QEX

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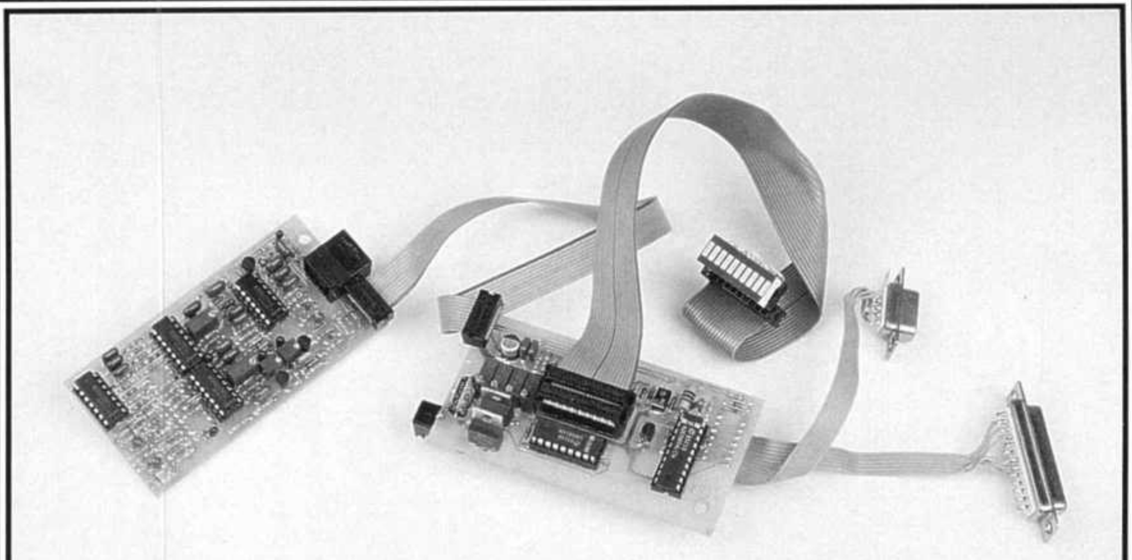
ARRL Experimenter's Exchange

May 1994



A Simple but Effective HF Modem

QEX: The ARRL
Experimenter's Exchange
American Radio Relay League
225 Main Street
Newington, CT USA 06111

QEX (ISSN: 0886-8093) is published monthly by the American Radio Relay League, Newington, CT USA.

Application to mail at second-class postage rates is pending at Hartford, CT.

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Subscription rate for 12 issues:

In the US: ARRL Member $12, nonmember $24;

US, Canada and Mexico by First Class Mail:
ARRL Member $25, nonmember $37;

Elsewhere by Airmail:
ARRL Member $48, nonmember $60.

QEX subscription orders, changes of address, and reports of missing or damaged copies may be marked: *QEX* Circulation. Postmaster: Form 3579 requested. Send change of address to: American Radio Relay League, 225 Main St, Newington, CT 06111-1494.

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About the Cover:
KC7WW's HF modem performs nicely with low-cost software to make a budget HF digital station.

ISSUE NO. 147

Features

Columns

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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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Empirically Speaking

HF Digital Protocols

As you'll see in one of this month's articles, there is a new entrant in the suite of HF digital data protocols available to amateurs: G-TOR. We now find almost a half-dozen different protocols in use on the HF bands for sending digital data. These protocols date variously from the early days of digital data transmission (RTTY), from a period when tubes and mechanical teleprinters were still the norm (AMTOR), from early attempts at error-free data transmission (packet), and modern protocols that make use of both the available technology and of what we have learned about HF digital data transmission over the years (Clover, Pactor and G-TOR).

With the increasing range of modern protocols available to us, it becomes harder and harder to justify the use of older protocols that are error-prone and/or inefficient. That's not to say, of course, that no one should *ever* use older protocols. There are valid reasons to use older protocols, including just because they are fun. This is *amateur* radio, after all. But when the object of the communication is to get a message from point A to point B, the best way to do so is with a means that is conserving of spectrum.

The measure of spectrum conservation that best applies here probably is bits/s/Hz. That is, a signal that packs the most bits into the narrowest bandwidth for the least amount of time will use less of that renewable but hard-to-get resource, spectrum.

So, what are the characteristics of a protocol that we should consider for serious HF data communication work? First, it should provide error-free data transfer. There is just no reason in this modern era to accept errors in the data. That pretty much eliminates RTTY and AMTOR from consideration, as both allow delivery of uncorrected errors. Second, some form of data compression is required. One can argue endlessly about *which* form of data compression to use, because the efficacy of data compression algorithms depends to some degree on the nature of the data being compressed. But clearly, almost any form of data compression is better than none at all. Third, the protocol should be able to use forward error correction (FEC) to minimize the effects of short-term variations in propagation and noise. Finally, some means of adapting to longer-term changes in the channel is needed. Putting all of those characteristics together in a protocol can produce an efficient system, if the details are well thought out and well implemented.

Of the protocols mentioned earlier, Clover, Pactor and G-TOR include these four characteristics, to varying degrees. These modern protocols are, therefore, choices that make some sense for use on HF. (One unfortunate characteristic of G-TOR, though, is its operation at 300 bauds. That makes its signal wider than 500 Hz, which is undesirable since many radios have filter choices of only 500 Hz and 2.1 kHz or more.)

So the message is, it's time to retire RTTY, AMTOR and packet as workhorse protocols for our HF digital communication. Use them for fun, if you want, but stick to the newer, better protocols when you have to get the bits moved.

This Month in *QEX*

Simple HF digital stations often use "software TNCs" that use the main computer to generate and decode the bit streams. Unfortunately, these inexpensive stations often also use poor modems. A remedy can be found in "AN-93: An HF Modem for RTTY, AMTOR and Pactor Software TNCs," by Johan Forrer, KC7WW.

Kantronics' new HF digital protocol is "G-TOR: A Hybrid ARQ Protocol for Narrow Bandwidth HF Data Communication," described by Glenn Prescott, WBØSKX, et al.

Microwave doesn't have to be expensive—or exhibit poor performance—as Robert Cook, N2SB, shows us in "5760 MHz from the Junkbox."

Software tools for design are becoming more available at prices hams can afford. One such tool, *SysCalc*, is reviewed this month by David Newkirk, WJ1Z.

Finally, Zack Lau, KH6CP/1, presents a 2-W 13-cm amplifier design in his "RF" column.—*KE3Z, email: jbloom@arrl.org.*

AN-93, an HF Modem for RTTY, AMTOR and Pactor Software TNCs

Low-cost HF digital operation need not suffer from poor modem performance—using this design.

By Johan B. Forrer, KC7WW

This article describes the theory, operation and construction of a low-cost modem for use with software terminal-node controllers (TNCs). This conservative design is based on tried and tested principles that ensure good performance on RTTY, AMTOR and Pactor even under unfavorable band conditions.

Overview of the AN-93

A software approach to HF digital communication, especially RTTY, AMTOR and Pactor, offers several cost-effective advantages to amateurs interested in the digital modes. Such an approach requires only three components, a PC-compatible computer, an HF modulator/demodulator (modem) and software that performs the encoding and decoding in real time.

Microcomputers have become both affordable and, besides radio equipment, an essential commodity in the ham shack. Combined with inexpensive software for PC compatibles, they can perform a function similar to that of a terminal node controller (TNC).

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AN-93 is a modest modem with good performance even under adverse conditions. The design evolved over a number of years and is based on various proven and tested principles dating back to the mid sixties.

Pactor uses a "soft" error-correction technique called memory ARQ that benefits from digitized estimation of received signal strength in addition to "hard-threshold" data. This modem provides both.

No HF modem is complete without some form of tuning indicator. A simple but effective bar-graph tuning display is included here. Buffered audio outputs are also provided for an oscilloscope-type display. Alternatively, the digitized signal strength may be used by computer software for a real-time on-screen display.

Modulation and Demodulation

Before proceeding with modem specifics, a brief introduction to HF digital issues may be worthwhile. This introduction will give a representative, though not comprehensive, treatment of some important considerations for HF digital communications, without any rigorous proof.

HF digital communication is designed to transfer data from one station to another. Whether the intended purpose is for keyboard-to-keyboard QSOs or to handle traffic, the principle remains the same. As shown in Fig 1, modern HF digital links consist of a pair of computers (a) to store/retrieve data and serve as the machine-user interface; modems (b) that form the bridge between the HF radio transceiver and the computer; HF radio transceivers, HF antennas and ancillary equipment, (c) and; an HF path to provide a means to maintain two-way communications (d).

RTTY, AMTOR and Pactor use frequency-shift-keying (FSK) modulation that is classified as an F1 emission.

In order to transmit text or character symbols, each symbol is represented in some code alphabet, such as

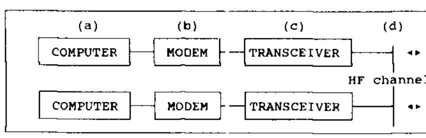


Fig 1—A two-way HF link.

ASCII or the CCITT ITA2, as a series of binary digits or bits. This binary system greatly simplifies RF modulation as there are only two states, logic high and logic low, corresponding to binary one and zero respectively.

These two logic states are commonly referred to in more generic terms as mark and space. This terminology, however, has to be interpreted in context—mark and space signaling may refer to +5 V and 0 V (unipolar keying), respectively, which are TTL levels, or −12 V and +12 V (bipolar keying), respectively, which are RS-232 levels. Mark and space may also refer to a pair of audio tones, such as 2125 Hz and 2295 Hz, or a pair of RF frequencies, such as 14.07500 MHz and 14.07483 MHz. The terms mark "channel" and space "channel" are often used in the context of frequency specification.

Data bits are eventually transmitted as one of two RF frequencies. For an F1 emission, only one state is transmitted at a time; it would not make sense to have both present simultaneously. By convention, the mark tone is transmitted as the higher RF frequency of the two states. Transmitting a space tone results in a downshift in RF frequency.

The frequency spacing between the mark and space channels is called the shift. For RTTY and AMTOR, a 170-Hz shift is commonly used. Pactor generally uses a somewhat wider shift, 200 Hz, for reasons that will become clear later.

A signaling element period is called a baud. For a binary system, one baud corresponds to a bit time. There are, however, other modulation schemes, such as Clover modulation, where each baud may be more than one symbol. This is called an *n*-ary code, where *n* denotes the number of possible signaling states, which is generally a power of 2. In this case the data rate in bits per second:

Data rate = signaling rate × $\log_2 n$ bit/s

The Nature of the HF Channel

HF modems are quite different from VHF/UHF or telephone modems, mainly because of the nature of the HF channel. McLarnon appropriately described propagation on the HF channel as containing elements of multipath propagation, multilayer propagation and noise.[1] These effects may cause selective fading of either the mark or space channel, or both. The resultant quality of the HF channel is thus variable: one moment the channel may be perfect, the next moment it may be seriously impaired. Depending on the type of propagation, a significant amount of phase shift may be introduced in the received signal. If phase-shift variations are of sufficient magnitude, bit timing will become erratic, resulting in sampling of adjacent bits, ie, intersymbol interference (ISI). It is intuitive that longer message packets have a greater probability of being corrupted than shorter ones.

Noise on the HF channel is also different from, for example, that on VHF/UHF channels. QRN, from atmospheric noise due to lightning, or QRM due to ignition noise, noise from power lines, electric motor hash and other man-made electrical interference, is commonly present.

Besides these confounding factors, the ever-increasing competition for radio frequency spectrum space and the resultant problems with congestion and adjacent channel interference has to be taken into account.

All these factors, either singly or in combination, may introduce errors in the received data.

A well-designed modem combined with an intelligent message delivery system can function under poor conditions, albeit at a slower data rate.

Practical observations over a number of years have shown that a good demodulator must have adequate dynamic range to cope with variable signal amplitude and good filters to reach into the noise, so to speak, to process weak signals.

Codes and Message Protocols

Besides the hardware, some form of data-link protocol must also be followed. The protocol is an important consideration because it affects the dynamics of some modem components such as time constants, AGC and adaptive thresholding.

RTTY uses asynchronous data transmission. No error detection or error correction of data is performed. Transmission speed is at fixed rates of 45.45, 50, 75, or 110 bit/s.

AMTOR uses two modes, Mode B, forward error correcting (FEC) and Mode A, which uses 21-bit data blocks with automatic error-request (ARQ) acknowledgment. Transmission speed is at a fixed rate of 100 bit/s. The code alphabet is based on a 7-bit, single-error-detection code. The block repeat rate is 450 ms. A shift of 170 Hz is commonly used. For a description of the AMTOR protocol, please see CCIR 476.[2]

Pactor uses data frames similar to AX.25. There are either 8 or 20 character symbols in a Pactor frame compared to three for AMTOR. Each frame contains a header, data and a block-check character (CRC). Two modes are used: a so-called "Unproto" mode that uses regular data frames without message acknowledgment, and an ARQ mode. A special type of error correction, called memory ARQ, is used. This allows the reconstruction of a good data block from several corrupted copies of the block.

Pactor uses either 8-bit ASCII coding or Huffman data compression. When using data compression, the number of bits per character becomes variable and thus allows the more frequently used characters to be coded more efficiently—with fewer bits for each character on average. Under favorable conditions, the effective character rate may thus be increased.

Pactor's transmission rate is adaptive and may operate at either 100 or 200 bit/s with 200-Hz shift. Some innovative schemes are implemented to balance channel space usage. One such scheme alternates the sense of logic high and low on each transmitted block. This way, bias in usage of

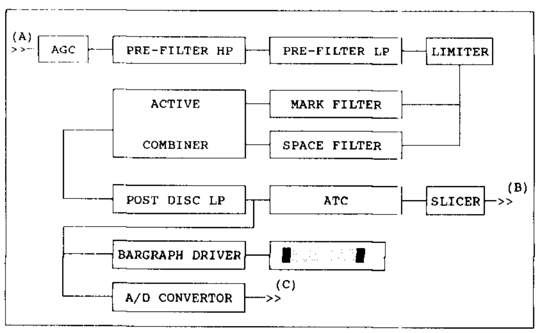


Fig 2—AN-93 modem block diagram.

[1]Notes appear on page 9.

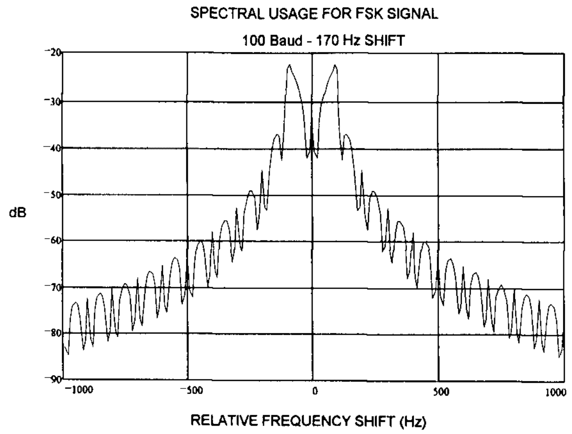


Fig 3—Spectral response of a 170-Hz shift, 100-bit/s FSK signal.

either above- or below-channel spectrum is minimized. A further measure that conserves channel bandwidth is the use of an idle data pattern of pairs of zeros and ones. This limits the occurrence of single signaling elements during idle periods—like when slow-typing operators are having a keyboard-to-keyboard QSO.

For an extensive discussion of the Pactor protocol, please see Helfert and Strate's definitive article.[3]

Data Rate and Bandwidth

One quickly learns from CW that keying waveforms with fast rising edges produce key clicks that use lots of bandwidth. The solution is found in shaping the waveform and thus effectively reducing bandwidth without loss of information. A similar situation exists for digital modulation.

If we consider an F1 emission—transmitting one frequency at a time—it appears as if we were keying the mark and space signals on and off. For example, if we were to send an alternating sequence of binary zeroes and ones at a rate of say 100 bit/s, the mark channel would be keyed on and off at a rate of 50 Hz. Similarly, the space channel will be keyed on and off at 50 Hz.

The resulting spectrum for 170-Hz shift, 100 bit/s is shown in Fig 3. The fundamental audio frequency, according to the *RTTY Handbook*, is shown as 100% with the first harmonic at 63% of the fundamental's amplitude, the third harmonic is 21%, and the fifth harmonic 12%.[4]

Note that in order to accurately preserve this on/off waveform (the fundamental and all of its harmonics), excessive bandwidth would be needed. Further, note the amount of overlap in the spectra of the mark and space channels. This may appear as too small a shift. This situation gets worse when the baud rate is increased. In such a case, a wider shift must be employed in order to recover more of the spectral components. There are, however, some limitations. The FCC regulations specify that the symbol rate on HF (below 28 MHz) is limited to 300 bauds or, alternatively, that the shift is limited to a maximum of 1 kHz.[5]

Limiting the bandwidth does not necessarily mean loss of information. It is of little consequence, for instance, to know that the original on/off modulation had very sharp transitions. All that needs to be determined is what the actual logic was and allow for some fair amount of tolerance.

Determining the needed bandwidth is quite simple. For the example shown, a 50-Hz bandwidth for each mark and space channel would recover the fundamental as well as the first harmonics of the 50-Hz signals. Taking into account that only part of the spectral content of the signal was recovered, the reconstructed waveform will appear with rounded edges. If the threshold applied to the signal is appropriate, this slight distortion will be eliminated and thus is of little consequence.

This leads to a formula for FSK often used by modem designers that defines occupied bandwidth (ie, spectrum usage at a specific signaling rate) as:

$$BW = B + DK$$

where

BW = necessary bandwidth
B = signaling rate (bauds)
D = frequency shift
K = shaping factor, typically 1.2

For an F1 emission, this corresponds to each channel filter having an approximate bandwidth equal to the signaling rate in bit/s.

Hoff suggested 65-90 Hz as "minimal bandwidth" channel filters for 45.45-bit/s, 170-Hz operation.[6] This amounts to approximately twice the signaling rate. Such a filter would include some tolerance to cope with frequency drift and slight off-frequency operation.

For those interested in a good discussion of wave-shaping and bandwidth reduction, see Miller as well as Bissel and Chapman.[7,8]

The Choice of Tone Frequencies

The choice of tone frequencies used by amateurs goes back to the era of mechanical teletype equipment. At one time, 85 Hz was a magic tone standard that was often used by communications engineers.

It is no accident that the mark frequency, 2125 Hz, is an exact multiple of 85 Hz (2125/85 = 25). Similarly, the space frequency, 2295 Hz (2295/85 = 27) is a multiple of 85 Hz. Note that the choice of these harmonics requires that they are odd and not harmonically related.

