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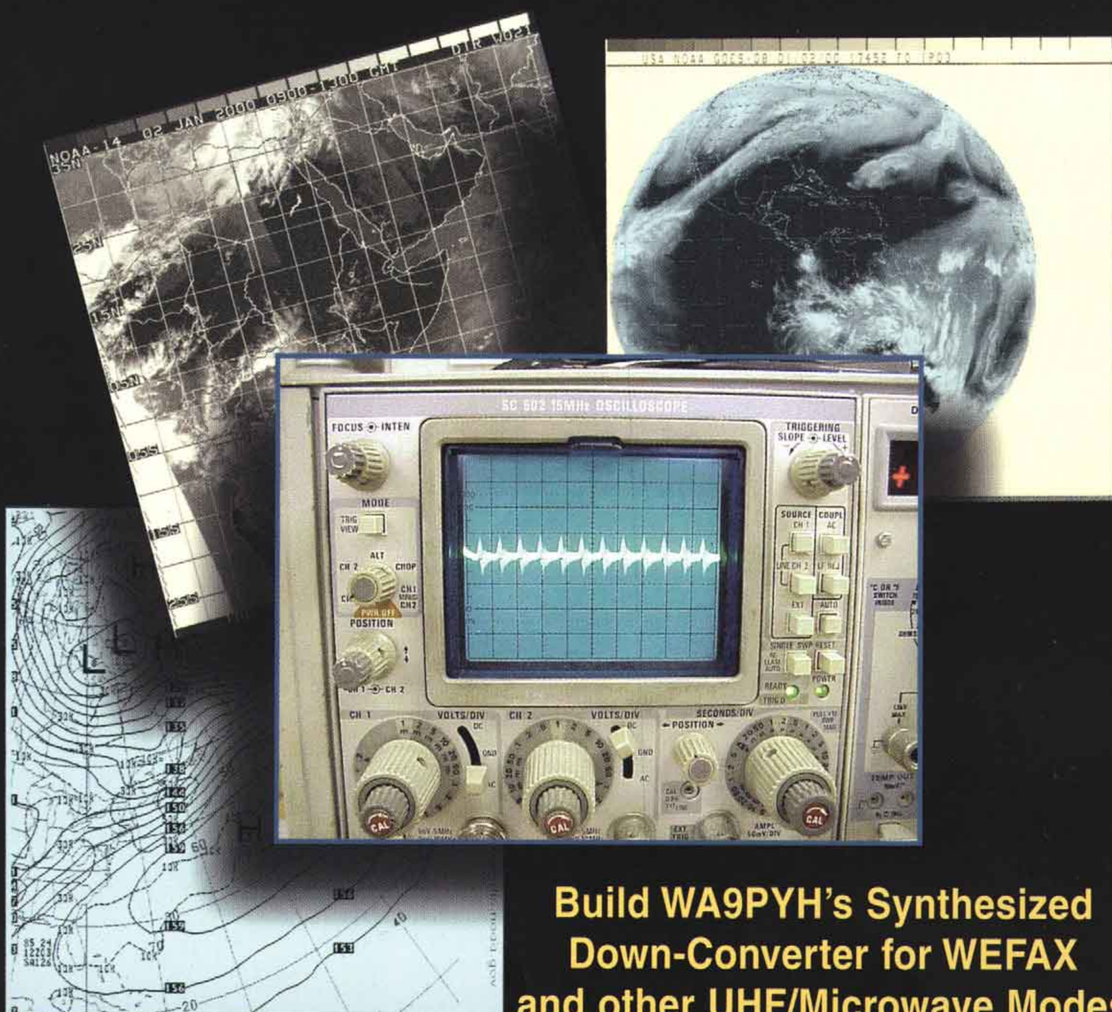

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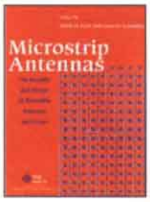

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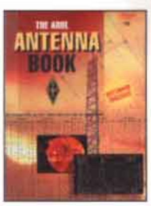

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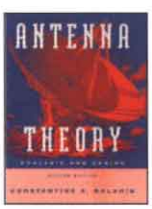

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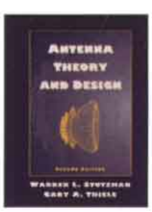

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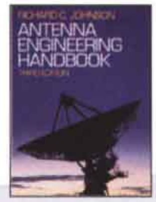

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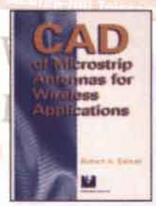

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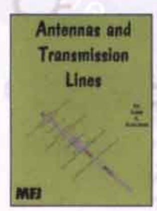

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QEX

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QEX (ISSN: 0886-8093) is published bimonthly in January, March, May, July, September, and November by the American Radio Relay League, 225 Main Street, Newington CT 06111-1494. Yearly subscription rate to ARRL members is $22; nonmembers $34. Other rates are listed below. Periodicals postage paid at Hartford, CT and at additional mailing offices.

POSTMASTER: Form 3579 requested. Send address changes to: QEX, 225 Main St, Newington, CT 06111-1494

Issue No 199

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Telex: 650215-5052 MCI
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Subscription rate for 6 issues:

In the US: ARRL Member $22, nonmember $34;

US, Canada and Mexico by First Class Mail: ARRL member $35, nonmember $47;

Elsewhere by Surface Mail (4-8 week delivery): ARRL member $27, nonmember $39;

Elsewhere by Airmail: ARRL member $55, nonmember $67.

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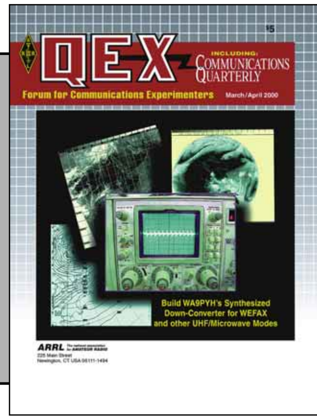

About the Cover

Sample output images and a test shot (phase/frequency-detector output) of the inexpensive WEFAX down-converter.

Features

3 The ATR-2000: A Homemade, High-Performance HF Transceiver, Pt 1
By John B. Stephensen, KD6OZH

16 A Homebrew Multiband Yagi
By Johan Van de Velde, ON4ANT

20 A Giant LF Loopstick
By Richard Q. Marris, G2BZQ

23 A High-Performance Homebrew Transceiver: Part 5
By Mark Mandelkern, K5AM

38 High-Quality SSB Audio
By Bob Heil, K9EID

45 A Class-B Audio Amplifier
By Parker R. Cope, W2GOM/7

48 A Synthesized Down-Converter for 1691-MHz WEFAX
By Jim Kocsis, WA9PYH

Columns

56 RF *By Zack Lau, W1VT*

59 Upcoming Technical Conferences

60 Letters to the Editor

63 Next Issue in *QEX*

Mar/Apr 2000 *QEX* Advertising Index

Almost All Digital Electronics: 64
Astron: Cov IV
Atomic Time, Inc.: 55
Communications Concepts Inc.: 15
Communications Specialists, Inc: 15
Computer Aided Technologies: 63
HAL Communications: Cov III
Roy Lewallen, W7EL: 44
Nemal Electronics International, Inc.: 44
Noble Publishing: Cov II
No Walls Inc: 37
Renaissance Radio: 47
Shoc: 62
Tucson Amateur Packet Radio Corp: 63
TX RX Systems Inc.: 55
Universal Radio, Inc.: 44
Z Domain Technologies, Inc: 37

The purpose of *QEX* is to:

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

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

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted on IBM or Mac format 3.5-inch diskette in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX*. Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Empirically Speaking

Welcome to the combined *QEX/Communications Quarterly*. Special greetings go to those of you who haven't been getting *QEX*. As a Forum for Communications Experimenters, you will find this magazine to be a level medium for the exchange of ideas and information about a wide variety of topics. We are here to help document advanced work in telecommunications and support efforts to forward the state of the art. We also embrace the cause of technical education and try to solicit and select a mixture of articles encompassing a broad range of skill levels. Here we are much less formal than *QST*, for example. Authors need not explain every detail, nor provide complete bills of materials; they exchange ideas through their articles and letters. I think this isn't so different from what you've come to expect.

