[3] This circuit may oscillate with some terminations, as it is only conditionally stable. You could add negative feedback to fix the problem, but the noise figure will degrade. An MAR-6 MMIC is a bit more stable, though it has an NF close to 3 dB with a gain of 20 dB. If you are using a GaAs FET preamp ahead of the mixer, the MAR-6 makes more sense. Both circuits are usable from 30 MHz to 1 GHz unless you have parallel resonance problems with coupling capacitors. (A parallel tuned trap can seriously degrade the noise figure of an amplifier.) You might use smaller or larger capacitors than shown to better cover a particular range. If the capacitor value is too small, the reactance at low frequencies will degrade performance. If the capacitor value is too high, parallel resonances will degrade high-frequency performance. I used leaded chip capacitors that have very high self-resonant frequencies.

HF Down Converter

The circuit shown in Fig 6 is a 30 MHz to 10.7 MHz down converter

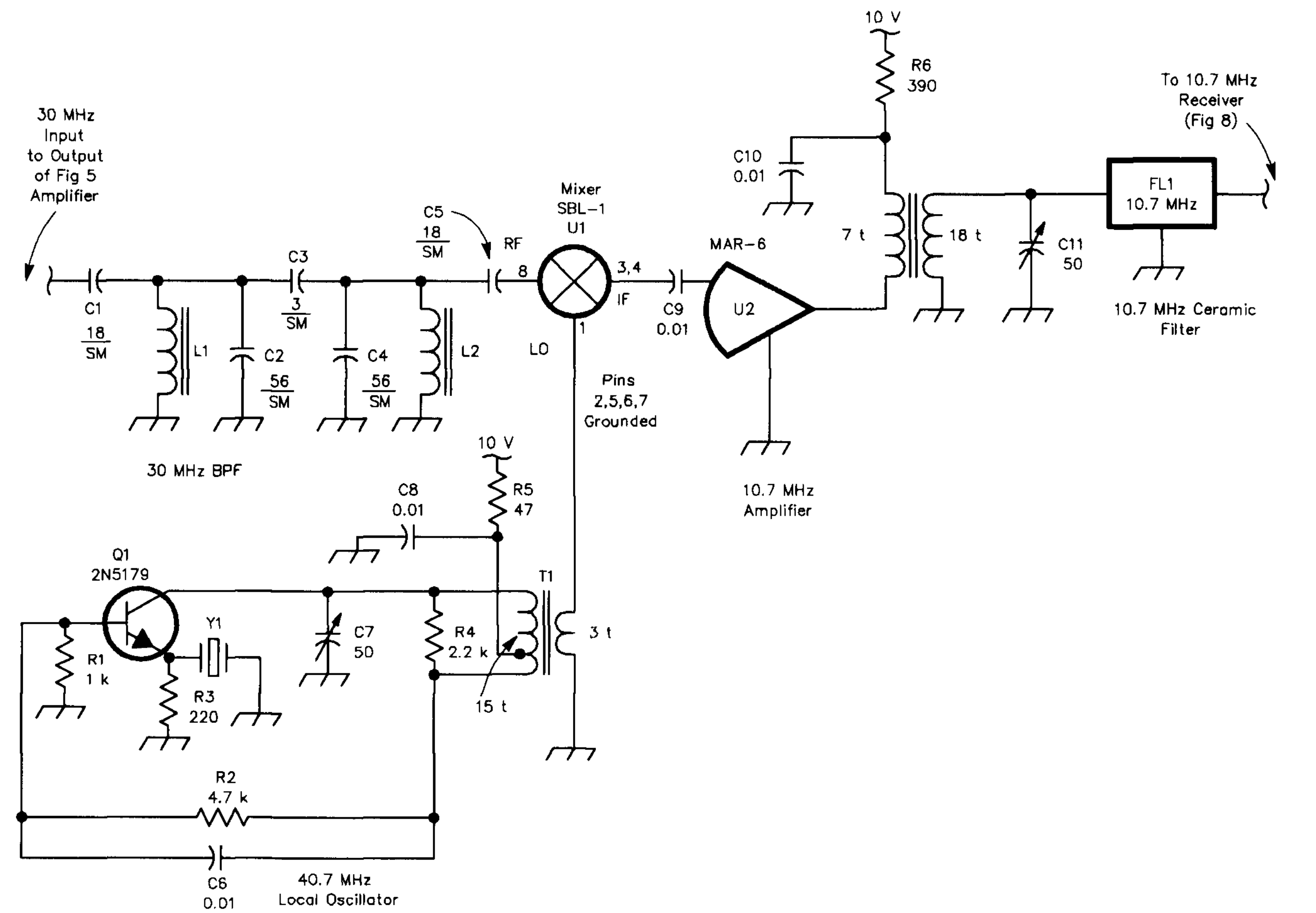

Fig 6—Schematic diagram of a down converter from 30 MHz to 10.7 MHz.