The tone pair 2125/2295 Hz also appeared to be ideally suited to the intermediate frequency filters in old radio receivers. One further reason for this choice was the bandwidth requirement for 45.45- or 50-bit/s RTTY, as discussed above.

Is this a good standard? European RTTY operators and designers did not think so. Their argument was that modern filters and audio passbands of receivers actually favor lower frequency tones. Thus the use of the tone pair 1275/1445 Hz that has become popular in Europe. Note that these tones also are multiples of the magic 85-Hz tone standard.

Which is better? There is a trade-off between radio receiver specifications—filter bandwidth and audio passband on the one hand and demodulator efficiency on the other. Since most demodulators use rectification of channel energy in the discriminator, a higher channel carrier frequency means more cycles to be rectified per signaling element than for a lower channel carrier frequency. However, this difference is rather marginal and could be seen as insignificant.

Modulation Schemes—FSK and AFSK

The two popular ways of generating

an F1 signal are direct frequency-shift keying (FSK) and audio frequency-shift keying (AFSK). Both will emit an F1 signal. Direct FSK generally uses some method to shift the frequency of an oscillator, usually at RF. With AFSK, audio tones (2125 Hz mark and 2295 Hz space) are generated and applied to the microphone input of an SSB transceiver. Note that in order to emit the space tone at a lower RF frequency, LSB mode must be selected for AFSK.

Most modern transceivers provide for direct FSK. This is the preferred method as generally such operation allows for special-purpose crystal filters to be used for reception.

AFSK is generally more complex and requires some careful thought in order to emit a clean signal without clicks. It is essential that phase-synchronous keying be used. This means that the AFSK signal generator should shift from one tone to the other while maintaining phase continuity. If the phase of the emitted tone is changed randomly, abrupt phase changes in the output will result. In the case of unshaped CW keying, excessive bandwidth is used. Further, be aware that the SSB speech modulator may involve a number of subcircuits—speech processors, balanced modulator, etc. In some instances there may be time delays associated with these that may play havoc with some link protocols.

Table 1—Prefilter Band-Pass Requirements for RTTY, AMTOR and Pactor.

	Signaling Rate bit/s	*Shift Hz*	*Occupied Bandwidth Hz*
RTTY	45.45	170	216
	50	170	220
	75	170	245
	110	170	280
AMTOR	100	170	270
PACTOR	100	200	300
	200	200	400

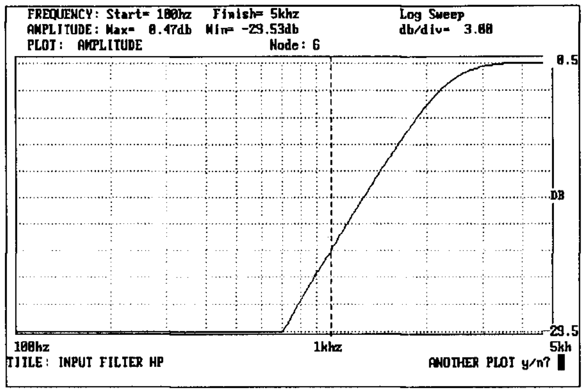


Fig 4—Input filter high-pass response.

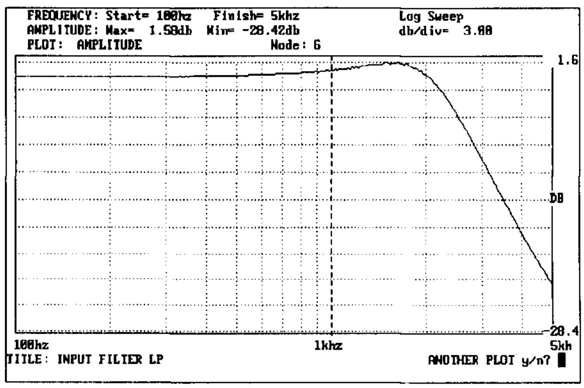


Fig 5—Input filter low-pass response.

AN-93 Architecture and Technical Description

Specifications for AN-93 are given below. For filter cutoff values, approximate values are provided for comparative purposes.

- AGC at > 100-mV audio input amplitude.
- Prefilter bandwidth > 1 kHz.
- Discriminator channel filter bandwidth 120 Hz.
- 170-Hz channel separation.
- Data low-pass filter cutoff 200 Hz.
- Automatic threshold corrector.
- Tuning indicator.
- A/D converter for "soft" error correction.

Analyzing the schematics of different types of modems makes the differences in HF modems quite apparent. Coping with the variability of HF signal characteristics and with different protocols presents a special challenge to modem designers.

A block diagram depicting AN-93's architecture is shown in Fig 2. The circuits for the modem and display electronics are shown on pages 8 and 9.

AGC

The first stage is an automatic gain control (AGC) using U1A and Q1. This stage copes with an extremely large dynamic input range. The attack/decay R-C constant is set at approximately 0.1 second in order that AGC action does not interfere with the rhythm of either AMTOR or Pactor operation. AGC action starts at roughly 100 mV. The merits of AGC have been debated in the past. However, without some form of signal level control, severe distortion can occur in the filter stages.

Prefilter Band-Pass

The prefilter band-pass is implemented with separate third-order high-pass, U1B, and low-pass, U1C, active filters. The required bandwidths of prefilter band-pass filters for RTTY, AMTOR and Pactor are summarized in Table 1.

The effective response of the HP/LP prefilter is somewhat wider than the required maximum. Theoretical response curves are shown in Figs 4 and 5. The prefilter response was measured by applying a calibrated RF signal to the author's transceiver and is shown in Fig 6. The measured –6-dB (half voltage point) bandwidth is in excess of 1 kHz. It was not possible to determine the complete prefilter bandwidth as the transceiver's IF filter attenuated signals above 2600 Hz.

This prefilter may be improved, as it is not possible to obtain good filter skirts with such low-order filters. Stark describes an alternative design for such an improved prefilter (see Note 1).

Limiter

The output from the prefilter is hard limited by U1D before being applied to the mark and space channel discriminator filters.

Channel Filters

A pair of biquad active filters, U2 and U3, are used in the discriminator. There obviously is a compromise for optimal bandwidth of the channel filters to accommodate a range of bit rates between 200 and 45.45 bit/s and 220-Hz and 170-Hz shifts.

The actual measured discriminator output response is shown in Fig 7. This response is somewhat similar to Hoff's famous 170-Hz filters. The occupied channel bandwidth is approximately 300 Hz. This bandwidth will reliably decode 130 bit/s at 170-Hz shift. Decoding 200-bit/s, 170-Hz Pactor will require careful tuning. For 200-Hz shift, the maximum data rate would correspond to 100 bit/s, making 200-bit/s, 200-Hz shift Pactor just possible, but requiring very critical tuning and strong signals.

Scope Output

Outputs from the channel filters are further buffered by U2D and U3D for oscilloscope tuning purposes.

Detection and Post-Detection Filter

Recovery of the modulation is though rectification and magnitude combining by U4A. The combined signal is further low-pass filtered to remove the carrier tones, thus producing

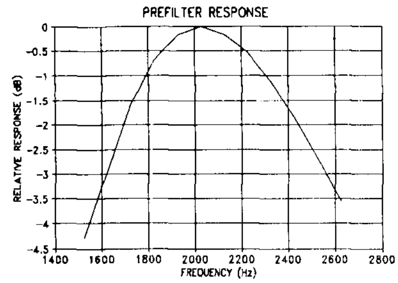


Fig 6—Measured prefilter response.

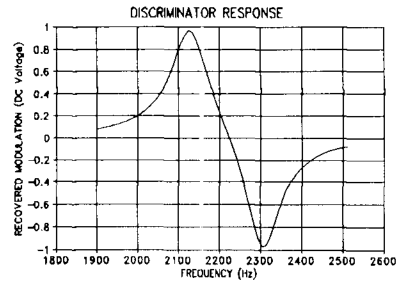


Fig 7—Measured discriminator response.

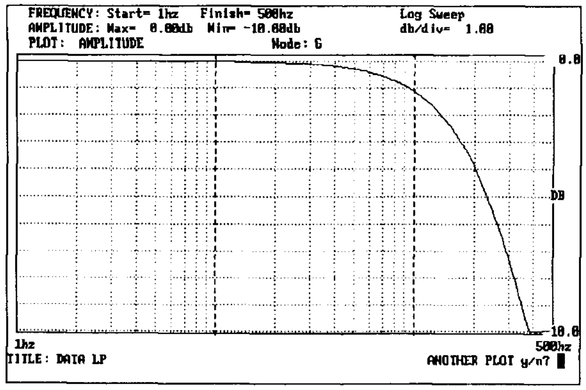


Fig 8—Data filter low-pass response.

the recovered modulation signal. The post discriminator low-pass filter, U4B, is a third-order Chebychev active filter with a cutoff of approximately 200 Hz. This allows high-speed Pactor demodulation. The theoretical filter response for the data low-pass filter is shown in Fig 8.

ATC and Slicer

To cope with signals that are slightly off frequency, an automatic threshold corrector (ATC) is used. This circuit will track asymmetry in the discriminator output signal caused by slightly off-frequency signals. Note that the ATC has a relatively long R-C constant in order not to interfere with the rhythm of AMTOR or Pactor.

The slicer stage, U4C, neatly squares and level-shifts the recovered signal to RS-232 levels.

Tuning Display and A/D Converter

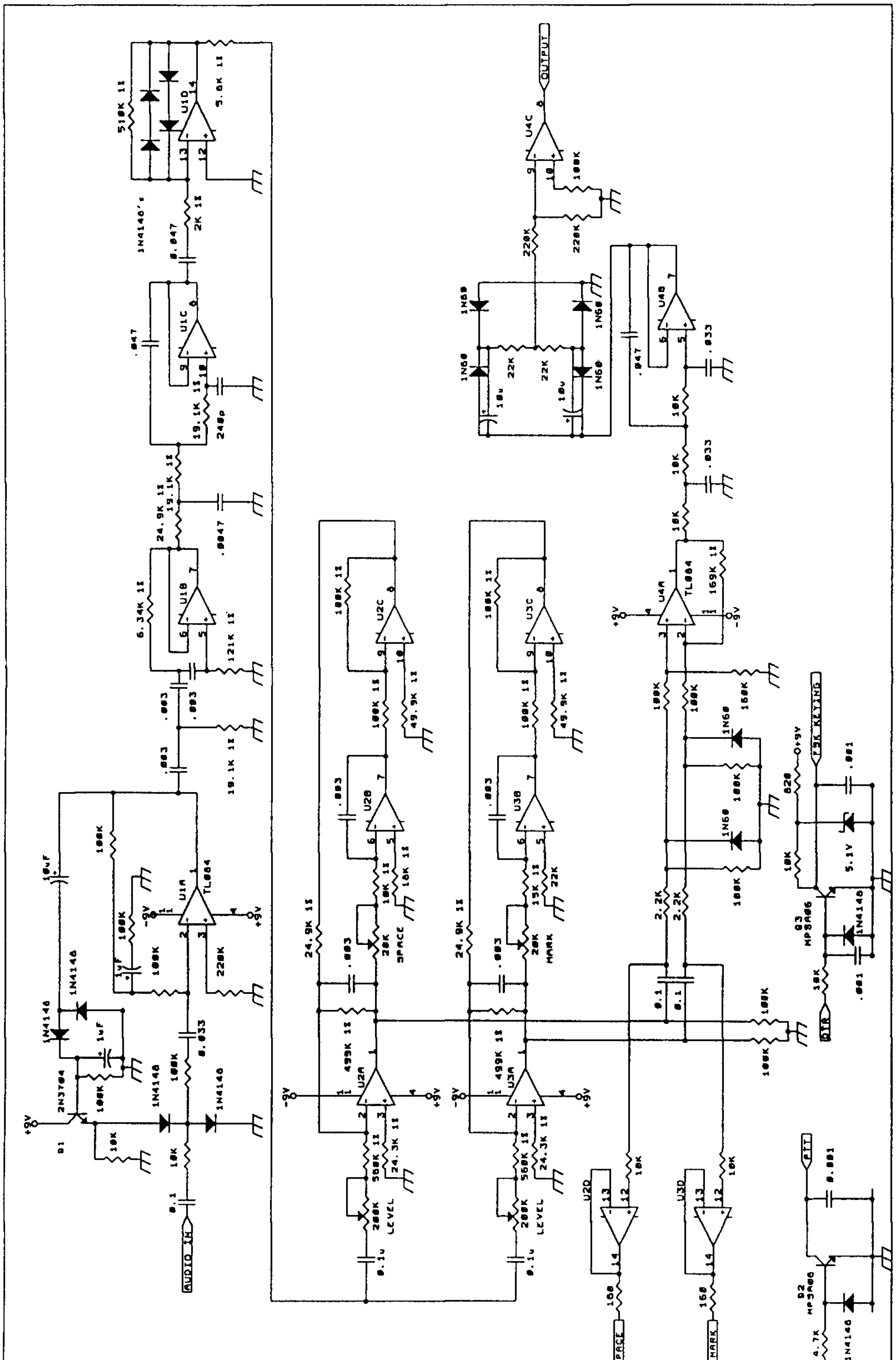
The recovered modulation signal varies between approximately 9 V and –9 V corresponding to mark and space, respectively. This signal is applied to a bar-graph display driver for tuning purposes. For a well-tuned signal, the frequency of occurrence of mark and space elements increases and thus appears as brightly modulated dots at the extreme ends of the display. Noise, on the other hand, switches much more rapidly than mark and space elements and has variable amplitude and thus appears as a smeared dot pattern around the display's center dots.

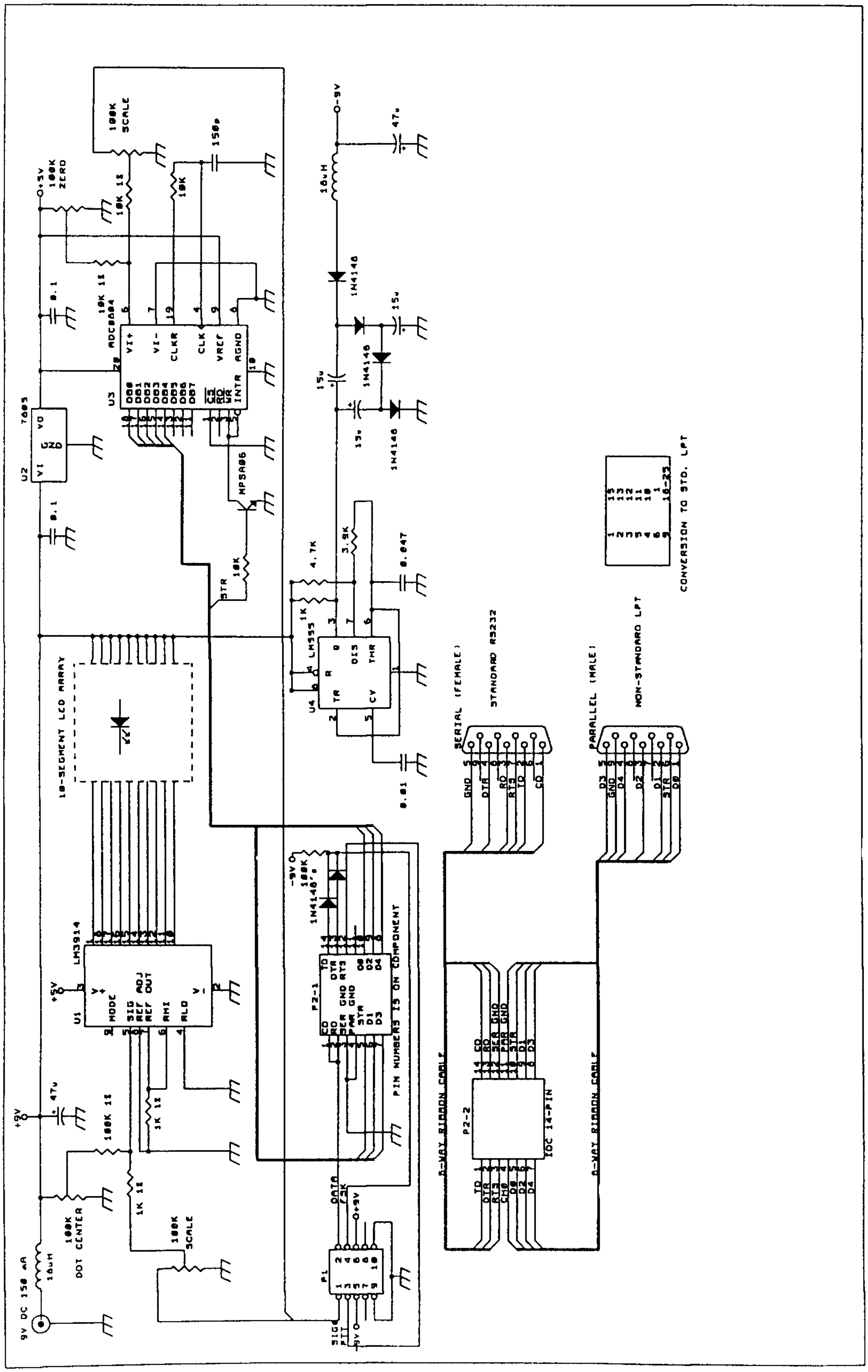
Computer Interface

The output of the slicer varies between 9 V and –9 V. These voltage levels are adequate for use with RS-232 circuits.