To *Communications Quarterly* contributors, whom we're also quite happy to have aboard: Most of your articles will be scheduled and run over the next several issues; a couple are still pending queries from me as of this writing. We compensate authors at the rate of $50 per published page or part thereof, on publication. We would be delighted to receive your further submissions at the address listed on page 1 and look forward to hearing from you. Acceptance decisions are made by me with the able assistance of the ARRL Technical Advisory (TA) staff. Most of these volunteers are top professionals as well as radio amateurs. Our editorial criteria are well documented on our Web page (**www.arrl.org/qex/**) and are available through regular mail for the asking.

We also welcome your letters, inquiries, comments and suggestions. Civil discussion is always well received. Long-time readers know that this publication may be whatever you want to make it. As Dick Ross, K2MGA, put it, at least now it has reached the "critical mass" it needs to flourish. I'm happy to be a part of it, and I look forward to serving you in the future.

In this QEX/Communications Quarterly

We begin with a look into John Stephensen, KD6OZH's home-brew rig, the ATR-2000. This synthesized, 20-meter monobander uses a common-sense approach to reasonably large receiver dynamic range on HF. John's circuits address the needs for better phase-noise performance and reduction of receiver-induced QRM.

Johan Van de Velde, ON4ANT, has produced some interlaced, multiband Yagis that ought to interest DXers and anyone else who wants a strong signal. The tradeoffs among element spacing, boom length and electrical parameters are duly considered in Johan's computer modeling. Richard Marris, G2BZQ, is working on a somewhat smaller scale; he brings us his "giant" loopstick antenna. It's good for broadcast and other LF reception where larger antennas aren't practical.

Mark Mandelkern, K5AM, wraps up his series with emphasis on his control system and the permeability-tuned oscillators (PTOs). Mark has also promised to document for us the design of the front ends—the parts that change from band to band.

Audio pioneer Bob Heil, K9EID, contributes his analysis of high-fidelity audio reproduction as it pertains to Amateur Radio. He discusses what hardware is necessary and how it should be adjusted for best results under a variety of conditions. Like his extremely popular seminars on the subject, this piece benefits from Bob's engaging style. Frequent *QEX* contributor Parker Cope, W2GOM/7 explains class-B transistor amplifiers and gives circuit examples for audio applications.

Jim Kocsis, WA9PYH, has been receiving high-resolution photographic telemetry (HRPT) WEFAX and other signals in the S band and contributes his down-converter design. According to Jim, the frequency may be moved upward or downward to accommodate nearby Amateur Radio bands, as well. In *RF*, Zack Lau, W1VT, discusses PA matching at 144 MHz and a QRP dipole center insulator—*73, Doug Smith, KF6DX, **kf6dx@arrl.org**.*

The ATR-2000: A Homemade, High-Performance HF Transceiver, Pt 1

Homebrew is alive! This interesting high-performance transceiver draws on traditional techniques and implements them with contemporary IC building blocks and control by an external PC.

By John B. Stephensen, KD6OZH

Six years ago, I completed a transmitter and receiver for the 2-meter Amateur Radio band that was more or less "state-of-the-art" for its time. The most-difficult parts of this project were the LOs, which required multiple PLLs, mixers and filters to achieve adequate resolution and stability. In the intervening time, direct digital synthesizers (DDSs) have become available in chip form, allowing the LO complexity to be reduced considerably. In addition, new gain-controlled and log amplifier chips have become available, which can provide improved AGC performance.

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Los Angeles, CA 90049
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Since my license privileges would soon extend below 28 MHz and I would have a lot of spare time for six months, I decided to try some new ideas.[1] I would build a transceiver for the 20-meter amateur band that could later be expanded to other bands. I wanted a high-performance transceiver optimized for use on the crowded HF bands with the following features:

- High dynamic range to allow working close to local stations
- Low-phase-noise LO to preserve the shape factor of CW and PSK31 filters
- At least 90 dB of image rejection
- An effective noise blanker to counter local line noise
- Effective speech processing
- Clean audio without the "mushiness" found in many off-the-shelf products
- An accurate S-meter
- 100 W PEP output or more
- Computer control
- Use of an external frequency standard

[1]Notes appear on page 15.

The dynamic range requirements initially seemed very severe at my Los Angeles location with two other Amateur Radio stations within a one-kilometer radius. I initially calculated the line-of-sight path loss and determined that a station running 1 kW with a three-element beam could generate +18 dBm (5 V, P-P) at the antenna terminals of the receiver. This seems to be mitigated, however, by the fact that amateur HF antennas are used at low heights (in wavelengths), so much of the power goes upward rather than travelling close to the ground.

The effective antenna-height restriction here is 45 feet, so the peak radiated power on 20 meters occurs at an elevation of 22° or more. I performed a test with one of these stations running 100 W on 28.45 MHz, where the elevation angle is lower. With his three-element beam pointed at me while my five-element beam was pointed at him, his signal was only –37 dBm (36 dB over S9) at the antenna, 45 dB below the predicted value. A kilowatt would have produced –27 dBm (46 dB over S9), so this is the maximum signal level that should be encountered at my location.

Atmospheric noise determines the required receiver noise figure. I assumed that I would always be using at least a full-size dipole for reception. Although I am located in a high-noise environment now, I picked a low-noise environment to determine receiver sensitivity. A low atmospheric noise level on 20 meters is 28 dB above thermal noise or about 0.6 µV at the receiver antenna terminals as shown in Fig 1. To avoid degrading the sensitivity of the system by more than 1 dB, the receiver noise figure should be at least 6 dB lower, or 22 dB above the thermal noise. On the 10-meter band the atmospheric noise is 10 dB lower so the receiver noise figure should be 12 dB or less.

The HF dynamic range requirement is greatest on 10-meters, as signals are strong and the noise level is low.[2] The maximum signal level is –27 dBm. If we want spurious responses to raise the received noise level by only 1 dB they must be 6 dB below the atmospheric noise level, which is –122 dBm in a 2.5-kHz bandwidth. The spurious-free dynamic range (SFDR) required is therefore 101 dB = –27 – –126 + 6, which requires the use of a very good mixer circuit.

The other component that affects the dynamic range is the local oscillator. Because of reciprocal mixing, the noise sidebands of the LO will appear on every signal received. The LO noise floor must be low enough so that—when mixed with the combined power of all signals arriving at the receiver—it does not raise the noise floor. We can confidently assume that the total power of all arriving signals will never exceed S9 + 70 dB or –3 dBm. Since the noise floor at the mixer input is –164 dBm/Hz, the LO noise floor must be less than –161 dBc/Hz, and this is easily achievable.

Reciprocal mixing also makes the filters in the receiver appear wider than they are, and the problem is most severe with narrow CW or PSK filters. The narrowest filter that I will use is a 250-Hz filter that is 1 kHz wide at the –80-dB points. To degrade the response by only 3 dB, the LO phase noise must be 80 dB down for a signal to appear at the –80 dB point on the passband. Since the noise over a 250-Hz range will appear in the receiver output, it should be less than –114 dBc/Hz at a 500-Hz offset. Phase noise usually decreases at a 9-dB/octave rate at low frequencies. This requires –134 dBc/Hz at a 5 kHz offset. At ±5 kHz, spurious responses from reciprocal mixing are then 110 dB below the interfering signal level in a 250-Hz bandwidth or –100 dB in a 2.5 kHz bandwidth. Reception will be unimpaired 5 kHz or more away from the strongest possible CW, PSK and FSK signals if they are clean.[3] SSB signals will have splatter that is greater than –100 dBc so the receiver poses no limitation in that case.