FL-1 —10.7-MHz center frequency, 230-kHz bandwidth ceramic IF filter (Digi-Key TK2306).
L1, L2—10 turns # 20 enam wire wound on a T-50-10 powdered-iron toroid core.
T1—Toroidal transformer. Primary: 15 turns # 28 enam wire wound on a T-37-10 ferrite toroid core. Tap 3 turns from R2. Secondary: 3 turns # 28 enam wire.
T2—Toroidal transformer. Primary: 7 turns # 28 enam wire wound on a T-44-2 toroid. Secondary: 18 turns # 28 enam wire.
U1—SBL-1 Mini-Circuits diode mixer.
U2—MAR-6, MSA-0685 MMIC.
Y1—40.7-MHz third-overtone crystal (International Crystal 471190). The case size isn't critical, but I prefer the miniature FM-5 case style.

and narrow IF filter. 30 MHz is pretty much the standard IF for Gunnplexer work, though big-gun contest station W2SZ/1 and their rovers use 10.7 MHz to minimize complexity. 33 MHz is used by some surplus distance-measuring equipment found in Canada.

The band-pass filter at the mixer input in Fig 6 is 2.1 MHz wide with 2.3 dB of insertion loss in a 50-Ω system. With a sweep setup to tune this filter, it is entirely practical to use fixed-value capacitors and just move the turns on the toroids around for a nice passband. Or, you might use 39-pF fixed capacitors and trimmer capacitors to make tuning easier. The band-pass filter is needed to eliminate the mixer image; otherwise the noise figure will be degraded by 3 dB. The mixer used is a Mini-Circuits SBL-1, a low-impedance mixer with excellent dynamic range. The 10.7-MHz ceramic IF filter that follows the mixer sets the system bandwidth. You don't actually need this filter for short-range contacts if you can accept roughly 10 dB of degradation in noise figure. The wider bandwidth might actually *improve* system performance if you are attempting to use either fixed-tuned units or 24-GHz Gunn oscillators, which tend to drift quite a bit more than 10-GHz units. A narrow bandwidth is pretty useless if you can't keep the signal in the passband long enough to make a contact.

The local oscillator, Q1, is a Hartley crystal oscillator with a tuned tank circuit. If needed, a capacitor or inductor can be added in series with the crystal to get the oscillator exactly on frequency. A receiver that is slightly off frequency will degrade the range of the Gunnplexer setup.

If you are really serious about performance, you can gain additional receive sensitivity by reducing the image noise. With a 30-MHz or lower IF, this is quite difficult to do with a front-end band-pass filter; you have to separate two 10-GHz signals that are 60 MHz different in frequency. This can be done with an image-reject mixer, however, as shown in Fig 7. This technique uses phasing to eliminate the unwanted image noise. Whether or not this is worthwhile depends on the amount of noise generated externally compared to that generated internally. As Rick Campbell has pointed out, there is little sense in using a filter to eliminate image noise if you have a noisy receiver in the first place.[4] In fact, it is possible for the filter to actually hurt, rather than improve sensitivity. On the other hand, a sensitive receiver can benefit from such a filter.

With a 100-MHz or higher IF, filtering to reduce image noise becomes practical, at least for fixed frequencies. The lowest-loss 10-GHz filters are those made out of waveguide, though cheap pipe-cap filters offer acceptable performance. The British seem to prefer filters made with irises, though I prefer filters made with posts; I find them easier to construct. But you can make iris filters from waveguide scraps. The little odds and ends can be filed or sanded down to exactly the right size, then soldered to sheet metal with the appropriately sized iris holes. A suitable filter using posts was published in *QEX* by Glenn Elmore, N6GN.[5] Glenn even includes the dimensions for WR-90-to-SMA transitions. But I recommend use of two- or four-hole flange SMA connectors instead of threaded connectors with lock nuts. The waveguide is tapped to accept 2-56 screws, so the SMA connectors can be easily replaced if damaged. More important, the connectors don't work loose from the waveguide as easily.