Project Construction and Alignment

Please refer to the board layout diagrams for assembly and parts placement. (Note: These diagrams are available from ARRL. Send an SASE to: Technical Department, ARRL, 225 Main St, Newington, CT 06111. Ask


AUDIO IN
OUTPUT
SCOPE SPACE
SCOPE MARK
FSK KEYING
PTT
RTS
DTR
SPACE
MARK
LEVEL
LEVEL


ADC0804
LM3914
LM555
MPSA06
7805
10-SEGMENT LCD ARRAY
SERIAL (FEMALE)
STANDARD RS232
PARALLEL (MALE)
NON-STANDARD LPT
CONVERSION TO STD. LPT
PIN NUMBERS IS ON COMPONENT
IDC 14-PIN
8-WAY RIBBON CABLE
9V DC 150 mA
DOT CENTER
100K SCALE
100K ZERO

Parts list

Demodulator board (AN93-D)

Qty	*Description*
Resistors 1% metal film ⅛ W	
1	2.0 kΩ
1	6.34 kΩ
1	5.6 kΩ
1	10.0 kΩ
1	15.0 kΩ
1	18.0 kΩ
3	19.1 kΩ
1	22.0 kΩ
2	24.3 kΩ
1	24.9 kΩ
2	49.9 kΩ
1	121.0 kΩ
2	499 kΩ
1	510 kΩ
Resistors 5% carbon film ⅛ W	
2	160 Ω
1	820 Ω
1	4.7 kΩ
9	10 kΩ
2	22 kΩ
12	100 kΩ
3	220 kΩ
Multiturn potentiometers	
2	20 kΩ
2	200 kΩ
Capacitors	
1	240-pF ceramic
3	0.001-µF ceramic
7	0.003-µF stacked metal film
1	0.0047-µF stacked metal film
3	0.033-µF stacked metal film
3	0.047-µF stacked metal film
5	0.1-µF ceramic
2	1.0-µF tantalum
3	10-µF tantalum
Semiconductors	
10	1N4148 silicon signal diodes
6	1N60 germanium signal diodes
1	5.1-V Zener diode, 400 mW
4	TL084LCN quad op amp
2	2N3904 NPN silicon switching transistors

Supply and display board (AN93-P)

Qty	*Description*
Resistors 1% metal film ⅛ W	
2	1.0 kΩ
1	100.0 kΩ
2	10.0 kΩ
Resistors 5% carbon film ⅛ W	
1	1 kΩ
1	3.9 kΩ
1	4.7 kΩ
2	10 kΩ
1	100 kΩ
Multiturn potentiometers	
4	100 kΩ
Capacitors	
1	150-pF ceramic
1	0.01-µF stacked metal film
1	0.047-µF stacked metal film
3	0.1-µF ceramic
2	15-µF electrolytic
2	47-µF electrolytic
Semiconductors	
1	LM555CN timer
1	LM3914N bar-graph display driver
1	ADC0804LCN Analog-to-Digital converter
6	1N4148 silicon signal diodes
1	LED bar graph
1	2N3904 NPN silicon switching transistor
1	7805 5-volt regulator
Connectors	
2	10-pin IDC ribbon connectors (AMP 746610-1)
1	14-pin IDC connector
1	dc socket
1	RJ-11 socket
Miscellaneous	
2	10-µH RF chokes

for the "AN-93, *QEX* 05/94" package.) I recommend that a low-wattage, temperature-controlled soldering iron be used. A small tip is required. Use only good-quality solder. Finally, do not work on the project when you are overstressed or tired. In the event of an error, it is quite difficult to remove these small components, and the result is most likely a ruined component or damage to the circuit board.

The AN-93 project consists of two printed-circuit boards. One board, AN93-D, contains all the components for the demodulator, and the other, AN93-P, contains the tuning display and the A/D convertor electronics. These two boards are interconnected through a 10-wire ribbon cable.

Connections to the RS-232 serial and parallel ports are made through a 14-pin connector located on the display board. Power is applied though a dc receptacle located on the display board. Interfacing to the radio is through a four-wire RJ-11 connector mounted on the demodulator board.

Assembly of the project starts by mounting the largest components first. I suggest you use IC sockets in case it becomes necessary to replace integrated circuits.

The smaller components require a great deal of care, especially in bending leads and soldering. All resistors must be shaped for 0.3-inch spacing. Most capacitors will readily fit in their holes, although further lead-shaping may be required. Pay careful attention when shaping the leads of glass diodes and some of the smaller capacitors as they are very fragile. It also is a good idea not to bend leads directly against a part's body. Excessive lead strain may cause a part to fail during soldering or, even worse, lead to unreliable operation.

As a guide, I suggest you populate the boards starting from one edge, one column at a time. Insert only a few parts at a time, soldering them into place, then crop leads. Double-check your work.

Once completed, with everything double-checked, interconnect the two boards and apply 9-V dc power. Check the operation of the negative voltage supply at P1 pin 5. It should be approximately –9 V. The positive supply will be approximately +9 V as measured at P1 pin 6.

Filter Alignment

To align the filters, a calibrated audio signal generator is needed. An audio tone and a frequency counter may be used instead. First, apply a 2125-Hz sine-wave signal with an amplitude of approximately 450 mV p-p to the modem's input. Adjust both level adjustment multiturn pots (200 kΩ) for maximum signal at pins 1 of U2A and U3A. If all is well, a clean unclipped signal should be present at U1A pin 1.

The mark filter is then peaked by adjusting the 20-kΩ multiturn pot adjacent to U3 until the output at U2A pin 1 peaks.

Similarly, peak the space filter's 20-kΩ multiturn pot while applying a 2295-Hz signal to the input.

The filter output level should next be balanced by adjusting the 200-kΩ pots so that the signal on U2A pin 1 with 2295 Hz is in the same amplitude as that seen on U3A pin 1 when 2125 Hz is applied.

Bar-Graph Display Setup

Adjustments are provided for full-scale (range) as well as zero (center). These adjustments are often interactive, and have to be repeated a few times to obtain the desired settings. The idea is that with no signal at the modem's input, the bar graph's middle two LEDs should be on. With a mark or space tone at the input, the LEDs at the corresponding extreme end should be on. It is often easiest to start by adjusting the center, then vary between the mark and space to adjust the range. Repeat the procedure a few times.

Due to some minor unsymmetry in the positive and negative 9-V supplies, this adjustment will hold only for a particular 9-V power supply. There will be slightly different settings for a different supply.

As a final check, the output of the modem should be checked at the RS-232 transmitted data output (P1 pin 2). This signal should be –9 V in the mark state and +9 V in the space state. This voltage convention is required for the serial ports of computers.

A/D Setup

The modem may be used without the A/D chip when memory ARQ is not required. However, for full-memory ARQ capability, as well as the on-screen tuning display of the *PCTOR* software, the A/D chip and parallel port (LPT 2) cable must be present and connected.

The A/D converter has adjustments for full-scale and zero settings. The actual A/D converter settings are set using a computer program that continuously displays A/D values. This program is named ADTEST.EXE and is supplied with the *PCTOR* documentation.

Run ADTEST. Note that a continuous display of A/D readings will be displayed. The first is a 2-digit hexadecimal number that represents the 2's complement 5-bit A/D value, the second is a converted decimal equivalent. This decimal number corresponds to a scaled voltage with a magnitude between approximately 14 at the mark tone and –14 at the space tone.

To adjust the A/D range and center settings, apply a quiet audio signal to the modem. As with the bar-graph display, several passes at these adjustments will be required, as there is some interaction between adjustments. It is convenient to first adjust the center setting by setting the audio input to approximately 2210 Hz. (This should correspond to the zero setting on the LED bar graph as well.) The A/D decimal value should alternate between zero and one. Then alternate between mark and space tones. Note that a numerical rollover occurs at mark—a voltage greater than 14—and will be displayed as a negative number. Similarly, a numerical rollover at space will be shown as a positive number.

The actual maximum and minimum decimal voltage values are not critical. If possible, adjust for a maximum value of 12 at mark, and –12 at space. It is important that no rollover occurs under maximum signal conditions.

Using AN-93 with *PCTOR* and *PC-PACTOR*

The model AN-93 modem was designed for use with *PCTOR* and *PC-PACTOR*. Shareware versions of these programs are available from several sources. The HamNet forum on CompuServe, TAPR, and the ARRL BBS are a few.

AN-93 allows the use of *PCTOR*'s on-screen tuning display and advanced error-correction capability.

[The author has several sets of AN-93 boards available for purchase. Contact him for details.]

Notes

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

[2]CCIR Recommendation 476-4 (1986), "Direct Printing Telegraph Equipment in the Maritime Mobile Service," *Recommendations and Reports of the CCIR*, 1986, Volume VIII-2, Maritime Mobile Service, Geneva; International Telecommunications Union, pp 60-69.

[3]Helfert, H. P., DL6MAA, and Strate, U., DF4KV, "PACTOR Funkfernschreiben mit Memory-ARQ und Datenkompression," *CQ-dl*, November 1990. (Also, English translation in October 1991 *QEX*, p 3.)

[4]Green, Wayne, W2NSD, *RTTY Handbook*, TAB Books, Blue Ridge Summit, PA 17214, 1972. ISBN: 0-8306-2597-6.

[5]*Code of Federal Regulations, CFR 47*, Part 80 to end. (1991) United States Government Printing Office, Washington DC, pp 583-586.

[6]Hoff, Irvin M., W6FFC (K8DKC), "High-Performance RTTY Filters, Improved Designs for Better Reception, Part 1," *QST*, August 1966. pp 16-19. Part 2, *QST*, September 1966, pp 34-37.

[7]Miller, James, G3RUH, "The Shape of Bits to Come," *Proceedings of the 10th ARRL Computer Networking Conference*, San Jose, CA, September 1991, pp 112-120.

[8]Bissel, Chris C. and Chapman, David A., *Digital Signal Transmission*, Cambridge University Press 1992.

[9]Stark, Peter A., K2OAW, "Design An Active RTTY Filter," *73 Magazine*, September 1977, pp 38-43.

G-TOR: A Hybrid ARQ Protocol for Narrow Bandwidth HF Data Communication

This new protocol supplies efficient, error-free HF communication while being simple enough to run on a general-purpose microprocessor.

by Glenn Prescott, WBØSKX
Phil Anderson, WØXI
Mike Huslig and
Karl Medcalf, WK5M

Introduction

On New Year's Day, 1994, WØXI and WK5M transmitted a series of 9718 byte files from Kansas to WA4EGT in California using a new digital mode. The exchanges took place on a very noisy and crowded 20-meter band, and the average transmission time for each exchange was 6 minutes and 30 seconds (23.7 characters/second average throughput). The channel was weak and the band overly crowded, as usual. However, the new mode, equipped with modern digital-communication signal-processing features was able to minimize retransmissions and maintain data throughput by using data compression, interleaved frames, forward error correction and fault tolerant ACKs. Throughout the month of January these tests were repeated with over one-million bytes transferred error-free. After comparing the performance of the new mode to Pactor and AMTOR, we knew we had a winner. The new mode is called G-TOR, short for Golay-TOR, an innovation of Kantronics, Inc. It's a new HF digital-communication mode for the amateur service, which was inspired by HF Automatic Link Establishment (ALE) concepts outlined in Federal Standard 1045 and in MIL-STD-188-141A.[1,2] Although it is not our intention to use the G-TOR protocol for ALE applications, it is structured in such a way as to be compatible with ALE systems when they become available; and the fundamental word-structure and error-coding techniques are similar.

Our original intention was to develop a new HF protocol that would serve as an evolutionary step-up from AMTOR and Pactor. We wanted a protocol that used more advanced signal-processing techniques than are available with current modes, yet we wanted it to be simple and inexpensive to implement on existing multimode TNC hardware. The key features of G-TOR set it apart from existing TNC HF modes:

- Golay forward-error-correction coding
- Full-frame data interleaving

[1]Notes appear on page 19.

Kantronics and RF Concepts, Inc
1202 E 23rd St
Lawrence, Kansas 66046-5006
Email: 73744,457 (CompuServe)

- Huffman data compression—two types, on demand
- Embedded run-length data compression
- CRC error detection with hybrid ARQ
- Link-quality based signaling rate: 300, 200, or 100 bauds
- Hybrid, 2.4-second ARQ cycle
- Fuzzy acknowledgments
- Reduced overhead within data frames
- Standard FSK tone pairs (mark and space)

The primary benefit of these innovations is increased throughput—ie, more bits communicated in a shorter amount of time. This is achieved because the advanced processing features of G-TOR provide increased resistance to interference and noise and greatly reduce multipath-induced data errors. The Amateur Radio community might question the need for another HF data communication protocol, especially since the relatively recent introductionof Pactor andClover. Before describing the G-TOR features in detail, we will first address this issue.

The Need and Motivation for G-TOR

Over the past several years many excellent articles and editorials have appeared in *QEX* and other ham radio journals discussing the deficiencies of current HF data communications protocols. For examples, see Notes 3 and 4. The central theme of these articles is that current protocols fall short of their potential because they fail to make use of modern signal-processing techniques such as forward error correction, data compression, and interleaving; and they are not able to adapt to the changing propagation conditions of the HF bands. AMTOR is an example of this type of system. AMTOR is effective primarily because of its very short transmission cycle, but it relies completely upon an automatic repeat request (ARQ) protocol. Pactor, on the other hand, extends the transmission cycle time considerably with the help of a technique called *memory ARQ*, which is a form of forward error correction that employs soft decision, majority logic decoding. Additionally, Pactor adapts simply to channel conditions by automatically selecting 200- or 100-baud transmission rates based on good/bad frame counts.

Clover, the new HF data communication system marketed by HAL Communications, Inc, is a system that makes use of these advanced features and can do so adaptively, responding to changing channel conditions. However, the Clover system requires sophisticated and expensive digital signal-processing hardware, and is generally too complex to be implemented on existing TNCs. One of our primary goals in developing the G-TOR protocol was to provide a system which can take advantage of several advanced digital communications techniques for maximizing throughput on the HF bands. However, an equally important goal was to keep this protocol straightforward and relatively easy to implement on existing multimode TNCs.

The purpose of the G-TOR protocol is to provide an improved digital radiocommunication capability for the HF bands, and the propagation characteristics of this medium and the operating procedures employed in the Amateur Radio service place some challenging constraints on HF protocol design. The HF bands have largely been dominated by voice (AM and SSB) communication for many years. The human voice has two notable features that make it suitable for HF. First, the bandwidth of voice is relatively narrow (less than 3 kHz) and the ultimatereceiver for voice transmissions (ie, the human brain) is an extremely effective system for copying the voice signal in the presence of noise, interference and fading. On the other hand, for data communication, the miserable propagation conditions that are characteristic of the HF band make effective communication a nightmare. Received signals, weakened over long distances, are often subject to multipath fading caused by differential propagation path lengths; and the ever present QRM can devastate a receiver's ability to copy the received data. With digital communication, we no longer have the human brain to help interpret the signal. Therefore, we need to incorporate a great deal of ingenuity into the receiver system. This ingenuity, in the form of modern communication signal processing, will help us figure out how best to transmit error-free data across such an inherently poor communication medium.

Background: Limitations of HF Data Communication

The G-TOR protocol was designed to accommodate worldwide communication over propagation paths that may experience QRM, multipath fading, random noise from atmospheric and galactic sources, and burst noise from lightning and other distant man-made and atmospheric sources. For data communication over the HF bands, there are three dominating factors: limited available bandwidth, limited signaling rate, and the dynamic time behavior of the channel.

The overcrowded amateur HF bands are characterized by frequent interference to one transmission by another from stations operating on the same frequency or a closely adjacent frequency. These conditions make bandwidth an extremely precious resource. In fact, transmission bandwidths in excess of 500 Hz are not easily accommodated in this environment. Most modern HF receivers have IF filters that can be selected to be as narrow as 500 Hz, and all digital modes on HF should use signals that are contained within this bandwidth. G-TOR employs up to a 300-baud transmission with a maximum of 200-Hz separation. While the theoretical null-to-null bandwidth is 500 Hz, in practice it is slightly wider, but not appreciably so.

Another important factor is the symbol rate allowed on the channel. This issue is not independent of the signal bandwidth requirement. However, it is noteworthy that the FCC does not currently permit digital transmissions at symbol rates greater than 300-symbols per second (bauds). This is a reasonable limitation since multipath propagation can become a serious problem with faster rates. The real culprit here is the symbol duration versus differential propagation path length. When the multipath propagation path length approaches several milliseconds (which is common on long propagation path lengths), shorter symbols (ie, higher baud rates) experience a devastating amount of destructive interference, and digital communication will be poor or non-existent. By limiting the symbol rate to no more than 300 bauds, worldwide communication should be possible under severe multipath fading conditions.

The third dominating factor for the HF channel is its characteristic dynamic time behavior. The HF channel is in a constant state of flux. This time-variant behavior imposes an upper limit on the message transmission time. Research has shown that the HF channel can change significantly over several seconds; therefore, message durations of one second or less are usually advisable to ensure a relatively stationary propagation medium. G-TOR, on the other hand, uses a message duration of nearly 2 seconds because it employs signal-processing techniques which negate these effects.

Overview and Objectives of the G-TOR Protocol

The purpose of the G-TOR protocol is to provide an improved digital radio communication capability for the high frequency (HF) bands. Since one of the objectives of this protocol is ease of implementation in existing TNCs, the modulation format consists of standard tone pairs (FSK), operating at 300, 200 or 100 bauds, depending upon channel conditions. FSK is used in part for reasons of economy and simplicity, but primarily because of the existence of a large equipment base among Amateur Radio operators. Each transmission consists of a 1.92-second frame transmitted in a synchronous ARQ mode. The cycle time for the system is 2.4 seconds, which includes a 0.48-second interval to allow for propagation time and the receipt of ACK transmissions. All of the advanced features of this protocol are implemented in the signal-processing software. For example, data compression is provided in three different forms, automatically selected on a frame-by-frame basis for best throughput, and disabled when it will not increase throughput

The real power of G-TOR resides in the properties of the (24,12) extended Golay error-correcting code, which can provide correction of up to 3 random errors in 3 received bytes. The (24,12) extended Golay code is a half-rate, invertible error-correcting code—a code in which the original information bits can be completely recovered from the parity bits alone. To help the error-correcting code alleviate the effects of burst noise and interfering transmissions, each transmitted frame is interleaved to the maximum depth allowed by the packet size. The mathematical structure of this code makes it an ideal choice for this application, as we will describe later.

An important feature of the G-TOR protocol is that a combination of forward error correction and error detection with ARQ is used to correct errors within a frame. This technique is known as *hybrid ARQ* since any errors are corrected by a combination of retransmission and the error-correction (parity) bits, which are not used as such unless they are needed. The error-detection code transmitted with each frame is a 2-byte cyclic redundancy check (CRC) code—the same one that is used in the AX.25 protocol. The CRC code is used to determine if the packet was received correctly before error correction is initiated; it is also used after error correction has been completed to ensure that the error-correction process has successfully removed all errors in the packet.

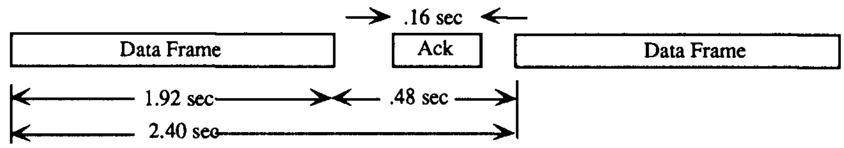


Fig 1—G-TOR ARQ system timing.