I wanted to meet these requirements with a transceiver that was relatively simple to construct and not too costly. Before starting the design, I reviewed numerous articles in *QST* and *QEX* going back into the 1950s and 60s, when homebrew ham stations were more common. In fact, the basic architecture dates from the 1960s, but uses 1990s technologies to reduce IMD and increase frequency stability.

The first decision was to eliminate the possibility of general-coverage reception and to use a single-conversion architecture. This minimized complexity without sacrificing any performance in the amateur bands. Limiting the frequency range to amateur bands allows the use of front-end filters optimized to reject interference from high-power international broadcast stations on adjacent frequencies.

The single-conversion architecture

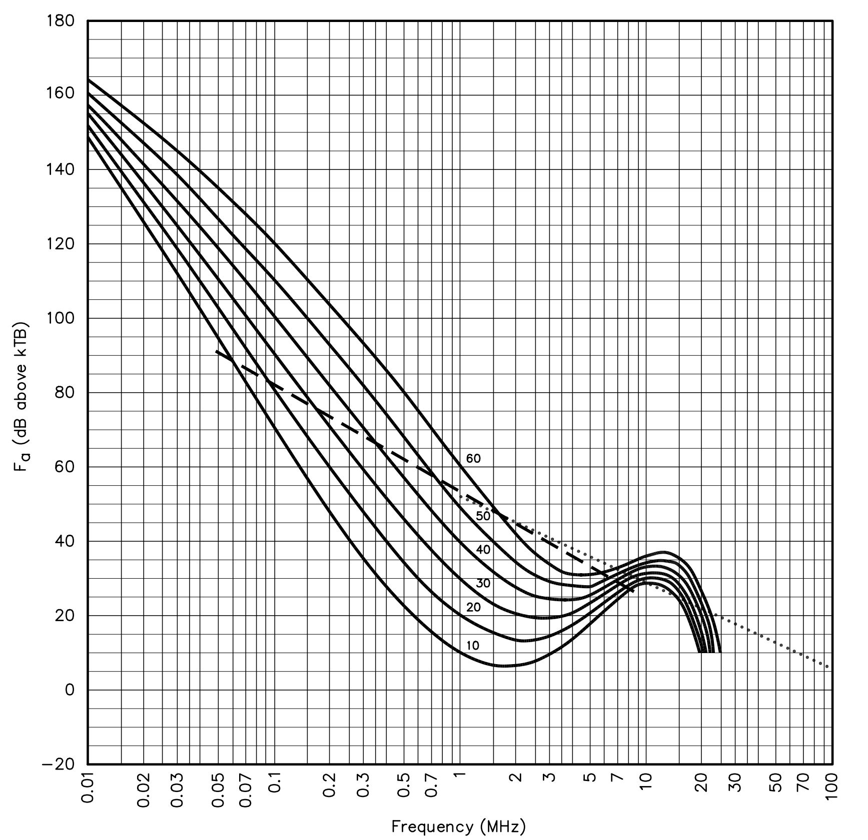

Fig 1—Atmospheric noise levels in a quiet location. Solid lines are the expected levels of atmospheric noise (in dBN_0) with curve 10 showing the quietest locations on Earth. The dashed line is the expected level of galactic noise and the dotted line is the expected level of manmade noise. The peak at about 13 MHz is due to ionospheric propagation of noise (source: CCIR Report 322).

was first described in *QST* by Goodman[4] and later by Squires,[5] it has the same advantages today as described then. The one disadvantage at the time was the complexity of building a stable HF LO. Consequently, double-conversion architectures with a crystal-controlled converter followed by a MF tunable IF became popular. This architecture predominated in a generation of commercial[6] and homebrew amateur equipment[7] alike from the 1960s through the 1980s. Although stable synthesized HF LOs became available in the 1980s, commercial transceiver designs have become even more complicated. They often have a VHF first IF and a wide-bandwidth VCO in the LO to allow general coverage, plus a third IF to provide reasonable gain distribution. This also created some well-known deficiencies[8, 9] that still exist in many products.[10] Single conversion eliminates two stages of mixing, thus reducing intermodulation distortion (IMD) and spurs. It also allows the use of narrow-band VCOs, which dramatically reduce phase noise, eliminating the mushy audio associated with many modern transceivers.

Today, it is possible to build a simple crystal-locked LO with a DDS chip and a PLL chip. The use of DDS chips has been described previously in *QST*;[11] it can vastly simplify the construction of homebrew transceivers. The output of a DDS should not be used directly as the LO, however, because its total spurious output power is only 40-70 dB down and spurs can be very close to the output frequency. These spurs will result in spurious responses in the receiver and spurious outputs from the transmitter.

The single-conversion architecture also requires that we distribute gain more equally between the IF and audio sections of the receiver. In the past, a problem has been the requirement to amplify as much as possible at the IF to be able to drive an AGC rectifier. The only alternative was AF-derived AGC with its pops and clicks. A better solution is the use of dual AGC loops[12] with low-level log detectors at both IF and audio.

I also decided to make this transceiver a "black box," controlled by software on a PC. I needed some of the "bells and whistles" of the commercial equipment, such as multiple memories, dual VFOs, IF shift and RIT. In the past, adding even minimal features vastly increased the complexity of the project—with multiple mechanically tuned VFOs, complex switching and mixing schemes. With notable exceptions,[13] very few seem to