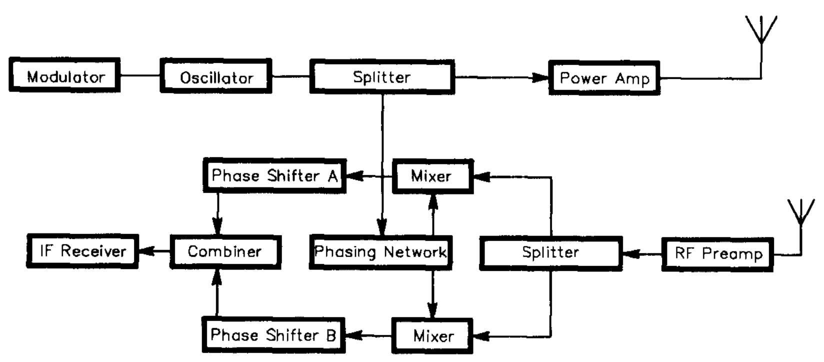

Fig 7—Block diagram of a sophisticated Gunn transceiver using an image-reject mixer.

Using Gunn Oscillators Made for Negative Supplies

I've seen surplus Gunn oscillators designed for negative-polarity supplies at hamfests. Unless you have a high-power diode that needs good heat sinking, these work just fine with positive supplies if you reverse the Gunn and varactor diodes. These units can be identified by noting the polarity of the electrolytic bypass capacitor. I made a simple spanner wrench out of a piece of brass rectangular stock and 1/16-inch stainless steel rod—this allows me to unscrew the plastic nuts found on M/A-COM units.

FM Receiver Without Squelch or AFC

For serious contest work, I don't find either squelch or AFC useful. They are actually a hindrance, since these features add additional controls that can be set incorrectly. You don't want to miss a contact because your squelch was set too high! Similarly, it is easy to set the AFC to the wrong polarity. AFC has a polarity because normal Gunnplexers can receive signals offset either above or below the transmit frequency. But the biggest problem with AFC is the difficulty of making long-distance contacts with stations whose IF is slightly off frequency. This occurs when L-C oscillators are used at a temperature considerably different from that for which the radio was designed. This isn't unusual in northern New England, where the weather can be warm and sunny one day and snowing the next. I've found that even inexpensive ceramic filters will vary a bit: it isn't unusual to find some of these to be 10 or 20 kHz off frequency. It might not be a bad idea to sweep them if you have a suitable measurement setup. Alternately, if you are using an L-C local oscillator, you might vary the frequency a little and see if a slightly different frequency improves the sensitivity.

Fig 8 shows a 10.7-MHz wideband FM receiver that is basically a copy of the design developed by Jay Rusgrove that has appeared in *The ARRL Handbook*. You can vary R1 to match different IF-filter resistances. And you can omit R3 and R5 if you want to keep the parts count low. R4 is used to optimize the Q of the quadrature network. There is a trade-off here between distortion, audio