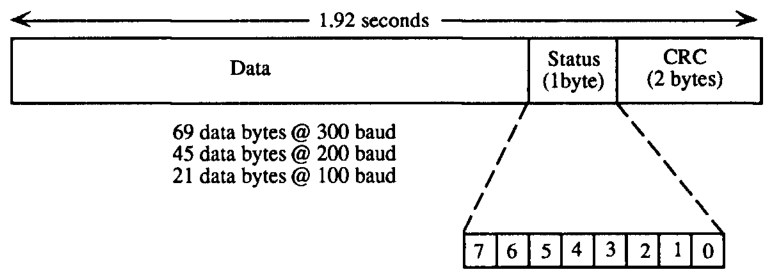


Fig 2—G-TOR frame structure before interleaving.

One of the primary causes of reduced throughput on synchronous ARQ links is errors in the acknowledgment (ACK) signal. To help reduce needless retransmissions due to faulty ACKs, G-TOR uses *fuzzy ACKs,* which allow the receiving station to tolerate a small number of errors in the ACK signal. All of these features work together in an efficient manner to increase the throughput of G-TOR.

Waveform Structure

G-TOR employs a binary-FSK transmission mode that allows data to be transmitted at three different rates, depending upon channel conditions:

- 300-baud data transmission is the default speed, and it is used until conditions begin to deteriorate,
- 200-baud data transmission is used when channel conditions deteriorate further, and
- 100-baud data transmission is used for transmitting ACKs, initiating contacts and when channel conditions are poor.

The G-TOR waveform consists of two phase-continuous tones (BFSK) spaced 200 Hz apart (mark = 1600 Hz, space = 1800 Hz); however the system can still operate at the familiar 170-Hz shift (mark = 2125 Hz, space = 2295 Hz), or with any other convenient tone pairs. The optimum spacing for 300-baud transmission would normally be 300 Hz. However, in the interest of keeping the bandwidth as close to 500 Hz as practical, some small amount of performance is traded off to save bandwidth.

G-TOR System Timing

The basic frame structure for the G-TOR protocol is based on the use of multiple 24-bit (triple-byte) words in order to be compatible with the Golay encoder. The frame timing for G-TOR is shown in Fig 1. From this diagram one can see that G-TOR is similar to Pactor and AMTOR in that it uses short-duration synchronous transmissions.

Synchronous operation is used to increase the throughput of the system in the presence of multipath fading, and to keep the number of overhead bits required to a minimum. No synchronization headers are used in this protocol; synchronization is performed using the received data and precise timing.

Research has shown that a message duration of about one second is an upper limit unless some sort of forward error correction is employed. G-TOR can operate with a 2.4-second cycle, unlike AMTOR and Pactor, because error correction and interleaving are utilized. This extended cycle time is used to increase throughput, in terms of bits/sec, and to gain efficiency, in terms of the ratio of data bits to overhead bits in each frame.

In the following sections we describe the data frame, the acknowledgment (ACK) frame and the change-over frame in detail. The connect frame and the disconnect frame, which are essentially identical in structure to the data frame, contain the call sign of both the called and the calling stations.

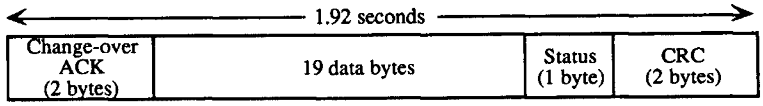


Fig 4—G-TOR changeover frame structure.

Data Frame Structure

The frame structure for a typical G-TOR data frame (before interleaving) is shown in Fig 2. The data frame is 1.92 seconds in duration. Each data frame is composed of either 72 bytes (at 300 baud), 48 bytes (at 200 baud), or 24 bytes (at 100 baud), depending upon channel conditions.

A single byte near the end of the frame is devoted to command and status functions. The status byte is interpreted as follows:

- status bits 7 and 6:
  Command
  - 00 - data
  - 01 - turnaround request
  - 10 - disconnect
  - 11 - connect
- status bits 5 and 4:
  Unused
  - 00 - reserved
- status bits 3 and 2:
  Compression
  - 00 - none
  - 01 - Huffman (A)
  - 10 - Huffman (B)
  - 11 - reserved
- status bits 1 and 0: Frame no. ID

ACK Frame Structure

The G-TOR acknowledgment (ACK) frame is shown in Fig 3. ACK frames are used to acknowledge correct or incorrect receipt of frames from the information-sending station. They also are used to request changes in transmission speed and to initiate a changeover in information flow direction. These frames are not interleaved and do not contain error-correction (parity) bits. There are five different ACK frames:

- Data acknowledgment—frame received without error (send next data frame)
- Data acknowledgment—frame error detected
- Speed-up acknowledgment frame
- Speed-down acknowledgment frame
- Changeover acknowledgment frame

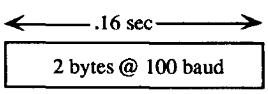


Fig 3—G-TOR ACK frame.

The ACK codes are composed of multiple cyclic shifts of a single 15-bit pseudorandom noise (PN) sequence (an extra 0 bit is appended to the sequence for balance, so the total ACK word length is 16). A pseudorandom noise sequence is used for the ACKs because PN sequences have powerful mathematical correlation and distance properties that facilitate the identification of the appropriate ACK code, even in the presence of noise and interference. We refer to this concept as a *fuzzy ACK*, in that it allows us to tolerate up to 3 bit errors within a received ACK frame.

Changeover ACK Frame Structure

The changeover frame is shown in Fig 4. This frame is always transmitted at 100 bauds and it is never interleaved. It is essentially a combination of the ACK frame structure and the data frame structure in which the first 16 bits of the frame represent one of the five ACK frames described in the previous section.

Data Compression

The Huffman data-compression code is used to remove redundancy from source data in such a way that the original data can be completely recovered or restored. Therefore, fewer bits will be traversing the channel to convey any given message. Use of data compression codes will increase data throughput and decrease the required transmission time. Both of these are valuable features to use on the HF channel. The G-TOR protocol uses run-length coding and two types of Huffman coding during normal text transmission. Run-length coding is used when more than two repetitions of an 8-bit character are sent. It provides an especially large savings in total transmission time when repeated characters are being transferred.

The Huffman code is the optimum code when the statistics of the data being compressed are known. G-TOR applies the Huffman code in two forms. The first code (Huffman A) is based on the upper- and lower-case character set. Many operators prefer to use only upper-case text. In this case a different Huffman code is used (Huffman B). The result of either type of Huffman code is to reduce the average number of bits per character. Situations do occur, however, in which the data transmitted will not benefit from Huffman coding. In these cases, the encoding process will be disabled by the transmitting station. The decision to use (or not use) Huffman coding is made on a frame-by-frame basis by the information-sending station.

Error-Correction Coding and Interleaving

A central feature of the G-TOR protocol is the Golay forward error-correcting code. This code was invented in 1949 and named after its inventor, M. J. Golay. The most useful version of this code is the *extended* Golay code, which is a (24,12) binary linear-block code capable of correcting three or fewer errors in a received 24-bit code word. The extended Golay code has numerous useful properties that have led to its use in communications and telemetry over the past 45 years. Its most notable application was in NASA's *Voyager* space mission where it was used to transmit color images of Jupiter and Saturn in the early 80s. Our reasons for choosing the extended Golay code for G-TOR are two-fold. First, the encoder and decoder are simple to implement in software, that suited our goal of developing a system that can be implemented on existing HF TNCs. The more powerful and sophisticated classes of codes, such as Reed-Solomon codes and convolutional codes, are far more complex to implement in that they require elaborate decoding schemes that are quite demanding on the TNC processor. Secondly, the Golay code has certain interesting mathematical properties that make it an ideal choice for a short-cycle synchronous communication protocol.

The valuable properties of the Golay code, other than its triple error-correcting ability, are as follows:

• The code is a rate 1/2 code, which means that the encoder generates one error-correction bit (called a *parity* bit) for every data bit to be transmitted. Furthermore, the code can be implemented in *systematic* form, which causes the 12 input data bits to appear in the first 12 positions of the code word, as shown in Fig 5. This is advantageous because it allows us to separate the data bits from the parity bits and to form two frames: one frame consisting of data bits and one of parity bits.

• The code has the rather rare property of *self-duality*. This property has several mathematical consequences, but the most valuable of these for our purposes is that the code is *invertible*. This means that the original data can be recovered by recoding the parity bits. Therefore, if we transmit a frame consisting only of parity bits, the original data frame can be recovered at the receiver by recoding the parity frame. This feature is used in the hybrid ARQ procedure.

Because of the linear block code structure of the Golay code, the encoder and decoder can be implemented simply using a table look-up procedure. An alternative implementation for the decoder, which requires far less memory, uses the well-known Kasami decoding algorithm.

Error-correction coding inserts a controlled amount of redundancy into each (triple-byte) word so that errors that occur in the receiving process due to low signal-to-noise ratio or multipath propagation can be corrected. However, most error-correcting codes are effective at correcting only random errors. When burst errors occur due to lightning or interference, the error-correcting capability of most error-correcting codes is exceeded. Therefore, if burst errors are expected to be a problem (and on the HF band, they certainly are) then some protection must be included in the protocol to help alleviate the resulting errors. The conventional approach to solving this problem is called *interleaving*. Interleaving, which is the very last operation performed on the frame before transmission and the first operation performed upon reception, rearranges the bits in the frame so that long error bursts can be randomized when thedeinterleaving is performed. The interleaving process reads 12-bit words into the registers by columns and reads 48-bit words out of the registers by row, as illustrated in Fig 6. The deinterleaver simply performs the inverse, reading the received data bits into the registers by row and extracting the original data sequence by reading the columns. If a long burst of errors occurs—say, 12-bits in length—the errors will be distributed into 48 separate 12-bit words before the error correction is applied, thus effectively nullifying the long burst. Both data frames and parity frames are completely interleaved.

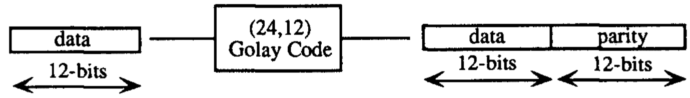


Fig 5—The (24,12) rate 1/2 Golay encoder.

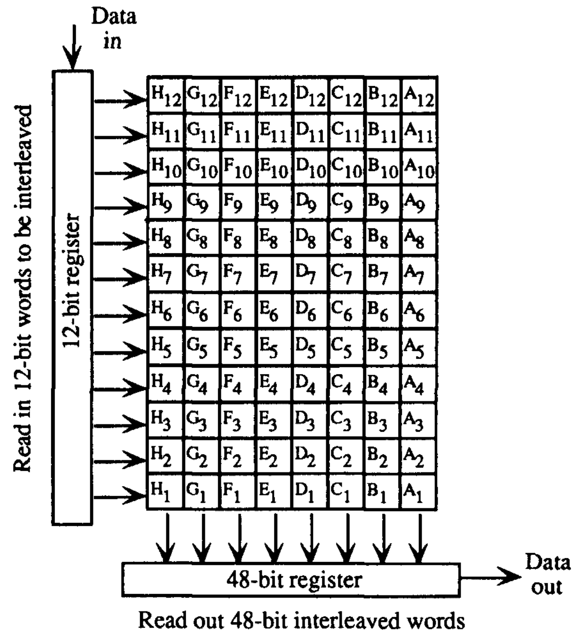


Fig 6—Interleaving the bits to be transmitted.

Error Detection and Hybrid ARQ

The ARQ concept is familiar to hams who work the digital modes. ARQ is based on the use of an error-detection code to determine if an error has occurred in the received frame. If the frame is received correctly, a positive reply acknowledgment (ACK) is sent to the sending station. However, if the frame is received in error, the receiving station sends a negative acknowledgment (NAK) to the sending station, which prompts a retransmission. Error detection is performed in G-TOR using the international standard 16-bit CRC code. This is the same CRC code that is used in the AX.25 packet protocol.

Hybrid-ARQ is a variation of the conventional ARQ concept. It uses a CRC error-detection code to detect the occurrence of errors in every frame; however, it allows the use of forward error correction when errors are detected. When no errors are detected, the forward error-correcting procedure is not used. This is a valuable concept since forward error correction is very costly in terms of throughput. For example, we previously described the (24,12) Golay code used in G-TOR as a half-rate code, with one error-correction bit required for every information bit. If this code were in use at all times, it would cause a 50% decrease in throughput.

The half-rate invertible Golay code provides an interesting dimension to the hybrid-ARQ procedure. After the message to be sent is processed through the Golay encoder at the transmitter, the message (data) bits are separated from the parity bits and two complete frames are formed. One frame consists of data bits and the other consists of parity bits. Keep in mind that the receiver can recover the message bits from the parity bits alone, so either frame can be used to transmit the data, as long as the receiver knows which to expect. *The procedure used in G-TOR is to alternate data frames and parity frames.* For example, we might expect the following sequence of frames to be transmitted when no errors are detected at the receiver:

D1 P2 D3 P4 D5 P6

Where D1 is the first frame transmitted (a data frame), and P2 is the second frame transmitted (a parity frame), etc.

When an error is introduced into a frame because of noise or interference, the receiver will detect the error and request a retransmission. When the sending station receives the request, the complementary portion of the frame (either data or parity) is sent. For example, the sequence illustrated below demonstrates a situation where two retransmissions are sent:

D1 P2 D3 P3 D4 P5 D6 P7 D7 P8

Retransmissions occurred for Frame 3 and Frame 7. In the case of Frame 3, one or more data bits were received in error and a retransmission was requested. Instead of resending D3, the corresponding parity frame P3 is sent. When P3 arrives at the receiver, the first action that occurs after deinterleaving is to convert the parity bits to data bits and check for errors. If no errors were detected, the data is accepted and a new frame (D4) is requested. If P3 fails the CRC check, this frame in its original form (ie, parity bits) is combined with D3 and the combination (D3P3) is decoded by the Golay decoder to correct the errors that occurred in both D3 and P3.

In the case of Frame 7, the parity bits were sent first in P7. After deinterleaving the frame, converting the parity bits to data bits and performing the CRC check, the presence of errors prompts the receiving station to request a retransmission. Since the parity frame has already been sent, the sending station transmits the data frame, D7. Upon receipt, the data frame is deinterleaved and checked for errors. If no errors are detected, the data bits are accepted and a new frame (P8) is requested. If errors are detected, data and parity bits are combined (D7P7) and decoded by the Golay decoder. Using this scheme, with two transmissions, the receiver can make three independent attempts to correct any errors that may occur in any frame. If this process still fails to produce an error-free frame, a retransmission is requested.

A functional block diagram of the G-TOR system is shown in Fig 7. In this diagram, S1 alternates between position 1 and position 2 as data and parity frames are alternated for transmission. S3 at the receiving station tracks S1 at the sending station to route the data or parity bits in the appropriate direction at the receiver. S2 is in position 1 when the system is functioning as the information-sending station, and position 2 is used to send ACKs when the system is the information-receiving station. S4 is in position 1 when a data frame is received, and in position 3 when a parity frame is received. Position 2 is used when the data and parity must be combined for

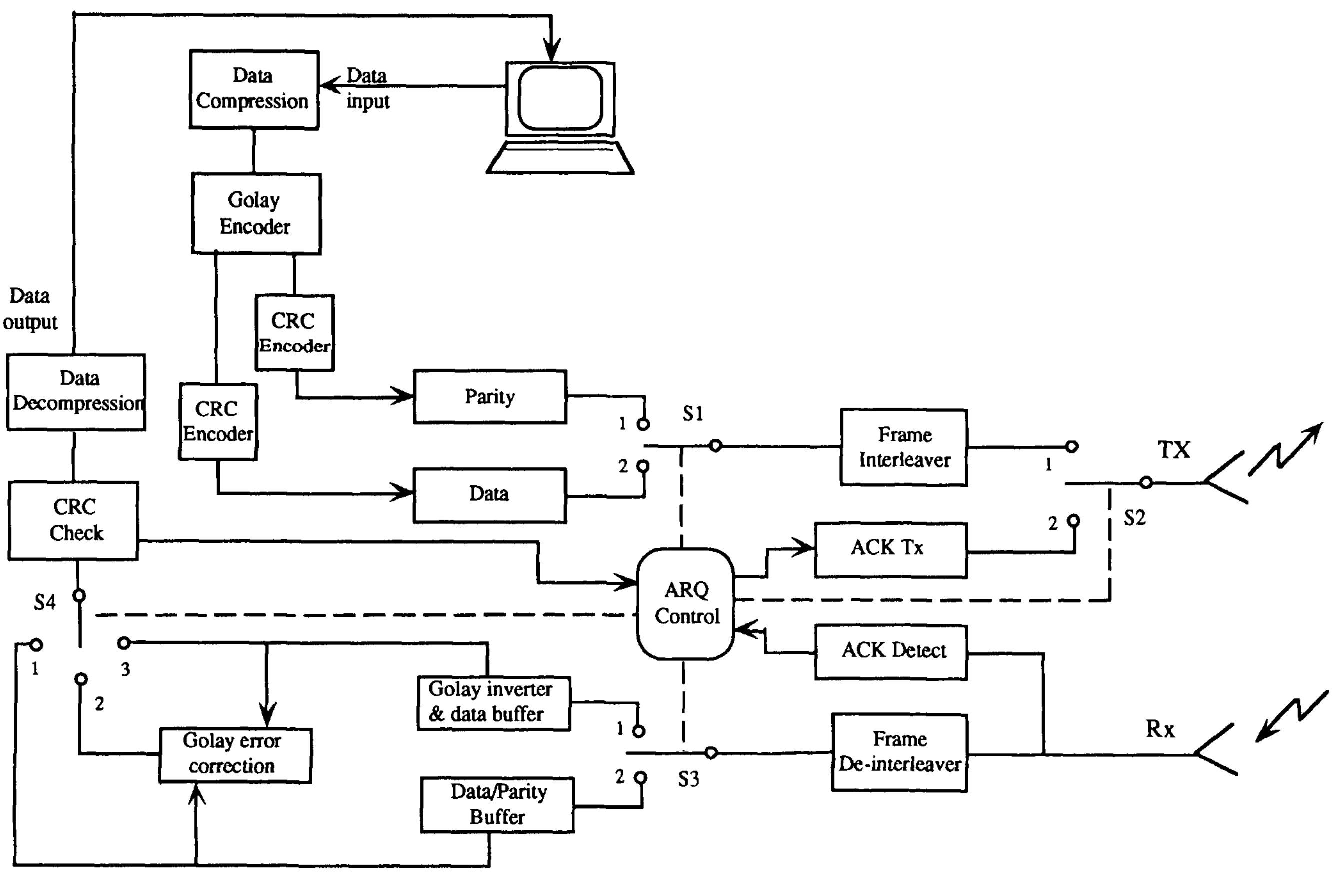


Fig 7—G-TOR simplified functional block diagram.