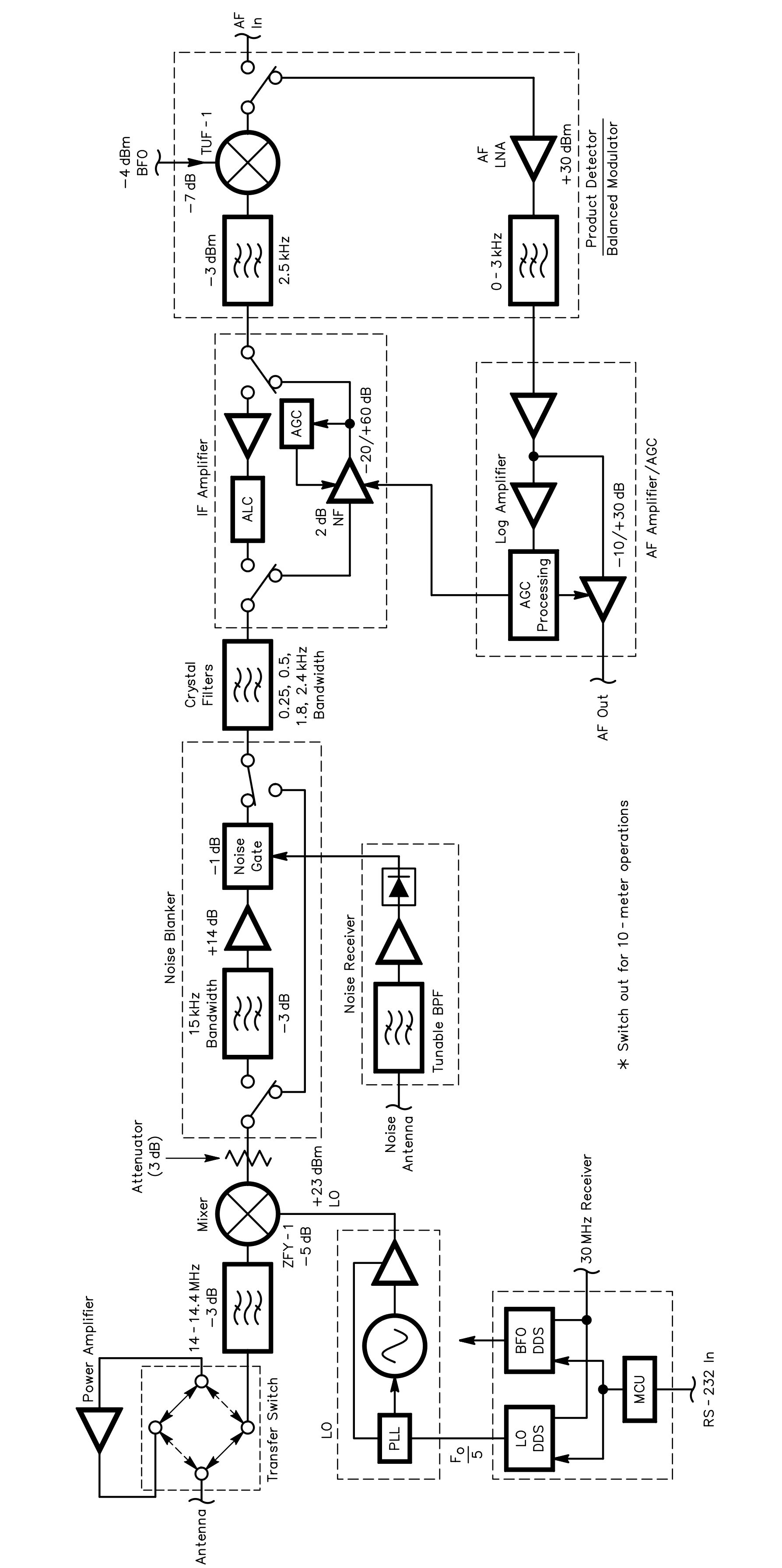

Fig 2—Transceiver block diagram.

have attempted such a project recently. However, software running on the PC easily controls this homebrew receiver through a serial port and provides these bells and whistles without adding hardware.

Block Diagram

The block diagram of the transceiver is shown in Fig 2. Dotted lines enclose the circuits built as individual shielded modules. The initial version is for 20 meters, but additional LO modules and RF filters can be added for other bands. The transceiver was designed with bidi-rectional circuits to allow sharing of the expensive components, including crystal filters and the high-level mixer.

The antenna is connected to a TR switch that bypasses the linear amplifier during reception. The linear amplifier would otherwise amplify receive signals to a level between about 0.025 mW and approximately 150 W PEP. Following the TR switch is a three-pole RF bandpass filter covering 13.9-14.5 MHz and a high-level balanced mixer. Note that no RF preamplifier is used since external noise sources dominate at 20 meters.

Driving the mixer is a LO that is indirectly phase-locked to an external 30-MHz reference that is tunable in 1-Hz increments. The LO is restricted to the range 22.95-23.45 MHz to cover the 20-meter amateur band (including the proposed extension to 14.4 MHz) with a 9-MHz IF. The LO is phase-locked to the output of a DDS running at one-fifth of the output frequency.

Two DDSs are used: one controls the LO and one provides the BFO directly. Accurate BFO control is needed to provide the IF-shift function. Both are controlled by a PIC that receives frequency commands via an RS-232 port. The external frequency reference is derived from either a surplus oven-controlled crystal oscillator (0.1-ppm accuracy) or a rubidium standard (0.001-ppm).

Following the mixer is a pre-IF noise blanker[14] that is bypassed on transmit. This starts with a 15-kHz-wide, two-pole crystal filter that delays noise pulses to be blanked and protects the post-mixer IF amplifier that follows from strong signals elsewhere on the band. The post-mixer IF amplifier reduces receiver noise figure overall and drives the noise gate, where noise blanking occurs.

The noise gate is driven by a separate receiver that amplifies and detects the noise pulses. This avoids noise-blanker triggering by strong, in-band signals from nearby amateur stations. This approach was originally used by Collins Radio[15] and more recently by Mandelkern.[16] It provides more effective blanking in an urban environment than other arrangements I have tried.

The main crystal filter follows the noise blanker. Its bandwidth is selectable for SSB (2.4 and 1.8 kHz), FSK (500 Hz) and CW (250 Hz). The filters are all 8- or 10-pole units providing 80-100 dB of stop-band attenuation, depending on the manufacturer.[17]

Next is a bidirectional IF amplifier. On receive, it provides 80 dB of gain control. Control comes from two sources. An internal envelope detector with a fast rise time suppresses noise impulses and other transients. Accurate long-term control is provided by a slower, AF-derived AGC loop. On transmit, amplification is provided with variable-time-constant ALC. A long time constant provides minimal distortion and a short time constant provides a clipping-like function to maximize intelligibility when signals are weak.[18]

A low-level DBM and five-pole crystal filter provide product detection on receive and SSB generation on transmit. From here, signals are split into separate transmit and receive paths. While receiving, low-noise audio amplification is provided by a PNP transistor[19] in a configuration identical to that used in direct-conversion receivers.[20] The signal is then filtered to remove high-frequency hiss. Zero-dBm trans-

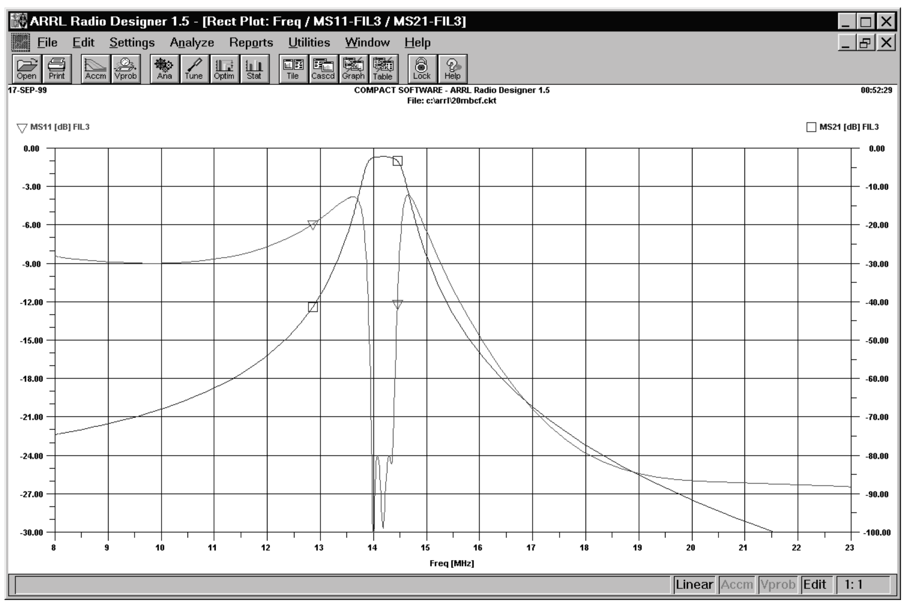

Fig 3—20-meter bandpass filter characteristics.

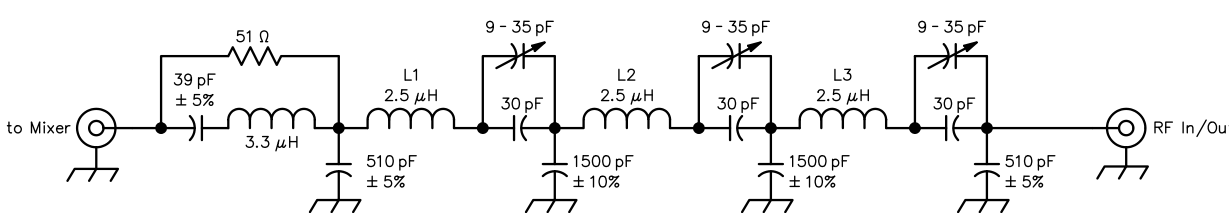

Fig 4—20-meter bandpass filter schematic diagram.
L1-L3—2.5 µH, 24 turns #22 AWG on T50-6 powdered iron core

mit audio is provided from an external microphone preamplifier and telephone coupler.