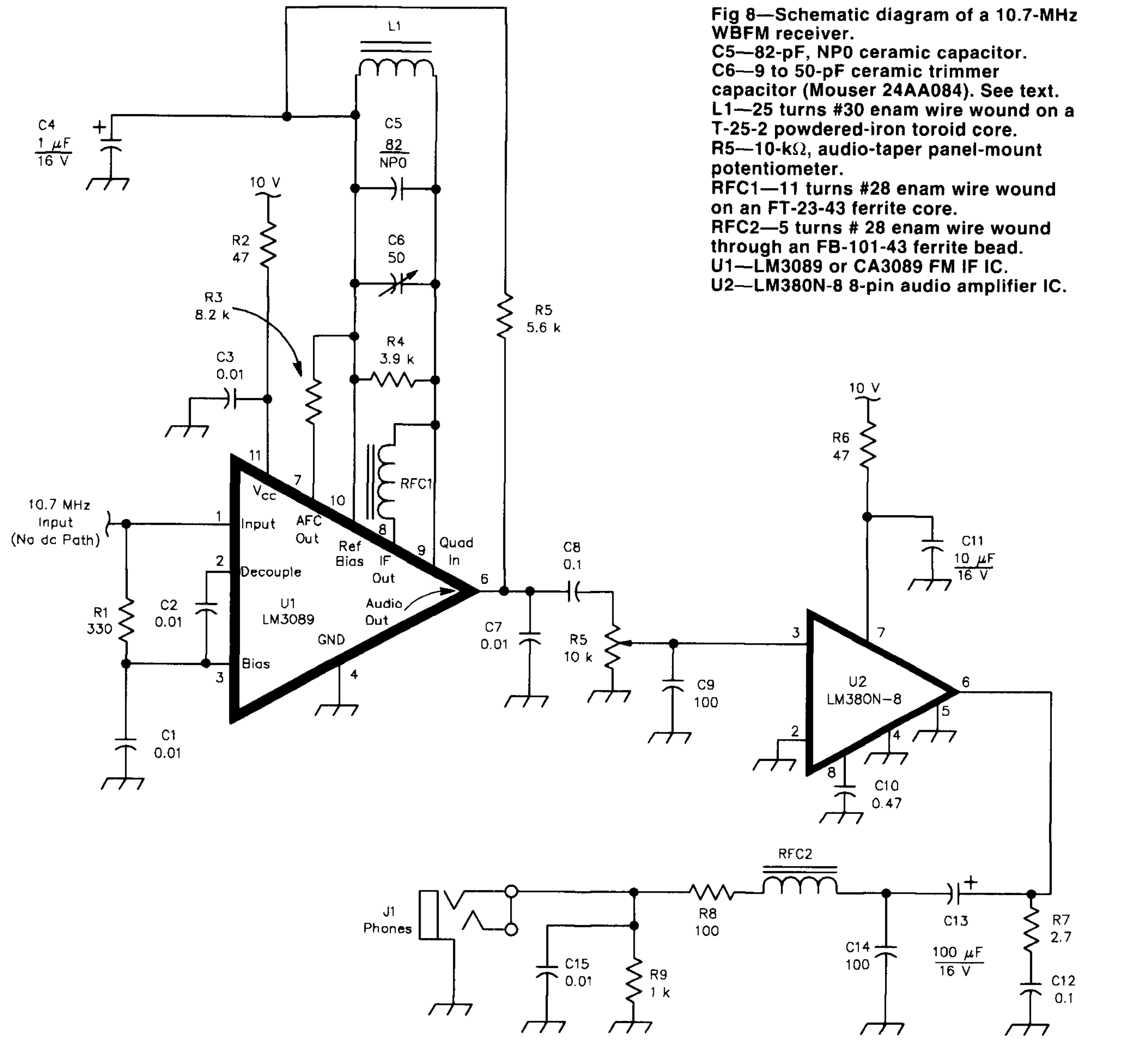

Fig 8—Schematic diagram of a 10.7-MHz WBFM receiver.
C5—82-pF, NPO ceramic capacitor.
C6—9 to 50-pF ceramic trimmer capacitor (Mouser 24AA084). See text.
L1—25 turns #30 enam wire wound on a T-25-2 powdered-iron toroid core.
R5—10-kΩ, audio-taper panel-mount potentiometer.
RFC1—11 turns #28 enam wire wound on an FT-23-43 ferrite core.
RFC2—5 turns # 28 enam wire wound through an FB-101-43 ferrite bead.
U1—LM3089 or CA3089 FM IF IC.
U2—LM380N-8 8-pin audio amplifier IC.

output, and maximum peak deviation. You can increase R4 to increase the Q and get more audio output, but this decreases the peak deviation and increases the distortion. Both film and ceramic trimmer capacitors will work for C6, but I prefer ceramic trimmers, which are less likely to be disfigured by an errant soldering iron.

Notes

[1]Lau, Zack, KH6CP, "Home-Brewing a 10-GHz SSB/CW Transverter," *Part 1: QST*, May, 1993, pp 21-28; *Part 2: QST*, June, 1993, pp 29-31.

[2]Wade, Paul, N1BWT, "Practical Microwave Antennas—Part 2," *QEX*, October, 1994, pp 13-22.

[3]Ward, A. J., "AT41486 Feedback Amplifier General Purpose Test Amplifier," *Proceedings of Microwave Update '89*, ARRL, p 197.

[4]Campbell, Richard L., "Low Noise Receiver Analysis," *Proceedings of Microwave Update '91*, ARRL, pp 1-13.

[5]Elmore, Glenn, N6GN, "A Simple and Effective Filter for the 10-GHz Band," *QEX*, July 1987, pp 3-5.