G-TOR vs. Pactor vs. AMTOR: A Nonscientific Throughput Test

By Steve Ford, WB8IMY
ARRL Assistant Technical Editor

As a KAM-Plus owner, I was intrigued to say the least when I read about the debut of G-TOR. The announcement made substantial claims about G-TOR's ability to surpass Pactor and even rival Clover in terms of overall throughput. Not that I disbelieved Kantronics' claims, but this was something that I had to witness in my own shack. I contacted Phil Anderson, WØXI, at Kantronics and purchased a replacement EPROM for my KAM Plus. It arrived within days and soon I was on the air.

My first test was with Karl Medcalf, WK5M, of Kantronics. We met March 3 at 2130 UTC on 14.085 MHz. I was running an ICOM IC-745 transceiver connected to a dipole approximately 30 feet off the ground. My average RF output was 100 W. WK5M was using a similar setup, although his antenna was a triband beam on a 50-foot tower. When we established communication, Karl gave me a 579 signal report. He was booming into my station with a 599+. There was moderate QRM. Neither of us were using signal filtering beyond that provided by our 2.3-kHz IF filters.

The first test was with Pactor. Karl sent a 9718-byte ASCII file that I received within 8 minutes and 15 seconds for an average data rate of 157 bit/s. We switched to G-TOR and Karl resent the file. This time I received the complete text in only 3 minutes and 35 seconds for an average data rate of 362 bit/s. To illustrate the robust nature of G-TOR, Karl sent the file a third time, but with his RF output reduced to only 5 W. Throughput suffered as you'd expect, but the entire file still made it across at an impressive rate of 290 bit/s (see Fig A).

G-TOR was clearly superior to Pactor under average conditions, but how would it fare under truly abominable conditions? (Karl and I didn't have time for an AMTOR test and I was curious to see how that would compare, too.)

Larry Wolfgang, WR1B, ARRL Senior Assistant Technical Editor, upgraded his KAM with G-TOR firmware and he met me on the air at 14.070 MHz on March 18 at 0200 UTC. His transceiver is a Kenwood TS-820 working into a triband beam pointed *away* from my location. Once again, neither of us were using filtering beyond that provided by our fixed IF filters (2.3 kHz).

The band had effectively shut down for the evening. Even though we enjoyed interference-free conditions, noise levels were very high. Since Larry lives only 40 miles away, I received his 100-W signal via groundwave. It was extremely weak. I could *just* hear his AMTOR FEC test transmissions above the noise (actual copy was spotty). Larry said that I was somewhat louder on his end, but barely. These were the awful conditions I was hoping for!

We began with AMTOR. I sent a 3934-byte ASCII text file which Larry received within 20 minutes for an average data rate of 26 bit/s. The link was maintained continuously throughout the test, although Larry noted several errors in the received text.

Pactor fared somewhat better. The same file was sent in 16 minutes and 6 seconds for an average data rate of 33 bit/s. Larry received the file error-free. (Both of us had Huffman encoding enabled.)

Finally, we tried G-TOR. Larry received the file in 6 minutes and 7 seconds for an average data rate of 86 bit/s—2½ times faster than our Pactor test! At one point during the transfer I was astonished to see my KAM indicate that it had momentarily surged to a data rate of 300 bit/s. The results of all three tests are illustrated graphically in Fig B.

Signal conditions remained fairly constant throughout the tests. We attempted to transfer the file using HF packet, but it was impossible to establish and maintain a connection. It would have been interesting to attempt a Clover test over the same path, but neither of us were equipped for that mode.

In my opinion, G-TOR turned in a remarkable performance—especially when you consider the relatively low cost of the mode. I'd be curious to see a thorough series of tests using all five error-detecting modes (G-TOR, AMTOR, Pactor, packet and Clover) under a variety of conditions. With only two test sessions to report, my results border on being anecdotal. According to Phil Anderson, however, they are representative of the overall performance of G-TOR.

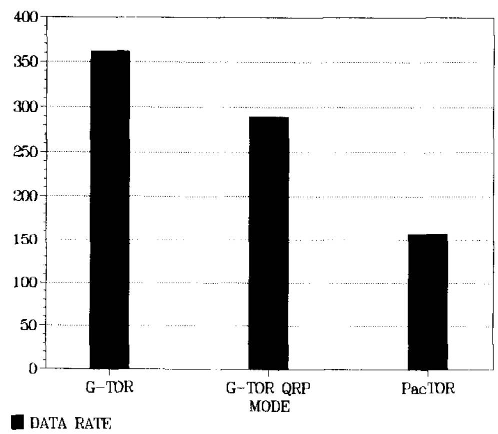


Fig A—A comparison of Pactor vs G-TOR in tests between WB8IMY and WK5M using a 9718-byte file. The test took place on 20 meters with good band conditions and moderate QRM.

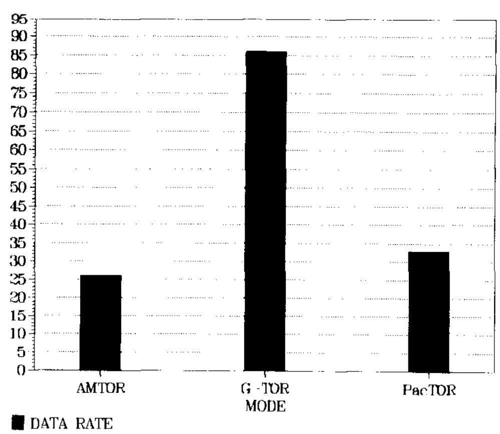


Fig B—Average data rates during tests on 20 meters between WB8IMY and WR1B. The size of the test file was 3934 bytes. Received signals were weak at both stations and noise levels were high. We attempted the file transfer using HF packet, but conditions were too poor to permit a connection.

error correction to be performed.

G-TOR Performance

Initial testing with G-TOR was conducted during the month of January 1994, between Lawrence, Kansas, and Laguna Niguel, California. During these tests, TRACE was set ON at each station, enabling the raw data display of frames received with and without the aid of forward error correction and interleaving. The results were somewhat surprising. While Pactor often dropped in transmission speed from 200 to 100 bauds, G-TOR nearly always kept on operating at 300 bauds. Enough frames were being corrected to keep the system running at maximum speed, regardless of man-made interference and mild multipath conditions. Transfer duration for the entire test files varied from 12 to 27 minutes for Pactor, but only 5.5 to 7.5 minutes for all but one G-TOR transfer. G-TOR simply maintained its highest pace better than Pactor, resulting in a substantial increase in average throughput.

Typical test results are shown in Table 1. In this table, *KAM-E* indicates the Kantronics All Mode (KAM) TNC with enhancement board; *KAM +* indicates the new KAM-Plus; and *SCS-PTC* is the German TNC that was used as a receiving station to observe the effect of using their particular implementation of Memory ARQ.

Table 1
Results of Recent G-TOR Testing.

Mode	Date Time	Local Unit	Distant Unit	Distant Call	Band	File size	Transfer time (m:s)	Char/sec
G-TOR	3/14/94 1600	KAM-E	KAM-E	WA4EGT	20 m	9718	4:18	37.66
Pactor	3/14/94 1600	KAM-E	KAM-E	WA4EGT	20 m	9718	8:10	19.83
G-TOR	3/14/94 1600	KAM-E	KAM-E	WA4EGT	20 m	9718	3:45	40.79
Pactor	3/14/94 1600	KAM-E	KAM-E	WA4EGT	20 m	9718	9:32	16.98
G-TOR	3/15/94 1530	KAM-E	KAM +	W5JTB	40 m	3269	1:35	34.40
Pactor	3/15/94 1530	SCS-PTC	KAM +	W5JTB	40 m	3269	3:07	17.40
G-TOR	3/15/94 1530	KAM-E	KAM +	W5JTB	40 m	3269	1:55	28.40
Pactor	3/15/94 1530	SCS-PTC	KAM +	W5JTB	40 m	3269	2:35	21.00
G-TOR	3/15/94 1530	KAM-E	KAM +	W5JTB	40 m	3269	1:10	46.00
Pactor	3/15/94 1530	KAM-E	KAM +	W5JTB	40 m	3269	2:35	21.00

G-TOR totals: Average ch/sec----> 37.45
Pactor totals Average ch/sec----> 19.24

Long before any on-air testing was performed, G-TOR was being developed using a combination of engineering analysis, software simulation and laboratory testing. After the initial system design, software simulation at the communications link level provided insight to system performance in noise. The next step was to take the fully implemented protocol and examine its performance using a real-time HF simulator. The simulator, which was developed by the University of Kansas, was implemented on a Texas Instruments TMS320C30 digital signal-processor system. The simulator was programmed to provide a noisy, fading multipath environment for the testing. In particular, we were able to test the G-TOR protocol by using the good, moderate, poor and flutter fading channels prescribed by the CCIR as recommended for simulator test channels.[5] The final test phase included on-air testing with beta testers distributed throughout the US.

On-air tests have shown G-TOR to be even better than our simulator predicted. G-TOR is found to have the ability—through a combination of coding and interleaving—to "hang in there" when channel conditions get tougher. The distribution of time to send a given binary file tends to be much narrower for G-TOR than for Pactor.

Conclusions

Advances in equipment state-of-the-art are usually incremental. Each inventor adds improvements to the work of past inventors. These changes give the new system its character and set the stage for further advances. New products and concepts are not necessarily *better* in every regard—perhaps they should be described as different or evolutionary. So it is with the history of innovation for HF digital-communication systems for the Amateur Radio service. G-TOR builds on Pactor, Pactor built on AMTOR, packet built on the X.25 protocol, AMTOR built on commercial TOR systems and RTTY, and so on.

G-TOR's evolutionary and distinct contribution to Amateur Radio is that it implements modern digital-communication techniques in order to increase data throughput on the HF bands. Furthermore, the protocol is simple enough to be implemented on multimode TNCs. We believe this protocol will continue to be valuable when DSP-based TNCs become widely available. With DSP-based TNCs, more advanced forms of modulation, more extensive use of the Golay codes (possibly to include memory ARQ), and adaptive equalization will be possible. G-TOR has the essential characteristics to be a useful protocol for years to come.

G-TOR is an innovation of Kantronics, Inc, and it is being shipped as a standard feature on the multimode TNC, the KAM-Plus. For all current owners of the KAM-Plus or the KAM with enhancement board upgrade, the new protocol is available as a ROM upgrade for a small fee. Although Kantronics has applied for a patent on G-TOR, the protocol in all its detail will be described to the Amateur Radio community through publications currently in the works.

Notes

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

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

[3] Wickwire, Ken, "The Status and Future of High Frequency Digital Communication, Part II: HF Modems and Their Performance," *QEX*, July 1992, pp 3-15.

[4] McLarnon, Barry, "More Ideas for a Robust HF Packet Signal Design," *QEX*, March 1990, pp 9-10.

[5] *Recommendations of the CCIR*, 1990, Volume III, (Fixed Service at Frequencies Below About 30 MHz), Recommendation 520-1, p 28.

5760 MHz from the Junkbox

Low cost and high performance in one design.

By Robert Cook, N2SB

The 5760-MHz ham band is an interesting one. It is close to a point in the spectrum where the natural noise level is at its lowest, and it is just below a heavily populated commercial microwave band. Until recently, hams interested in the microwave bands skipped over this band and went directly to 10 GHz.

In the last few years there have been a lot of changes in the commercial 6-GHz microwave bands, and a lot of goodies have shown up at hamfests at bargain basement prices.

In addition to lots of good surplus now becoming available, Al Ward, WB5LUA, has taken the grief out of microwave amplifier construction with his designs based on FETs.[1] His designs work first time and every time and are easy to reproduce.[2]

This article describes a transverter for 5760 MHz. Design and construction are based on what can be picked up cheap at hamfests and from other microwave enthusiasts' junk boxes. There are no sacrifices made to compromise performance and there was no attempt to make it small or portable. Every attempt was made to combine high performance and low cost.

Notes appear on page 24.

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Modular Building Block Approach

The construction technique used here is to build each stage as a standalone module with connections for dc

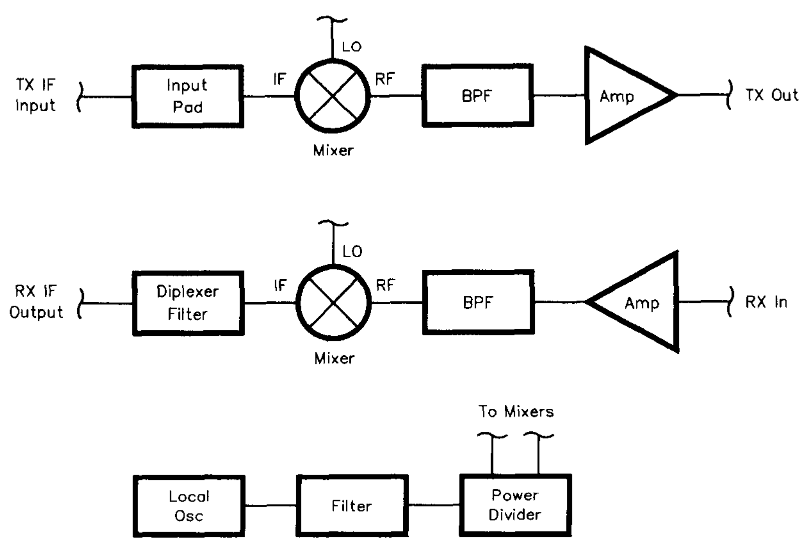


Fig 1—Block diagram of a conventional transverter, which needs two mixers and two band-pass filters.

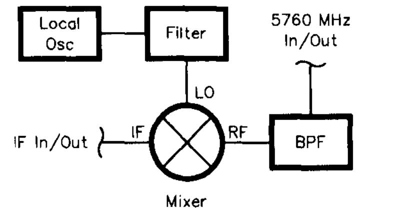


Fig 2—An "unconventional" transverter needs two coaxial relays for signal switching.

and RF. Each module can be tested or modified without disturbing the stages before or after it in the transverter. If something does not perform up to expectations or has a problem, it can be easily isolated and repaired or replaced. The design can be changed and the performance re-evaluated when a new part or assembly from the latest hamfest is added. This modular approach to building makes modification and experimentation very easy.

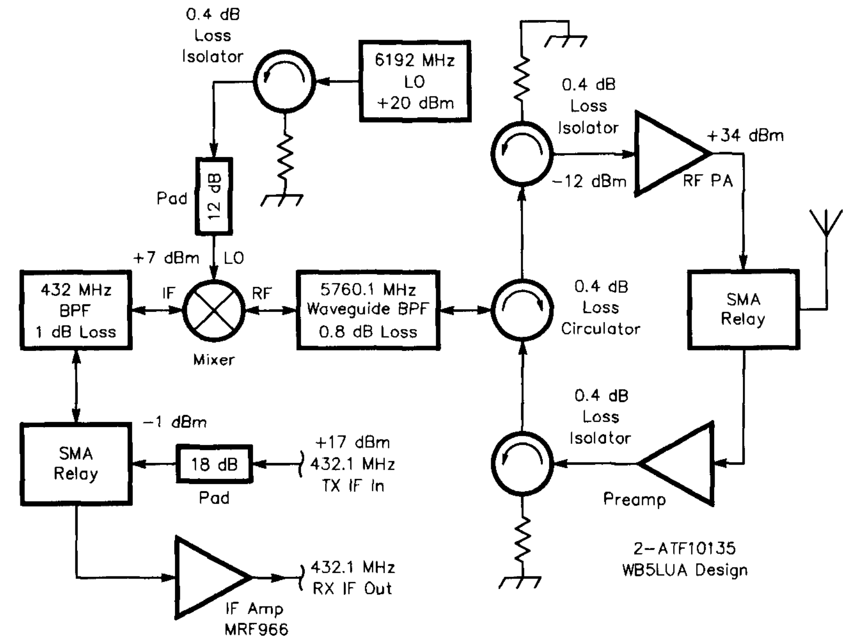


Fig 3—This transverter approach uses one mixer, one filter and one coaxial relay.

Design Considerations

Initially, the conventional transverter approach was the design goal (see Fig 1). This approach requires two band-pass filters after the two mixers, a power divider of some sort after the local oscillator to drive the mixers, and a local oscillator and filter.

The advantage of the conventional transverter approach is that you need only one good coaxial relay to switch between transmit and receive. The disadvantage is that you need two mixers and a power divider, which are notoriously difficult to find at hamfests.

A second approach is to use the bidirectional aspect of mixers and build an unconventional transverter (see Fig 2). This approach eliminates one mixer, one band-pass filter and a power divider from the parts line-up. The problem with this approach is the need for an additional coaxial relay to switch between the transmit and receive amplifier modules after the bandpass filter.

The third and final approach addresses the problems of the first two designs, incorporates the advantages of both and combines high performance with low cost. (See Fig 3) There is only one mixer, one band-pass filter and one coaxial relay. The second coaxial relay after the band-pass filter is eliminated with two isolators and one circulator. There is no performance sacrifice, and the parts have been readily available at hamfests for the last few years.

Module Construction and Modifications

The first item I acquired at a hamfest was a Frequency West microwave signal source. This is the fastest and easiest way to get a local oscillator functioning on the microwave bands. The time devoted to getting a traditional discrete-component local-oscillator/multiplier chain working can instead be devoted to actually looking for weak signals on 5.7 GHz.

These sources are cavity-type oscillators operating in the 1100-1500 MHz range, multiplied up to the operating frequency and phase-locked to a reference. The reference is either internal or external depending on the type of source. There is an informative article based on this type of oscillator in the *Proceedings of Microwave Update '87*.[3]

This particular source is a model MS-56X0-49, which runs about +20-dBm output after retuning and has an internal reference crystal. The crystal operating frequency was determined by dividing the chosen output frequency of 6192 MHz by 65. Other multiplier ratios can be chosen as long as the tuning range of the oscillator will extend to the new crystal and output frequency.

The cavity and output multiplier were then tuned for maximum output at 6192 MHz with a spectrum analyzer. The phase-voltage monitor point on the side of the source is about –7 V when the source is locked.

6192 MHz was chosen because there was a crystal available from another source on 95.2615 MHz (65 × 95.2615 = 6192). This is only 163 MHz outside of the tuning range marked on this source. It retuned easily to this new output frequency and phase-locked every time power was applied.