The AF output goes to a log amplifier and a variable-gain linear amplifier. The log amplifier is used as an AGC detector for the slow AGC loop. The slow-AGC control voltage is processed to provide optimal response for SSB and CW.[21] It has very low gain to minimize transient instability. The IF amplifier gain varies only 80 dB over a 120-dB signal range. The remaining 40-dB variation in signal strength is removed by controlling the gain of the last AF amplifier.

Housekeeping functions, such as TR switching and filter selection, are controlled by a second PIC that also communicates with the PC via a common 19,200-b/s EIA-232 link.

Circuit Details

Front-End Bandpass Filter and Mixer

The transceiver uses a high-side LO so that the image frequency is above the RF. Three poles provide adequate attenuation of the image frequency and other spurs from LO harmonics. Fig 3 shows the filter characteristics as modeled with *ARRL Radio Designer* and Fig 4 is the schematic diagram.

The bandpass filter is a Chebyshev with a 500-kHz bandwidth, 0.1 dB of ripple and shunt capacitive coupling between resonators. This type of filter was chosen because its pseudo-low-pass characteristic provides maximum attenuation at the image frequency with a high-side LO. Attenuation at the image frequency exceeds 100 dB. To achieve this level of attenuation, the filter must be constructed with a shield between each resonator. I used a small minibox with two copper partitions attached to one side of the box. Each 1500-pF capacitor actually consists of a 470-pF ±20% feedthrough capacitor penetrating the partition and a 1000-pF ±5% ceramic disc capacitor in parallel.

The first resonator on the mixer side of the filter, which is shunted by a 51-Ω resistor, is a diplexer. It shorts the resistor at frequencies near the passband. Above the passband, the resistor is shunted to ground by the 510-pF capacitor. At the LO frequency and above, the return loss of this port is greater than 20 dB. This diplexer dramatically reduces IMD in the mixer.

A Mini Circuits ZFY-1 mixer and a 3-dB attenuator follow the filter. Combined with the filter's loss, this level-23, doubly balanced diode mixer provides a 1-dB compression point of +23 dBm, which is 4 dB above the dynamic-range requirements. In fact, it is hard to do much better than this. At the 1-dB compression point, the signal level at the attenuator output is +10 dBm, or 10 mW, and the crystal filters are only rated to handle 10 mW of input power. IMD generated in the crystals would probably compromise the performance of a higher-level mixer.

The attenuator at the IF port of the mixer limits impedance excursions that degrade mixer IMD. The maximum SWR is 3:1, which should result in degradation within the 4-dB margin. A more complicated circuit could be used here with diplexers, hybrids and matched crystal filters to eliminate the attenuator, but this was deemed too expensive and complex. The 2-dB advantage in noise figure is not needed, as shown by the calculations below.

Given a 2-dB noise figure for the IF and 4-8 dB loss in the crystal filters, the estimated noise figure of the receiver is:

CW	SSB		
2	2	dB	IF amplifier noise figure
8	4	dB	8-pole CW/10-pole SSB filter loss
3	3	dB	Post-mixer attenuator
6	6	dB	Mixer loss
3	3	dB	20-meter bandpass filter loss
22	18	dB	Receiver noise figure

This is 6-10 dB below the level of 28 dB set by minimum atmospheric noise level,[22] so that the minimum discernable signal (MDS) detectable by the receiving system—including the antenna—is degraded by 1 dB or less:

CW	SSB		
–174	–174	dBm/Hz	Thermal noise at 290 K
+29	+28	dB	Receiver plus atmospheric noise figure
+24	+34	dB-Hz	Receiver bandwidth (2.4 kHz)
–121	–112	dBm	System MDS

The +35-dBm third-order intercept of the ZFY-1 mixer results in an intercept point of +38 dBm at the input to the 20-meter bandpass filter. The spurious-free dynamic range (SFDR) of the system is therefore estimated to be:

CW	SSB		
+38	+38	dBm	IP3
–(–121)	–(–112)	dBm	System MDS
159	150	dB	
× 2/3	× 2/3		
106	100	dB	System SFDR

This design should meet performance requirements without an RF amplifier or a post-mixer amplifier. On-the-air listening tests revealed that sensitivity is more than adequate and that the receiver is not "desensed" by stations running 1 kW on 20 meters just 1 km away.

On 10-meters where atmospheric noise is expected to be10 dB lower, the post-mixer attenuator must be removed and the 15-kHz filter/post-mixer amplifier must be switched in. This lowers the noise figure at the mixer input to 10 dB. Since I always have a TVI low pass filter connected to the transceiver, the RF bandpass filter requirements are minimal and a two-pole filter can be used. The insertion loss of the bandpass filter is 1.5 dB and the TVI filter is less than 0.5 dB, resulting in a system noise figure of 12 dB on SSB and CW. This is adequate for any receiving location. At my location, I have never seen atmospheric noise on 10 meters lower than +19 dBN$_0$ and it rises significantly when the band is open.

Local Oscillator

As described earlier, LO phase noise must be minimized to preserve the CW filter shape factor. The VCO was designed for minimum phase noise by narrowing the tuning range, maximizing the loaded Q of the resonator and operating it at the highest possible

```
1. L=  1.430uH     2. Q=   70       3. Cpad=  25.0pF     4. Cv=  60.0pF
5. Ev=     4.0V    6. Rs= 1.0       7. Fref=    36.0kHz  8. Fofs=     0.0MHz
9. Epd=    5.0V   10. t1= 200ms   11. NF=  6.0dB       12. Pi=  0.0 dBm
13. Fc= 20.00kHz 14. Cser=   10.0pF 15. Hyperabrupt: 0
t2=  84.12864ms                                      --------- L(Fm) dBc ---------
Vs(V) Fo(MHz) dF(Hz) Wn(r/s) damp. Tacq(s) BW(Hz)  50  500   2K  10K 100K   1M
  4.0   22.97    0.4    41.8  1.76  713.70    25  -75 -104 -122 -141 -163 -171
  6.0   23.05    0.3    37.3  1.57  893.62    21  -74 -104 -122 -141 -163 -171
  8.0   23.12    0.3    34.3  1.44 1059.85    18  -74 -104 -122 -141 -164 -171
 10.0   23.17    0.3    32.0  1.35 1217.18    16  -74 -104 -122 -141 -164 -171
 12.0   23.22    0.2    30.2  1.27 1368.22    14  -74 -104 -122 -141 -164 -171
 14.0   23.27    0.2    28.7  1.21 1514.55    13  -74 -104 -122 -141 -164 -171
 16.0   23.31    0.2    27.4  1.15 1657.22    12  -74 -104 -122 -141 -164 -171
 18.0   23.34    0.2    26.3  1.11 1796.96    11  -74 -104 -122 -141 -164 -171
 20.0   23.38    0.2    25.4  1.07 1934.32    10  -74 -104 -122 -141 -164 -171
 24.0   23.44    0.2    23.8  1.00 2203.40     9  -74 -104 -122 -141 -164 -171
```

Fig 5—VCO simulation.

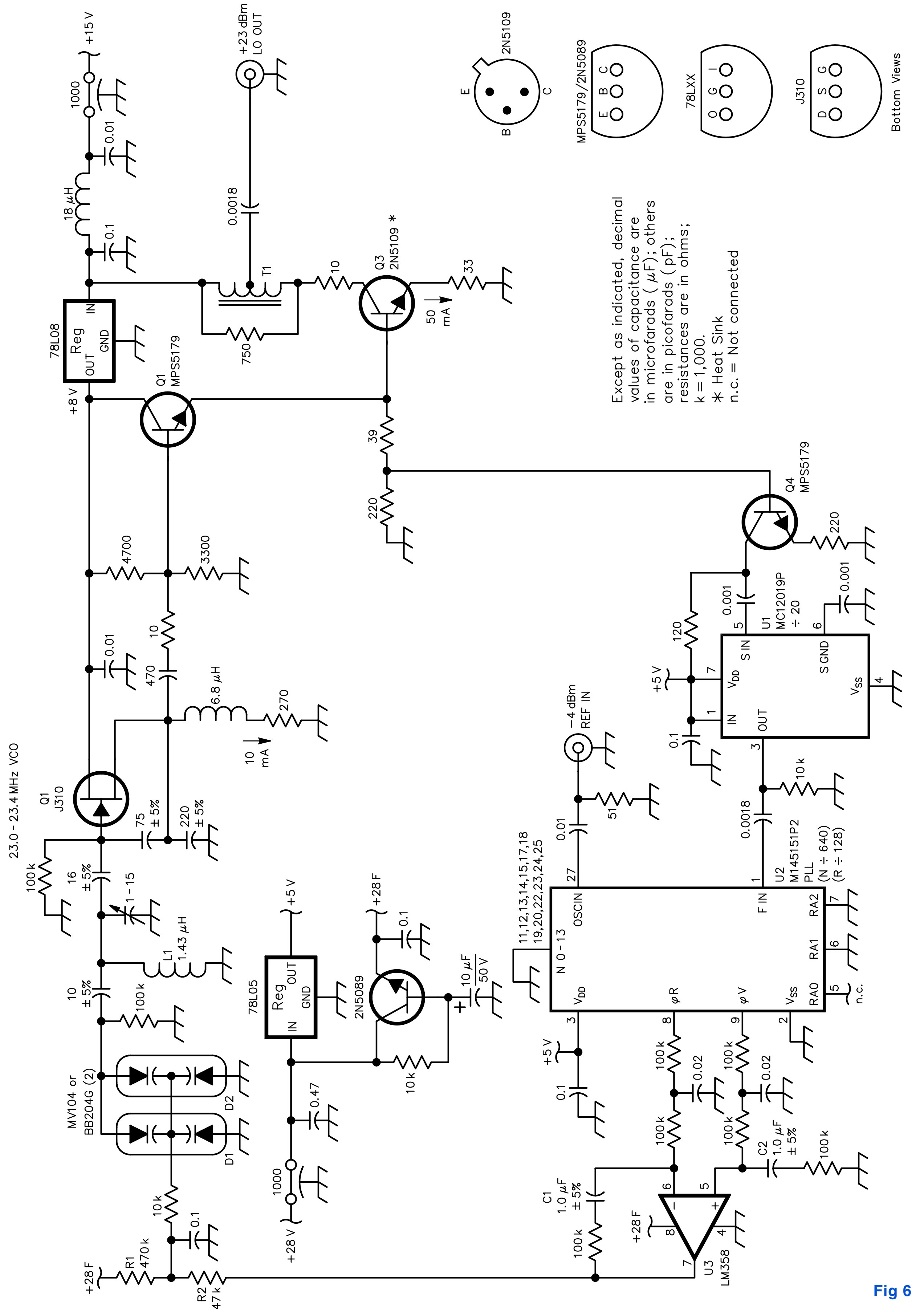
23.0 - 23.4 MHz VCO
Q1 J310
MV104 or BB204G (2)
D1
D2
L1 1.43 μH
78L08 Reg IN OUT GND
+8 V
+15 V
18 μH
Q1 MPS5179
T1
750
0.0018
+23 dBm LO OUT
Q3 2N5109 *
50 mA
10 mA
6.8 μH
Q4 MPS5179
2N5089
78L05 Reg IN OUT GND
+5 V
+28 V
+28 F
10 μF 50 V
−4 dBm REF IN
U1 MC12019P ÷ 20
U2 M14151P2 PLL (N ÷ 640) (R ÷ 128)
11,12,13,14,15,17,18 19,20,22,23,24,25
OSCIN
F IN
N 0 - 13
RA2
RA1
RA0
n.c.
C1 1.0 μF ± 5%
C2 1.0 μF ± 5%
U3 LM358
R1 470 k
R2 47 k
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
∗ Heat Sink
n.c. = Not connected
2N5109
E B C
MPS5179/2N5089
E B C
78LXX
O G I
J310
D S G
Bottom Views

Fig 6

power level. A 1.43-µH inductor with an unloaded Q of 220 was selected as the best available toroidal inductor at 23 MHz. A loaded Q of 70 and a tuning range of 500 kHz were selected after simulating the VCO on a program that I wrote several years ago. It uses Leeson's equation[23] to calculate phase noise based on several variables, including amplifier noise figure, resonator Q and varactor series resistance for several different varactor bias voltages. A second-order PLL is used with a long-time-constant (200 ms) integrator so that close-in spurs from the DDS are rejected unless they fall within the passband.

Fig 5 shows the results of the simulation. *T1* is the time constant of the integrator and *t2* is the time constant of the phase-lag network required for loop stabilization. F_o is the predicted VCO frequency for the voltage, V_s, applied to the varactor. *Bandwidth* is the loop bandwidth, ω_n is the natural frequency, $L(f_m)$ is the SSB phase noise, and *dF* is the residual FM component of the VCO output. T_{acq} would be the acquisition time if this were an analog phase-locked loop with no frequency detector; but it does not apply in this case, where a phase-frequency detector is used in the PLL chip.

The LO module, shown in Fig 6, contains the VCO, isolation amplifier, prescaler and PLL. A J310 junction FET was selected for the oscillator circuit because of its low noise figure and high drain-current capability. The Seiler oscillator circuit was used because it allows a high loaded Q with relatively large inductor values. This lets the inductor value be selected to maximize its unloaded Q. Another advantage is that the circulating current in the resonant circuit is split among three paths: the varactor tuning diodes, the parallel capacitance across the inductor and the capacitors coupling to the FET. This minimizes the effect of the varactor Q and allows the use of multiple smaller-valued capacitors, which in turn minimize the effects of lead inductance.

The capacitor values were selected to provide the required loaded Q and tuning range using *ARRL Radio Designer*. The oscillator was simulated by breaking the feedback loop at the source lead. Fig 7 shows the result for the lowest and highest VCO frequencies. During simulation, the loaded Q was set by varying the 16-pF capacitor value and the tuning range was changed by varying the 10-pF capacitor. Gain was set to approximately 4 dB to allow limiting using the square-law characteristics of the J310 FET. The ratio of the 75-pF and 220-pF capacitors