A new crystal can be ordered from International Crystal Manufacturing in Oklahoma City, 405 236-3741, for about $20. The part number is 585132.

When ordering a new crystal, keep in mind that the cavity and frequency multiplier of the source are designed to operate over a specific frequency range. Select your LO frequency to accommodate the tuning range of the source you obtain. Some sources will tune farther out the band marked on the case than others.[4]

The chosen LO injection frequency is higher than the operating frequency of 5760 MHz. This is referred to as high-side injection. High-side injection causes the IF rig to tune backwards and inverts the sidebands. LSB at the IF is USB on 5760 MHz, and tuning the IF rig lower in frequency increases the operating frequency on 5760 MHz. This small compromise did not affect the performance of the transverter, saved the price of a new crystal and stayed reasonably close to the original frequency range of this particular source.

In this design, the LO output goes into an isolator and then into a 12-dB attenuator to drop the level to +7 dBm to drive the mixer. With this arrangement, the LO is unaffected by changes in load on the output. The mixer used in this transverter is a Watkins-Johnson double-balanced type. I used it because that is what I could scrounge up. It has about 8 dB of conversion loss and 25 dB of LO-to-RF isolation. The mixer you select will be determined by what you can find. Look long and hard for the best mixer. The mixers I tried before the WJ mixer were all inadequate. The rat-race type and the diode-ring type were particularly sensitive to mismatches on the RF and IF ports, suffered from low LO-RF isolation and were a real pain to construct and tune.

I found the band-pass filter and the isolators/circulators at a hamfest in a box under a table. The box contained lengths of UT-141 semirigid coax, two waveguide filters and four waveguide-to-coax transitions and five circulators with loads that fit the ports. I bought the whole box for $3 with no haggling by either the buyer or the seller. The waveguide filters are 5-pole units tuned to the low end of the 6-GHz band, and the isolators/circulators are for the same frequency range but with a different bolt pattern. A few minutes with a drill fixed the problem of mating the filter to the circulator, and the isolators already fit the circulators. A few more passes with a drill allowed the waveguide-to-coax transitions to be mated to the filter on one end and the isolators on the other end.

The first filter was tuned with a signal generator and a spectrum analyzer. At 5760 MHz the insertion loss is 0.8 dB and the 3-dB bandwidth is 66 MHz. At the LO frequency and image frequency the filter rejection is more than 95 dB.

A second filter, tuned on a Hewlett Packard 8720 network analyzer, showed the same performance as the one tuned with the spectrum analyzer and signal generator, but the passband ripple was smoothed out to less than 0.3 dB over a 25-MHz bandwidth.

In both cases, the filters had a 35-dB or better return loss at 5760.1 MHz, and the return loss stayed at 25 dB or better at ±12.5 MHz from center frequency.

The image rejection, bandwidth and insertion loss of these two waveguide filters are typical of the types of filters found in commercial 6-GHz equipment. If you can find filters like these, 144 MHz could be used as the IF. An LO source that covers 5616 MHz for low-side injection or 5904 MHz for high-side injection would be necessary. Both filters measured more than 55-dB rejection at ± 144 MHz from 5760.1 MHz.

The addition of a circulator and two isolators to the end of the filter eliminates the need for a good coaxial relay at this point in the transverter and provides 50 dB of isolation between the transmit and receive paths.

On the receive path I used a 2-stage FET preamp constructed from a kit from Down East Microwave. This is a WB5LUA design and, other than setting the bias on the two FETs, there are no adjustments necessary for correct operation. The noise figure was measured at 1.2 dB and the gain at 17 dB. It worked from the first time it was turned on, and the gain and noise figure were not measured for more than a year after it was built. The measurements were made mainly out of curiosity; it was working just fine.

The transmit amplifier is one found in the junk box of a fellow Pack Rat. This amp is an Avantek model AWP64100. It has 46 dB of gain and its maximum output at 5760 MHz is +40 dBm. (*That's 10 watts.*) It draws 2.8 A at –24 V dc. A second amp, purchased at a later hamfest for $20, puts out +33 dBm. A nice unit for a spare. I picked up a third unit of the same type for $5. It was not working but had plenty of spare parts.

The transverter has –12 dBm out of the filter/circulator/isolator combination and, with the amp, transmit output is +34 dBm.

A small surplus SMA relay from another hamfest finishes the 5.7-GHz part of the transverter.

The 432-MHz IF side of the mixer goes to a small filter tuned to 432.1 MHz as the center frequency. I got this filter at a hamfest because it is small, has SMA-type connectors and a lot of Johanson-type piston trimmers. It looked like it might be for some microwave frequency above 2 GHz. It turned out to be tunable to 432 MHz. This filter was purchased years before I decided to get on 5.7 GHz and languished in the junkbox until a use for it could be found. It has a 3-dB bandwidth of 12 MHz and about 1.2 dB of insertion loss. Other filters could also be used here. There are examples of suitable filters in recent editions of the *ARRL Handbook*.

The filter is preceded by a relay to switch the IF between transmit and receive. A PIN diode switch is also suitable, but use what you have on hand. The transmit path includes enough attenuation to limit the IF drive to the mixer to about –2 to –3 dBm. This keeps the IF transmit signal from overdriving the mixer and causing distortion.

The receive path includes a post-mixer amplifier. (See Fig 4.) This amp is based on the Motorola MRF-966 dual-gate GaAs FET.[5,6] There is about 20 dB of gain, and the noise figure is about 1 to 1.5 dB. There is nothing original about this design. The input circuit was taken from a design in the 1986 *ARRL Handbook* and the output circuit is from an article by Joe Reisert, WlJR, in *Ham Radio* magazine from December 1987.

This post amp is easy to build and get operating and it is stable. But for use in the 5760-MHz transverter it has too much gain. Changing R1 to a 10-kΩ pot allows the gain to be adjusted to a comfortable level for the IF receiver. Or you can add attenuation to the output until the noise from the transverter lets the S-meter of your IF rig rest somewhere near zero. Or you can leave the S-meter resting at S6 and give everyone you work a good signal report!

Physical Construction

The MRF-966 amp module is built into a box made from old scrap G-10 circuit board. The 2-stage preamp has a small brass side rail all the way around it. These and the other modules are mounted on a small drawer designed with 19-inch rack-mount flanges on the front.

Dc power is provided by small, open-frame type, regulated supplies commonly found at hamfests for $5-10.

This transverter needs +12 V for the post amp and the preamp, –24 V for the transmit amp and relays, and –9 V for the LO.

The connectors are all good quality SMA or N type. Interconnecting cables are made from UT-141. UT-141 semirigid coax often can be picked up at hamfests with connectors already installed for 50 cents a length. SMA and N connectors can often be found for $2 each or less. Saving money by using cheaper connectors only leads to poor microwave connections and poor performance. There is nothing more frustrating than removing that BNC connector after the module has already

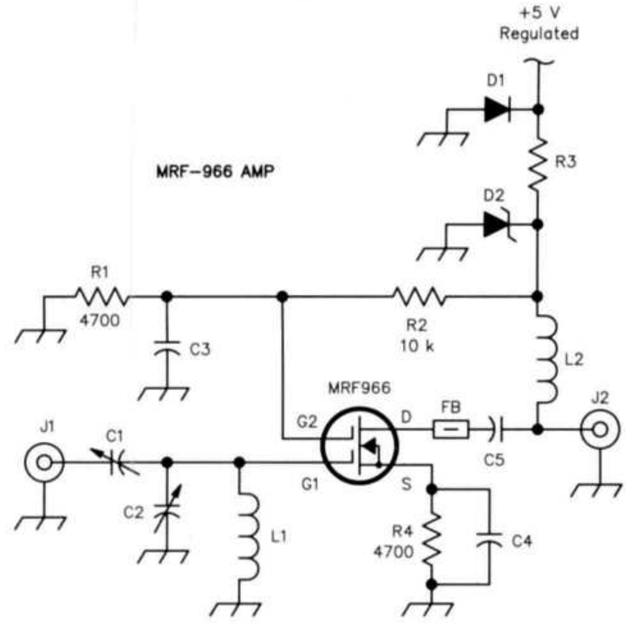
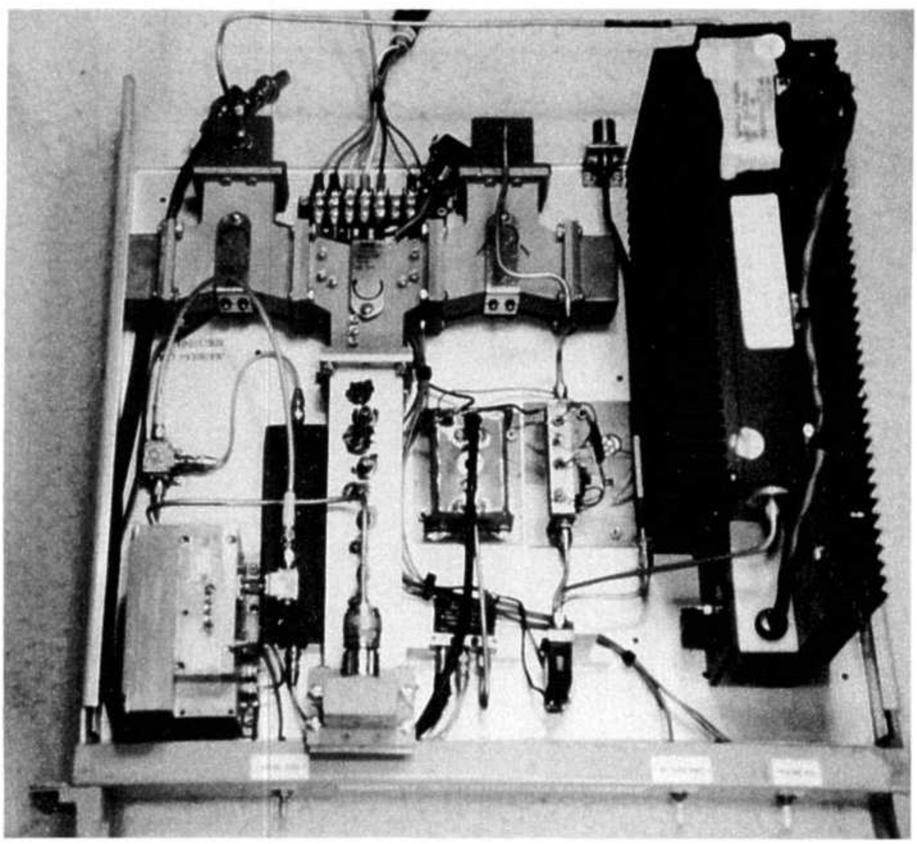


Fig 4—Schematic diagram of the MRF-966 post-mixer amplifier.

L1—4-turns, no. 20, 3/16-inch ID, spaced 1 wire diameter.
L2—5-turns, no. 24, 1-inch ID, 5 inches long.
R1—4.7 kΩ, 1/8 W
R2—10 kΩ, 1/8 W
R3—100 to 150Ω, 1/2 W
R4—220Ω, 1/8 W
C1,2—1 to 10 pF
C3,4—470 to 1000 pF chips
C5—3.9 pF
CR1—1N4007
CR2—5.6-V Zener, 1/2 W
FB—Ferrite bead



been built to replace it with a good connector. (It is really hard to make a hole smaller.)

Interfacing

My IF rig is an ICOM IC-451A. Its transmit power amplifier has been removed and the exciter drive brought out to a connector mounted in a spare hole in the back panel. There is +17 dBm available from the driver. The receiver antenna connection is brought out to the original antenna connector. Transmit and receive are separate now, and it is impossible to generate high power into the post amp or the mixer in the transverter.

The 5760-MHz transverter is keyed from the switched positive voltage which was originally used to run the IC-451A internal power amplifier. The circuit in Fig 5 is a simple dc switching circuit which will take an applied ground or an applied positive voltage and turn on a PNP transistor. It is used to pull in a small dc relay that applies +12 V on receive to the transverter's preamp and post amp and powers the antenna and IF relays on transmit. The PNP TIP-42 transistor can be changed if more current capacity is needed.

Performance

The transverter works well. The receive sensitivity measures –126 dBm for a 10-dB S/N in the 432-MHz IF radio, and image rejection is better than 95 dB. The transmit power is +34 dBm. The spurs from the LO are all more than 60 dB down, with most down below 80 dB.

The first contact with Ron Whitsel, WA3AXV, was anticlimactic; signals were S-9 plus. Ron's signal measured –73 dBm with a spectrum analyzer connected to my feed line. The path between our locations is about 17 miles and not totally free from obstructions. Signal levels drop significantly in the months when there are leaves on the trees, but the SSB signals are still Q5.

Conclusion

The 5760-MHz transverter described in this article combines high performance and low cost in a single unit. There were no sacrifices made to compromise performance.

Between attending a few hamfests and bartering with other members of The Mt Airy VHF Radio Club, the total cost of all the components came to less than $150.

The project does not need to be duplicated exactly as described. Judicious selection of components found at hamfests and in other microwave enthusiasts' junkboxes will determine exactly the form it will take on your workbench.

Acknowledgments

I would like to thank Ron Whitsel, WA3AXV, Dave Mascaro, WA3JUF, and Gary Hitchner, WA2OMY, for advice and ideas used in the design and implementation of this project and other microwave projects over the last few years.

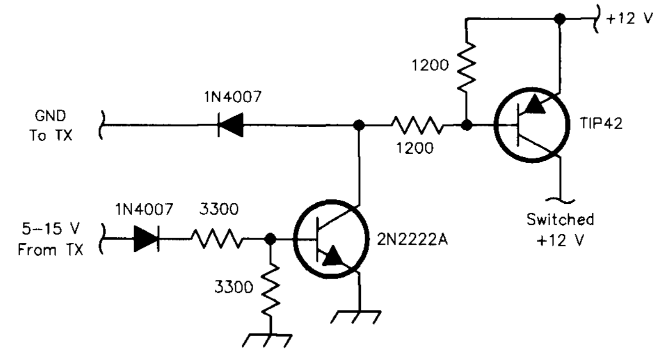


Fig 5—Simple keying circuit to provide keyed-line and switched +12-V control from a 5- to 15-V dc keying signal.

Notes

[1]Ward, Al, WB5LUA, "Simple Low-Noise Preamplifiers," *QST,* May 1989.
[2]WB5LUA-style preamps and preamp power supplies are available from: Down East Microwave, RR#1, Box 2310, Troy, ME 04987, tel 207 948-3741.
[3]Ericson, Keith R., KØKE, "Phase Locked Source Update," *Proceedings of Microwave Update '87,* pp 93-95.
[4]Subsequent attempts to tune the output filter on this source to 5328 MHz for low-side injection were unsuccessful. The PLL would lock on the new cavity frequency but the multiplier filter would not tune to the lower frequency.
[5]*The ARRL Handbook,* ARRL, 1986, Chapter 32, p 32-2.
[6]Reisert, Joe, WIJR, "VHF/UHF and Above," *Ham Radio Magazine,* December 1987, pp 74-75. □□

Software Review

SysCalc for Windows, Version 1.06a, Arden Technologies, 119 Cedarwood Court, Forest, VA 24551, tel 804 525-6837.

Reviewed by David Newkirk, WJ1Z
ARRL Senior Assistant Technical Editor

So—you're toying with different receiver-front-end options in your never-ending quest for the (most) crunch-proof receiver you can (afford to) build (with the time and components available). How would your baby's overall performance change if you used a stronger mixer with a slightly lower noise figure and replaced your current post-mixer single-ended-amplifier-and-pad combo with a diplexer and stronger push-pull amplifier? What if—on the bands where your radio's noise figure vies with atmospherics for setting system sensitivity—you also tossed a strong 10-dB gain stage in front of the mixer?

Radio technologists have long had sufficient math and theory in place for calculating this stuff; it's no big deal. Aside from envelope backs and sharp pencils, what they—all of us, for that matter—keep running short of is *time*. Whether the system change you propose is profound or prosaic, you have to go through the same dance every time you want to crunch the numbers. If the system is at all complex, tracking the effects of changes as they ripple from subsystem to subsystem is a pain at best.

This bit of *QEX* reviews a software product worth considering if you're interested in doing (and doing and doing and doing) just this sort of what-if dance with systems in which third-order-intercept, noise figure and gain—in other words, almost any radio system worth building—will not be left to chance. The software, a *Windows* application, is Arden Technologies' *SysCalc* 1.06a.

SysCalc 1.06a can be thought of as a combination system-performance spreadsheet and block-diagram drafting tool. You add/move/delete system blocks (amplifiers, mixers, filters, resistive pads, generics), label them appropriately, and enter their noise figure, gain and third-order intercept values in the appropriate cells. You provide system inputs: power of input carriers (dBm); level of third-order-IM products already present on the input carrier, if any (dBc); system bandwidth (MHz); and source (or antenna) temperature (K). You can do what-if scenarios at will, and you can print your results.

SysCalc calculates and displays, on a block-by-block basis: power out (dBm); NF+ (how much a given block degrades the system's total cascaded noise performance; dB); and IP3+ (how much a given block degrades the system's total cascaded third-order intermodulation performance; dB). On a given system sheet, *SysCalc* highlights in red the NF+ and IP3+ numbers that most degrade system performance.

On a sheet-by-sheet basis, *SysCalc* also returns: S/N ratio (dB); spur-free dynamic range (dB); minimum detectable (aka *discernible*) signal (MDS); effective system noise temperature (K); third-order input intercept (dBm); absolute IM-product level (dBm); and relative IM level (dBc).