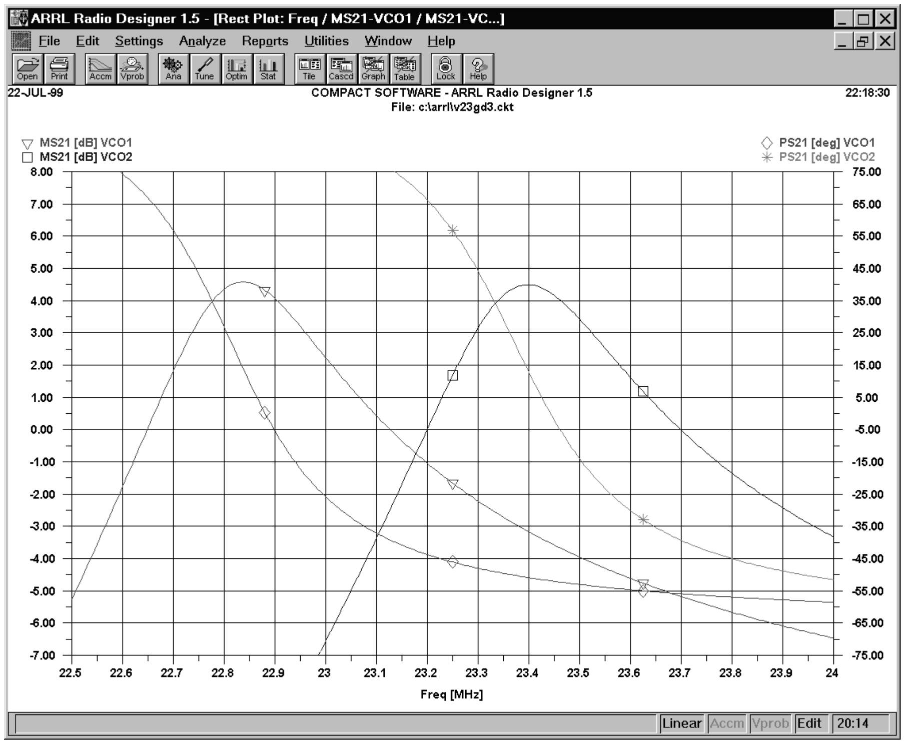

Fig 7—LO gain and phase plot.

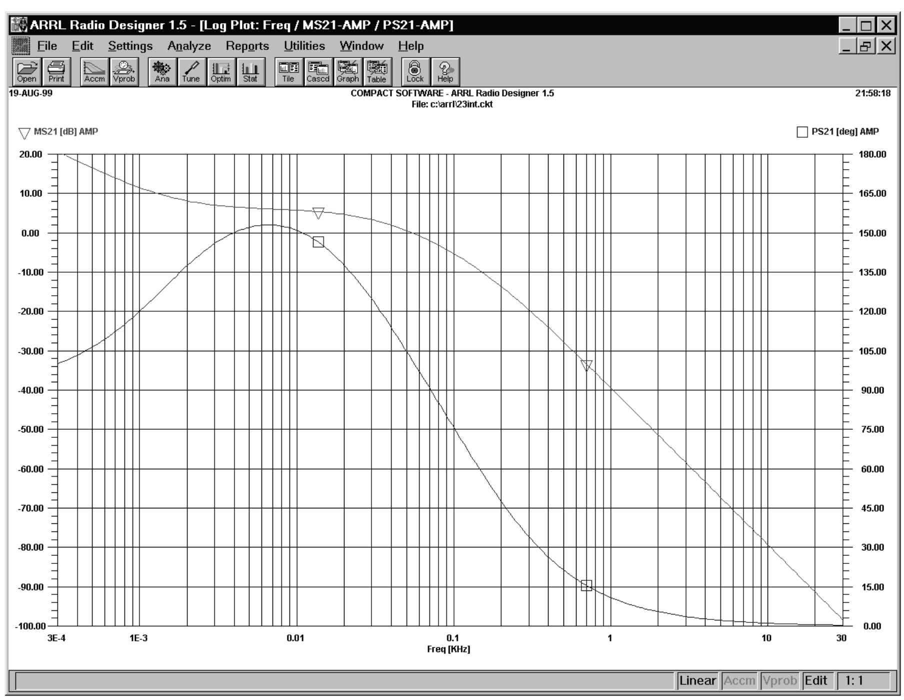

Fig 8—PLL filter amplitude and phase response.

Fig 6—LO schematic (opposite). Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. Capacitors are ±20% unless otherwise indicated.
L1—17 turns #24 AWG on T50-6 powdered iron core
T1—8 turns #28 AWG bifilar on FT37-61 ferrite core

sets the output voltage to approximately 3 V P-P.

Since the voltage across L1 is approximately 50 V P-P, varactor diodes cannot be placed directly across the inductor. A capacitive divider (10 pF in series with the varactors) is used to lower the voltage across the diodes, which provide 30-70 pF. The actual capacitance variation across L1 is 1.25 pF. A series-parallel combination of varactors is used to obtain the high capacity while minimizing distortion of the oscillator output. RF voltage is split across the top and bottom diode pairs; it affects the capacitance of each in opposite directions.

The VCO shown covers the 20-meter amateur band with a 9-MHz IF. It can be easily modified for a 10.7-MHz IF by removing one turn from L1.

The isolation amplifier consists of an emitter follower (Q2) to minimize oscillator loading followed by a common-emitter power stage (Q3) that provides 9 V P-P into a 50-Ω load. The 10-Ω and 750-Ω resistors in Q3's collector suppress spurs that occur during zero-crossings when driving a DBM. A second common-emitter output stage (Q4) drives U1, which divides the VCO output by 20 before application to the PLL, U2. The signal (N) and reference (R) counters in U2 are programmable by connecting pins to ground. R is set to divide by 128 and N to divide by 32. This results in division of the VCO output by 640 so the external reference input is one-fifth of the VCO frequency. The internal reference frequency is approximately 36 kHz.

U3 integrates the error voltage from the PLL chip and drives the varactor diodes, D1 and D2, to tune the VCO. RC filters are included on the input and output of U3 to suppress the reference frequency and prevent it from modulating the VCO. The cutoff frequencies are chosen to be less than 10% of the natural frequency of the loop to avoid instability. Fig 8 shows the frequency response.

The last potential source of noise is the power supply. Any 60- or 120-Hz hum modulation from power-supply ripple or ground loops must be suppressed. The VCO and PLL are powered from on-board 8- and 5-V regulators. Q5 and the associated RC components filter the power for U3,