Running It

Running *SysCalc* 1.06a and loading

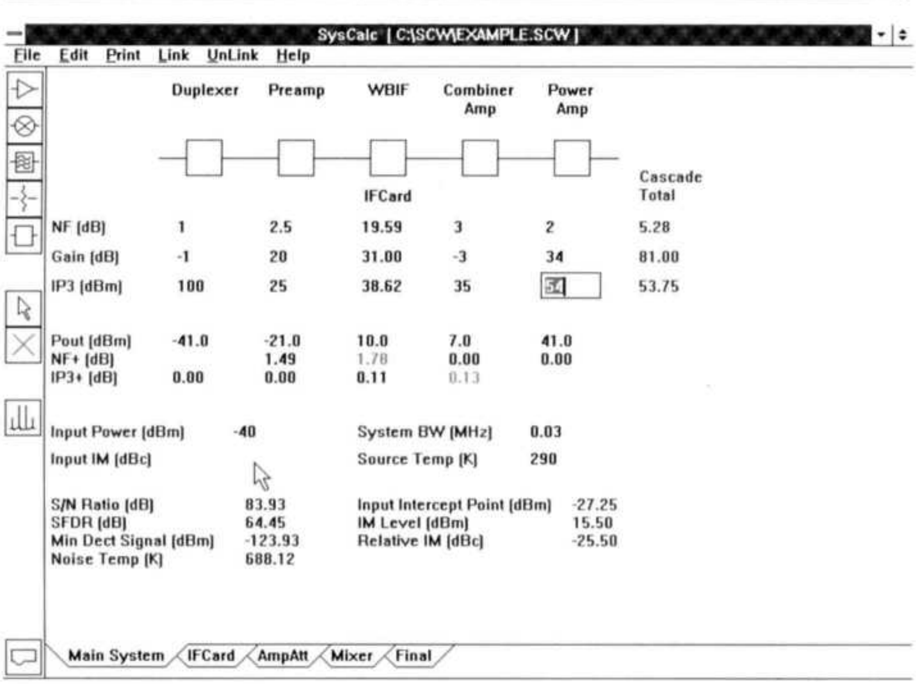


Fig 1—*SysCalc* 1.06 onscreen.

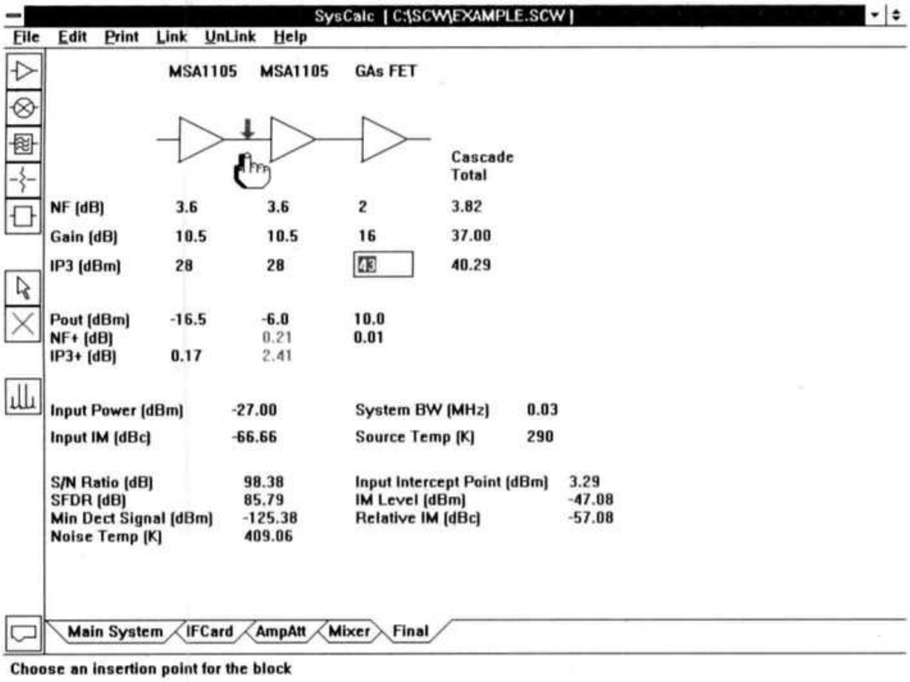


Fig 2—Releasing the mouse button at this point will drop in a mixer block.

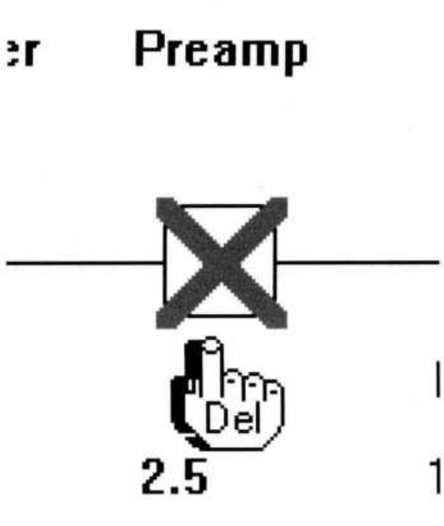


Fig 3—*X* marks the preamp for deletion in this view of part of the *SysCalc* screen.

the example file EXAMPLE.SCW produces the screen shown in Fig 1.

Rather than using the Multi-Document Interface (MDI) common to many *Windows* applications, *SysCalc* uses a tabbed-notebook scheme. You can drag and drop system blocks into place (Fig 2) or remove them (Fig 3) with a few mouse moves; you can add and delete sheets (see those tabs at the bottom of Fig 1); and you can merge multiple .SCW files into one.

SysCalc also can link a block or blocks on one sheet to another sheet; linked blocks are shaded yellow. Fig 1 shows the central (IF Card) block shaded—it's linked to the IF Card sheet. Double-clicking on a linked block switches you to the sheet it's linked to. (I miss being able to "unclick" to pop back to a sheet I just jumped from, but that's not allowed—you must click on the folder tab that corresponds to the original sheet.) No, you can't link the second box back to the first; trying to do so gives you a "circular link" error message.

SysCalc gives you two printing options: Notebook or Presentation output. The main difference between them is that Presentation generates a somewhat larger product you can use for producing transparencies. A limited choice of font sizes is available.

Nit-Picking It

If the system-planning concepts behind *SysCalc*'s crisp and intuitive presentation are cut and dried, can we have any bones to pick about the numbers it crunches and the numbers it returns after crunching? Yes, in the sense that *SysCalc* limits its IMD deliberations to cases of *third-order* IM, its intercept values to *output* intercepts, and its system impedance to 50Ω.

The commercial/military radio world is quite interested in *second-order* IMD; radio amateurs are catching on to its importance as well. Yet *SysCalc* 1.06a doesn't handle second-order IMD calculations.

SysCalc 1.06a includes no means of dealing with the effects of unequal test tones, so it can't simulate the effect of test tones falling on filter skirts.

SysCalc 1.06a can't think in terms of *input* intercept, but designers sometimes need to do so.

SysCalc 1.06a includes no means of system optimization. *You* provide those smarts!

SysCalc 1.06a doesn't speak the dual units (μV *and* dBm) commonly seen side-by-side in commercial, military and Amateur Radio specifications.

SysCalc doesn't tell you whether the power levels used in its calculations are average, peak, PEP or RMS. Inquiring minds need to know—and, sometimes, be able to choose among those options.

Modeling issues aside, *SysCalc* 1.06a's video smarts could be better. In one installation, *SysCalc*'s installation routine generated an error message because it didn't like the computer's Super VGA; merely ignoring the message nonetheless resulted in a *SysCalc* installation that played just fine. In two other installations, *SysCalc* 1.06a displayed its workspace text just fine in 640×480 resolution on a Gateway 2000 386/25 and squashed vertically in 640×480 on a Gateway 2000 4DX-33. (I think 800×600 resolution is the best thing to try if you have trouble at 640×480, because in 1024×768, wide *SysCalc* system sheets can extend beyond the screen borders—and no scroll bars appear to let you move there!) So *SysCalc* needs some work in the display department—font choices would be a start—but you can probably find a video resolution that plays.

Overall, *SysCalc* 1.06a is a slick, friendly CAE tool that frees you from paperwork and calculator-poking monotony so you can concentrate on the more important decision-making aspects of radio system planning and brainstorming.

I thank Dr. Ulrich L. Rohde, KA2WEU, for contributing to this review. □□

RF

by Zack Lau, KH6CP/1

A 2-W 13-cm Amplifier

This amplifier uses a resistively stabilized Hewlett Packard ATF-44101/AT-8140 power GaAs FET running class A to provide 14 dB of gain and a 2-W output. It wasn't too difficult to choose the device—there aren't that many linear devices that cover the 2.3 GHz band at this power level. Unfortunately, it is fairly pricey; the circa 1990 price was around $90. But since this transistor has been around a long time, it may be available surplus. Several hundred were sold at the bargain price of $3 each about a year and a half ago. I bought a few—I should have bought more.

Even if you do buy your FETs at bargain prices, you still will want to ensure that they are biased safely. As has been pointed out many times in the literature, having the negative gate bias supply fail while the drain supply is applied may result in the destruction of the device. Thus, designers have devised elaborate protection schemes to shut off the drain supply if the gate supply fails.

Fig 1 shows a simpler power-supply circuit. The drain supply is controlled by an ordinary 723 voltage regulator with current limiting. The current limiting protects the device—instead of the expensive FET getting hot and self-destructing, the cheap TIP 30 power transistor gets warm. Doing this eliminates the need for complex shutdown circuits to handle a failed bias supply. While there are more modern regulator chips than the 723, they aren't significantly better for this application. In fact, some are "improved" to the point where they are tougher to use. For example, chips with better current-sensing circuits often require special low-value resistors. While you could use the copper wire table in the *ARRL Handbook* to make your own low-value resistors

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Table 1—Measured Amplifier Performance

Input to MGF 1801 driver (dBm)	*Driver Output (dBm)*	*AT 8140 Output (dBm)*	*AT 8140 Gain (dB)*
-7.0	5.7	19.8	14.1
0.0	12.5	26.6	14.1
3.0	15.5	29.7	14.2
4.0	16.5	30.8	14.3
5.0	17.5	31.7	14.2
6.0	18.4	32.6	14.2
7.0	19.2	33.3	14.1
8.0	20.3	33.9	13.6
9.0	21.2	34.1	13.1
10.0	22.1	34.6	12.5
13.0	24.2	34.9	10.7

The 1-dB compression point is at 34.1/33.2 dBm output.
Comparing measurements made using an HP8563E spectrum analyzer against those made with an HP 453B/8481A power meter, a signal that measures 4.7 dBm on the spectrum analyzer reads out as 3.8 dBm on the power meter. The values listed are from the spectrum analyzer, taking into account a 30.2-dB, 25-W Bird attenuator inserted between the amplifier and the analyzer.

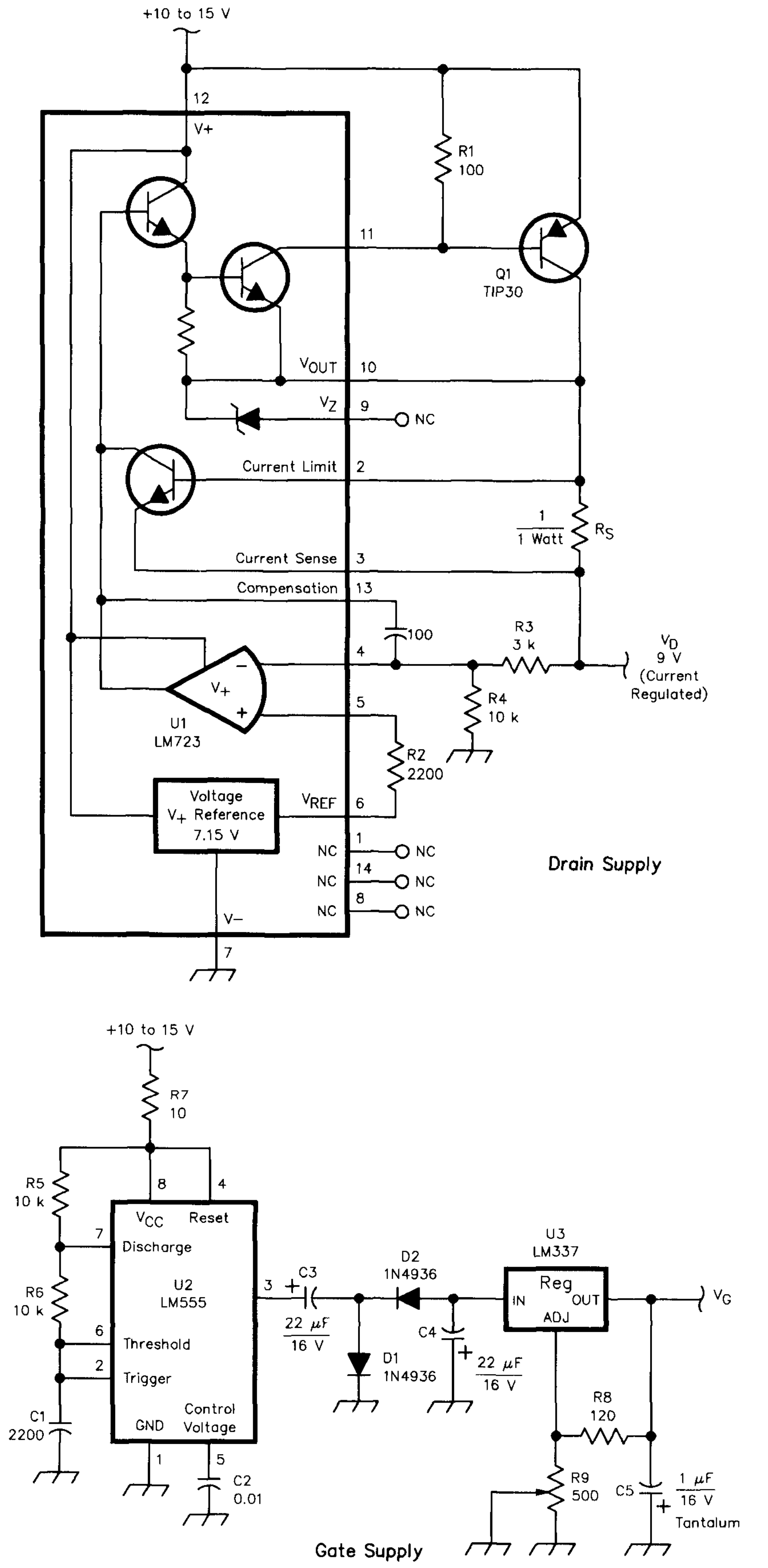

Fig 1—Power-supply circuit for the 2-W, 13-cm amplifier.

from wire, it makes more sense to me to use a 723 with standard, readily available parts.

The gate bias supply is produced by an NE555 timer running as an oscillator and driving a voltage inverter. An LM337 adjustable 3-terminal voltage regulator controls the gate supply. Now, you have to be a little careful using adjustable voltage regulators with bias supplies. While I haven't experienced this myself, it is entirely possible that a noisy potentiometer could momentarily present an open circuit between the wiper and the resistance element. Conceivably, in some circuits, this could result in voltage spikes at the output of the voltage regulator as the voltage is adjusted. The cure for this is to wire the potentiometer so that it never presents an open circuit, even with an intermittent wiper, as I've done with R9 in Fig 1.

For some other GaAs FET circuits, it may be advantageous to use an active-bias supply which operates inside the feedback loop of the current limited supply. This allows the bias point to be unaffected by temperature variations. The current limiting acts as a failsafe, protecting the transistor if a more negative gate supply voltage is unable to turn off the transistor. I've successfully used this circuit when only a milliamp of bias current was needed. Due to the high current required by the gate circuit of the 2-W amplifier, I decided it was impractical to active-bias the FET in this application; the ICL7660 can't provide the needed current. But there are higher-current versions of this chip, for those who don't mind the price.

RF Design

I found the AT-8140 rather difficult to get unconditionally stable. I ended up placing a 100-Ω chip resistor from gate to ground, which resulted in pretty good stability, except around 2.3 GHz. Unfortunately, it doesn't appear possible to dc decouple this resistor without adversely affecting stability, so quite a bit of bias current is needed. Because of this, it is conceivable that you may find a FET that needs too much gate voltage for a typical 100-Ω chip resistor to handle. Whether this is the case with a particular FET depends on both the threshold voltage and the transcon-ductance of the device. The threshold voltage can vary quite a bit, particularly in older devices, though the manufacturers have tightened up this specification in recent data books.

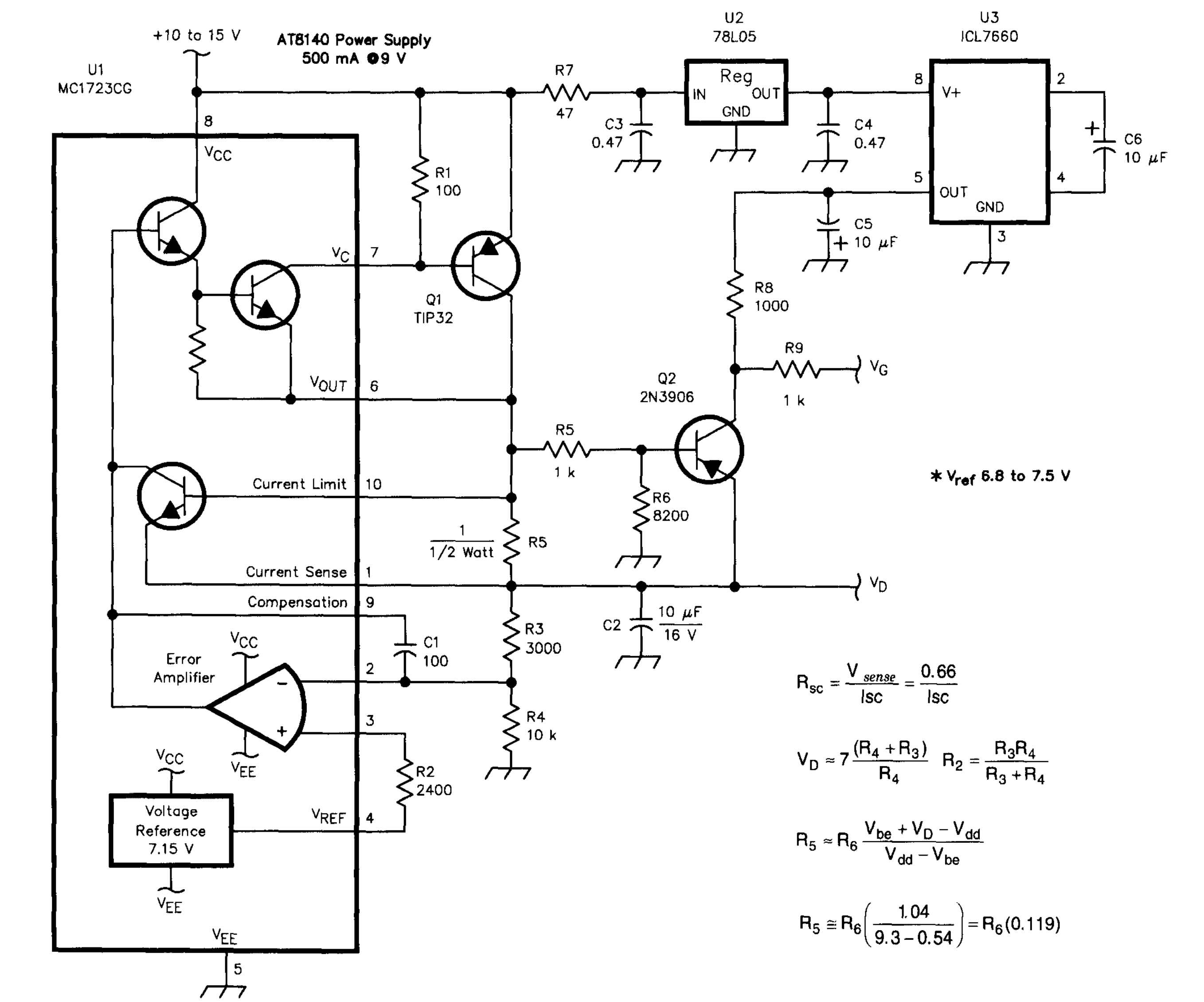

Fig 2—An active biasing circuit suitable for use with amplifiers that draw minimal gate current. Do *not* use this with the 2-W amplifier.

Worst case would be a large negative threshold voltage and a high transconductance. Stacking chip resistors in parallel might be a solution, though this might also invite unwanted parallel resonance effects, similar to what is seen with paralleled capacitors.

While in-band stability is not too critical, since the source and load impedance matching is usually pretty good at the operating frequency, I used a series resistor in the gate decoupling line to make the amplifier unconditionally stable, according to the computer model. This series resistor also improved the input match bandwidth, an important consideration if you don't want to tune the amplifier. Of course, improving the bandwidth with resistance also reduces the gain. The loss isn't too bad, however, seeing as this design is within a few dB of the maximum stable gain the data sheet specifies. The design was optimized with 150 Ω of series resistance, but I decided that such a high value was unwise. To accommodate such a high series resistance, it is entirely likely that the negative supply voltage would have to be more negative than the V_{GS} limit. Then, if the 100-Ω resistor were to fail open-circuited, the supply could damage the gate of the transistor. For this reason, I chose to use a 50-Ω series resistor. I didn't reoptimize the design for 50 Ω of series resistance, since the likely improvement was small compared to the variations I've seen when optimizing the unit on the bench.

Finally, I discovered that additional gate circuit bypassing was required at audio frequencies to prevent the circuit from oscillating. This was done with a 1-µF capacitor that is not shown in the computer model. This frequency range wasn't covered by the model.

Several dB of additional saturated output power was obtained by modify-

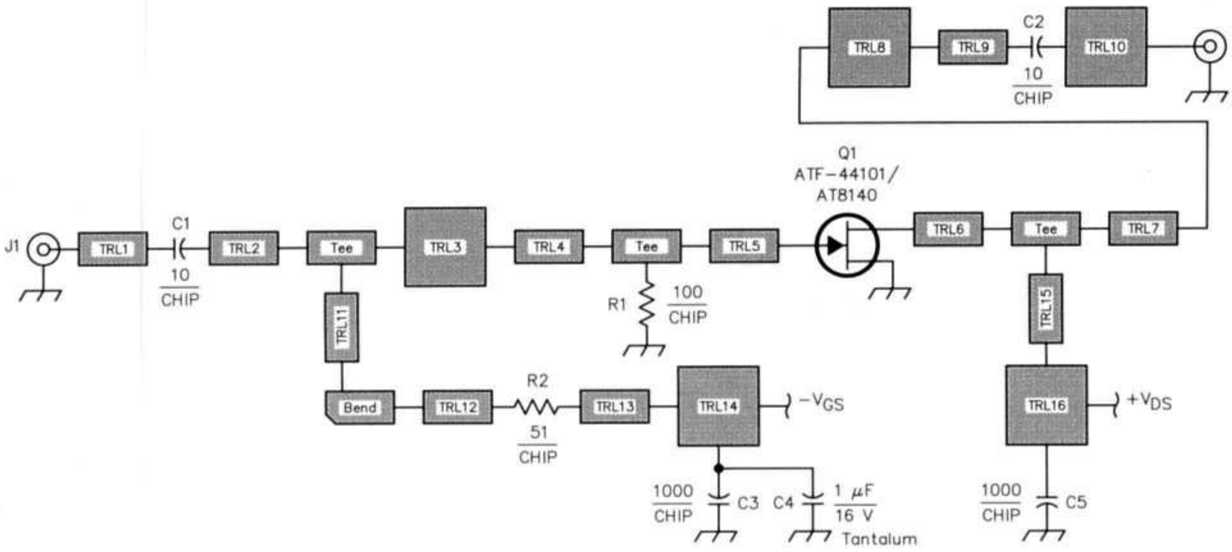

Fig 3—Schematic diagram for the 2-W amplifier.

C1, C2—10-pF, high-quality porcelain chip capacitors. I used 55-mil capacitors, although 100-mil capacitors should work just fine.
C3, C5—1000-pF NP0 chip capacitors
C4—1-µF, 16-V tantalum capacitor.
J1,J2—SMA jacks.
Q1—Hewlett Packard ATF-44101/AT-8410 medium power GaAs FET. This device must be properly heat-sinked. The absolute maximum channel temperature is 175°C. The case-to-junction thermal resistance is 23°C/W.
TRL1-16—Microstriplines etched on 0.031-inch-thick, ε_r=2.55 Teflon circuit board.

ing the output network with foil tabs. AT-8140s saturate at about 2.2 W output with around 200 mW of drive. An advantage to optimizing the circuit by modifying the 50-Ω striplines at the input and output of the amplifier is that the stability should not be adversely affected. But the board does have to be made larger to accommodate such tuning.

The original design was optimized using *Microwave Harmonica,* with the results shown in Table 2 and Fig 6. After I tweaked the output network to get maximum output power, I entered the changed circuit components into the computer and analyzed the circuit, resulting in the analysis shown in Table 3 and Fig 7. Table 1 shows the performance of the circuit in its final configuration.

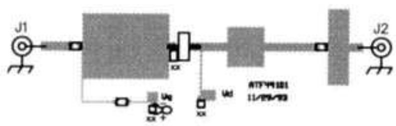

Fig 4—Parts placement diagram for the 2-W amplifier. Points labeled XX should be connected to the ground plane using copper foil.

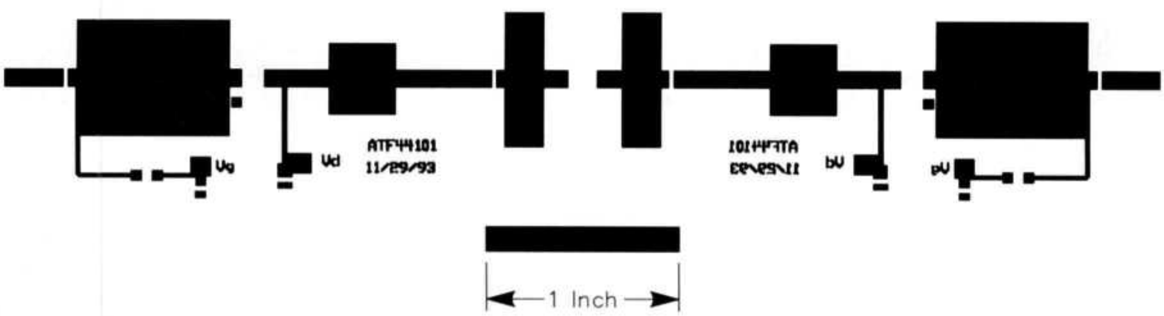

Fig 5—Etching pattern for the 2-W amplifier. Use 0.031-inch-thick, ε_r=2.55 Teflon circuit board.

Table 2—*Microwave Harmonica* Analysis (Original Circuit)

MICROWAVE HARMONICA PC V5.5 File: \mh\2304\qexamp.ckt 22-DEC-93 17:54:54

Freq GHz	MS11 dB AMP	PS11 deg AMP	MS21 mag AMP	PS21 deg AMP	MS12 mag AMP	PS12 deg AMP	MS22 dB AMP	PS22 deg AMP	MS21 dB AMP	K AMP
1.000	-0.70	-167.5	2.084	19.3	0.017	-53.7	-6.01	-128.7	6.38	1.64
1.100	-0.57	-177.2	1.988	0.1	0.018	-71.2	-5.35	-139.0	5.97	1.43
1.200	-0.47	173.3	1.941	-17.8	0.020	-87.3	-5.02	-148.2	5.76	1.27
1.300	-0.40	164.0	1.945	-35.0	0.022	-102.6	-4.96	-156.6	5.78	1.17
1.400	-0.36	154.7	1.999	-52.0	0.024	-117.7	-5.13	-164.3	6.01	1.09
1.500	-0.38	145.2	2.110	-69.2	0.028	-132.9	-5.59	-171.4	6.48	1.05
1.600	-0.48	135.2	2.288	-87.3	0.032	-148.9	-6.43	-177.6	7.19	1.02
1.700	-0.73	124.6	2.542	-106.8	0.039	-166.3	-7.84	178.8	8.10	1.01
1.800	-1.26	113.0	2.872	-128.6	0.047	174.0	-9.91	-177.3	9.16	1.01
1.900	-2.31	101.1	3.240	-153.5	0.056	151.3	-11.05	-156.6	10.21	1.01
2.000	-4.14	90.9	3.547	178.8	0.064	125.8	-8.31	-138.3	11.00	1.02
2.100	-6.75	87.5	3.682	149.7	0.070	98.9	-5.06	-141.0	11.32	1.04
2.200	-9.11	95.3	3.655	121.2	0.073	72.6	-3.15	-153.0	11.26	1.06
2.300	-10.01	108.4	3.587	93.9	0.075	47.6	-2.42	-167.3	11.09	1.07
2.304	-10.02	108.9	3.585	92.8	0.075	46.6	-2.41	-167.9	11.09	1.08
2.400	-10.08	120.9	3.583	66.5	0.078	22.4	-2.72	177.8	11.08	1.09
2.500	-9.15	137.4	3.638	36.0	0.082	-5.9	-4.55	162.9	11.22	1.11
3.000	-0.30	98.5	0.945	-114.5	0.025	-146.5	-1.26	-164.2	-0.49	1.17
4.000	-0.74	28.9	0.159	140.2	0.006	122.2	-0.11	150.2	-15.96	1.91
5.000	-2.01	-15.1	0.071	59.4	0.003	56.4	-0.08	123.6	-22.98	13.43
6.000	-0.78	1.8	0.044	120.4	0.003	125.4	-0.07	100.6	-27.20	9.68
7.000	-0.36	-60.0	0.158	68.8	0.014	86.8	-2.71	123.5	-16.02	8.12
8.000	-0.77	-109.6	0.086	-76.3	0.010	-50.3	-0.74	73.6	-21.26	14.53
9.000	-3.99	178.3	0.030	151.2	0.004	-174.8	-0.10	53.3	-30.56	50.93
10.000	-1.70	-138.8	0.016	5.4	0.003	45.4	-0.10	33.1	-36.06	75.19
11.000	-0.67	163.8	0.024	-91.2	0.006	-46.2	-0.47	12.8	-32.44	53.16
12.000	-0.95	116.5	0.031	106.0	0.010	152.0	-0.48	-0.6	-30.15	33.01

Table 3—*Microwave Harmonica* Analysis (Optimized for Output Power)

MICROWAVE HARMONICA PC V5.5 File: b:qexamp.ckt 22-DEC-93 18:04:28

Freq GHz	MS11 dB AMP	PS11 deg AMP	MS21 mag AMP	PS21 deg AMP	MS12 mag AMP	PS12 deg AMP	MS22 dB AMP	PS22 deg AMP	MS21 dB AMP	K AMP
1.000	-0.72	-168.6	2.386	39.7	0.020	-33.3	-17.28	-86.0	7.55	1.63
1.100	-0.67	-178.3	2.331	20.4	0.022	-50.9	-15.61	-117.3	7.35	1.42
1.200	-0.64	172.6	2.299	2.0	0.023	-67.5	-14.05	-142.9	7.23	1.27
1.300	-0.64	163.7	2.294	-15.7	0.026	-83.3	-12.69	-165.0	7.21	1.17
1.400	-0.66	155.0	2.314	-32.9	0.028	-98.6	-11.48	175.0	7.29	1.09
1.500	-0.71	146.5	2.365	-49.9	0.031	-113.6	-10.42	156.4	7.48	1.05
1.600	-0.79	137.8	2.453	-66.8	0.035	-128.4	-9.50	138.6	7.79	1.02
1.700	-0.91	128.9	2.588	-83.8	0.039	-143.3	-8.71	121.3	8.26	1.01
1.800	-1.10	119.4	2.786	-101.2	0.045	-158.5	-8.08	103.9	8.90	1.01
1.900	-1.42	108.8	3.071	-119.4	0.053	-174.6	-7.62	85.7	9.75	1.01
2.000	-1.98	96.3	3.482	-139.1	0.063	167.9	-7.46	65.7	10.84	1.02
2.100	-3.10	80.3	4.065	-161.7	0.078	147.5	-7.86	41.8	12.18	1.04
2.200	-5.77	57.1	4.815	170.7	0.096	122.2	-9.64	9.1	13.65	1.05
2.300	-14.58	10.6	5.418	135.8	0.113	89.5	-15.23	-52.9	14.68	1.07
2.304	-15.31	6.6	5.428	134.2	0.113	88.0	-15.55	-57.2	14.69	1.07
2.400	-10.67	-157.1	5.138	96.3	0.111	52.3	-13.32	174.5	14.22	1.09
2.500	-4.22	165.0	4.086	61.5	0.092	19.6	-7.41	119.7	12.23	1.11
3.000	-0.53	96.9	1.222	-35.2	0.032	-67.2	-2.33	21.6	1.74	1.17
4.000	-0.75	29.2	0.494	-155.6	0.019	-173.6	-1.06	-79.0	-6.13	1.69
5.000	-1.99	-15.1	0.261	93.6	0.013	90.6	-1.04	-169.9	-11.67	11.23
6.000	-0.78	1.9	0.137	130.1	0.010	135.1	-0.55	116.4	-17.28	7.89
7.000	-0.39	-60.0	0.101	160.7	0.009	178.7	-0.92	-3.1	-19.91	8.22
8.000	-0.75	-109.9	0.196	-1.0	0.023	25.0	-6.65	-66.0	-14.17	14.18
9.000	-4.01	178.4	0.126	-132.4	0.018	-98.4	-1.60	-147.6	-18.01	39.98
10.000	-1.70	-138.8	0.076	53.9	0.015	93.9	-2.18	130.1	-22.42	57.57
11.000	-0.66	163.8	0.071	-60.9	0.017	-15.9	-9.01	29.7	-22.92	50.28
12.000	-0.97	116.5	0.092	168.5	0.030	-145.5	-8.83	160.9	-20.71	31.18

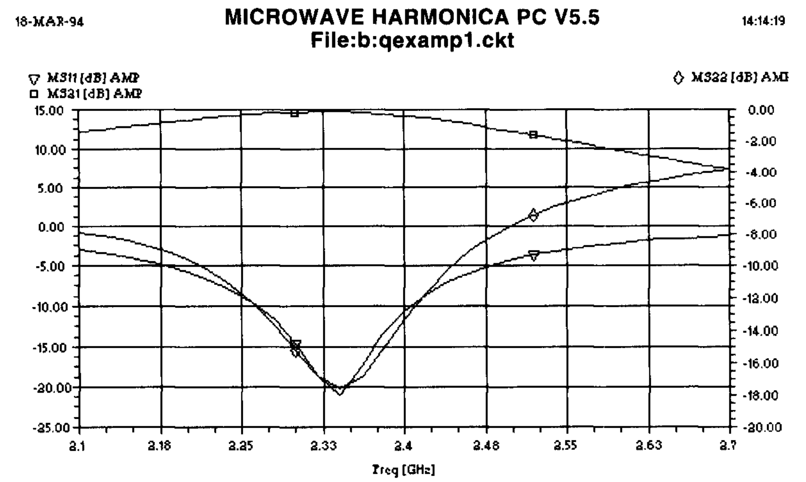

Fig 6—*Microwave Harmonica* analysis of the circuit's S parameters after computer optimization.

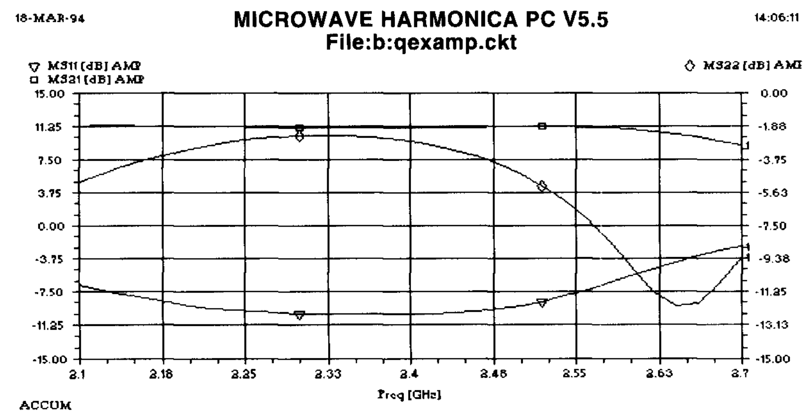

Fig 7—After the output network was optimized on the bench for maximum output power, *Microwave Harmonica* produced this analysis of the resulting circuit.

Construction Notes

To ground the drain bypass capacitor and the gate resistor, I cut slots in the board and connected the pads to the ground plane with 1-mil copper foil. Copper foil was also used to connect the gate bypass capacitor, although in this case I trimmed the board so that it wasn't necessary to cut a slot.

For high-quality grounding without lots of tiny screws, I decided to try using a heat spreader made out of 1-mil copper foil. I cut a hole in the circuit board for the FET and then carefully soldered the foil across the hole in the circuit board, taking care not to put any solder where it would interfere with heat-sinking action. I didn't want any air pockets, as they are extremely poor conductors of heat. Then, after attaching the board to a piece of 0.25-inch sheet aluminum with suitably tapped holes, I used a scribe to punch holes for the 0-80 screws to mount the FET. Ideally, clean holes would be punched into the copper, but I couldn't figure out a way to do that with the tools I have available. I used 0-80 screws, which are the largest screws I've found that will fit in the mounting holes. While I've used smaller screws in the past, I prefer using the largest screws possible since taps generally get tougher to use as they get smaller. I built one amplifier using 0.125-inch thick 6061 aluminum, but found that to be too thin to easily make the tapped holes for attaching the connectors and brass strips, although the finished mounting plate was stiff enough. □□