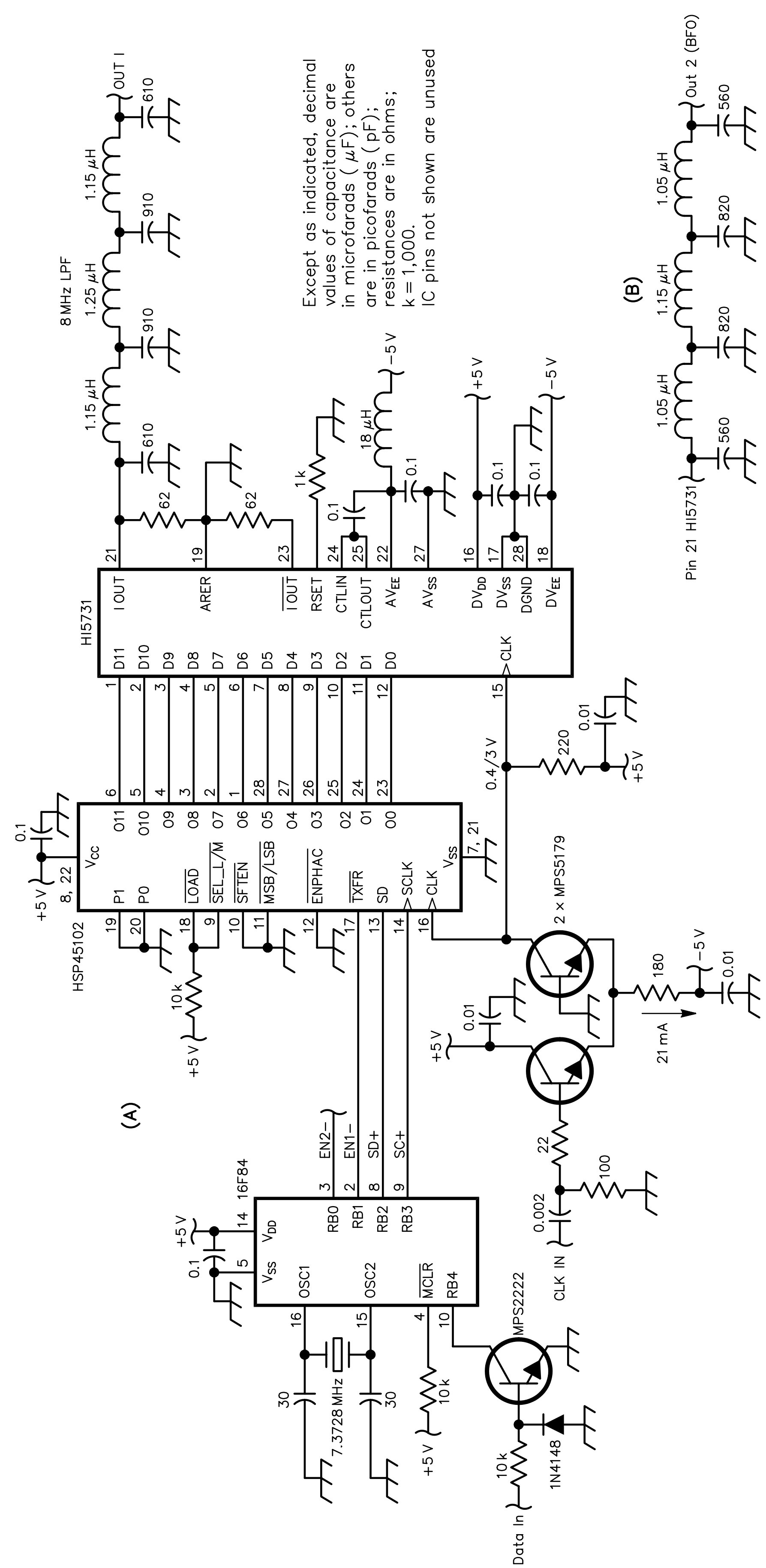

Fig 9 (right)—(A) PIC and LO DDS schematic diagram. The low/high clock levels to pin 15 of the HI5731 should be 0.4/3-V. (B) BFO low-pass filter schematic diagram.

which generates the varactor control voltage. This circuit allows the use of a much smaller capacitor value than would otherwise be required.

Direct Digital Synthesizer

Each DDS consists of a Harris HSP45102 numerically controlled oscillator (NCO) and a Harris HI5731 digital-to-analog converter (DAC). One DDS generates the one-fifth LO reference frequency and the other supplies the BFO signal. This combination results in all spurs being at least 69 dB below the output. The BFO DDS drives the product detector/balanced modulator directly. The schematic for one channel of the dual DDS is shown in Fig 9. The second channel is identical, except that the low-pass filter's component values are 10% lower to allow operation of the BFO DDS at 9 MHz.

A CMOS-level clock is required for the DDS chips. The square wave is created by a differential amplifier using two MPS5179 transistors. The amplifier is driven by an external 30-MHz, sine-wave reference at +5 to +7 dBm. The DAC output level following low-pass filtering is approximately +4 dBm.

The DDSs are controlled by a common Microchip PIC16F84. The firmware implements a UART running at 19,200 b/s to receive commands from a PC that set the transmit and receive frequency registers in the DDS. The MPS2222 and associated diode and resistors implement the RS-232 receiver. The commands implemented are character strings as shown in Fig 10. They are sent as seven-bit characters.

The ASCII STX and ETX characters are used to delineate the commands. Each command is followed by a longitudinal redundancy check (LRC) byte that represents an even-parity value computed over each of the lower seven bits. The most-significant bit is ignored in all received characters as drivers in many PCs alter the state of this bit. The first byte after STX is an ASCII character specifying the command. The remaining bytes before the ETX are the bits loaded into the DDS control registers. Only the lower four bits of each byte are used and they are transmitted LSB-first. A program on the PC calculates the appropriate DDS register values as follows:

$$\textit{DDS Register Value} = \frac{2^{32} F}{30 \times 10^{6}} \qquad \text{(Eq 1)}$$

The LO frequency must be divided by five in the transceiver-control software on the PC before transmission to the DDS.

Phase-Noise Results

The LO phase noise was measured using the test fixture in Fig 11. The LO was mixed with a low-noise crystal oscillator and the output passed through a narrow filter for measurement. I used a spare 500-Hz bandwidth, eight-pole crystal filter and some fixed-gain wide-band amplifiers to process the signal before application to a low-frequency spectrum analyzer. The low-noise crystal oscillator is described in Appendix A. A step attenuator was used to vary the input to the mixer from the crystal oscillator to prevent overload of the spectrum analyzer. The gain of the 9-MHz IF amplifiers and 500-Hz bandwidth filter was measured to be +46 dB. Losses in other components were known. This allowed measurement of phase-noise levels from the

Byte 1	*Byte 2*	*Bytes 3-10*	*Byte 11*	*Byte 12*	*Command*
STX	T	DDS register value	ETX	LRC	LO TX Frequency
STX	R	DDS register value	ETX	LRC	LO RX Frequency
STX	t	DDS register value	ETX	LRC	BFO TX Frequency
STX	r	DDS register value	ETX	LRC	BFO RX Frequency
STX	F	DDS register value	ETX	LRC	LO TX/RX Freq.
STX	f	DDS register value	ETX	LRC	BFO TX/RX Freq.

Fig 10—DDS command strings.

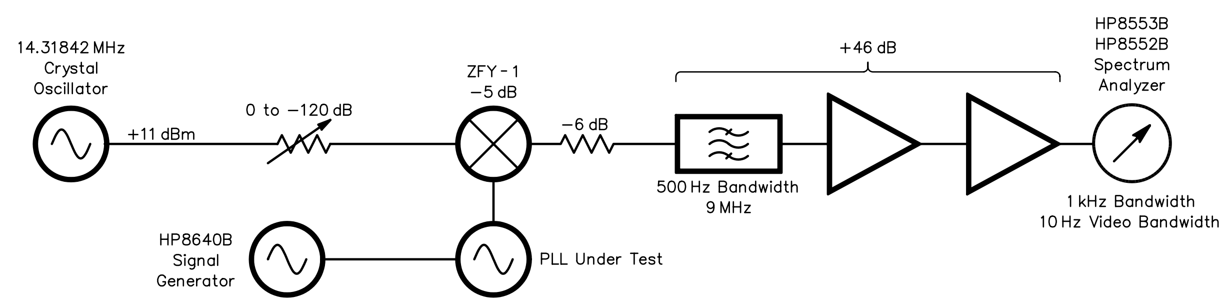

Fig 11—LO phase-noise test setup.

Table 1—LO Phase Noise Measurements

Offset (kHz)	*Expected Phase Noise (dBc/Hz)*	*Measured Phase Noise (dBc/Hz)*
1	–113	–110
2	–122	–121
5	–133	–134
10	–141	–144

Table 2—Local Oscillator Requirements

Band	*RF (MHz)*	*LO (MHz)*	*Min. Ql*	*Min. Qu*
160 m	1.80-2.00	10.80-11.00	28	83
80 m	3.50-4.00	12.50-13.00	33	100
40 m	6.90-7.30	15.90-16.30	41	123
30 m	10.10-10.20	19.10-19.20	48	144
20 m	14.00-14.40	23.00-23.40	59	176
17 m	18.07-18.17	27.07-27.17	68	204
15 m	21.00-21.45	30.00-30.45	76	229
12 m	24.90-25.00	33.90-34.00	85	255
10 m	28.00-30.00	37.00-39.00	98	293
6 m	50.00-54.00	41.00-45.00	113	338

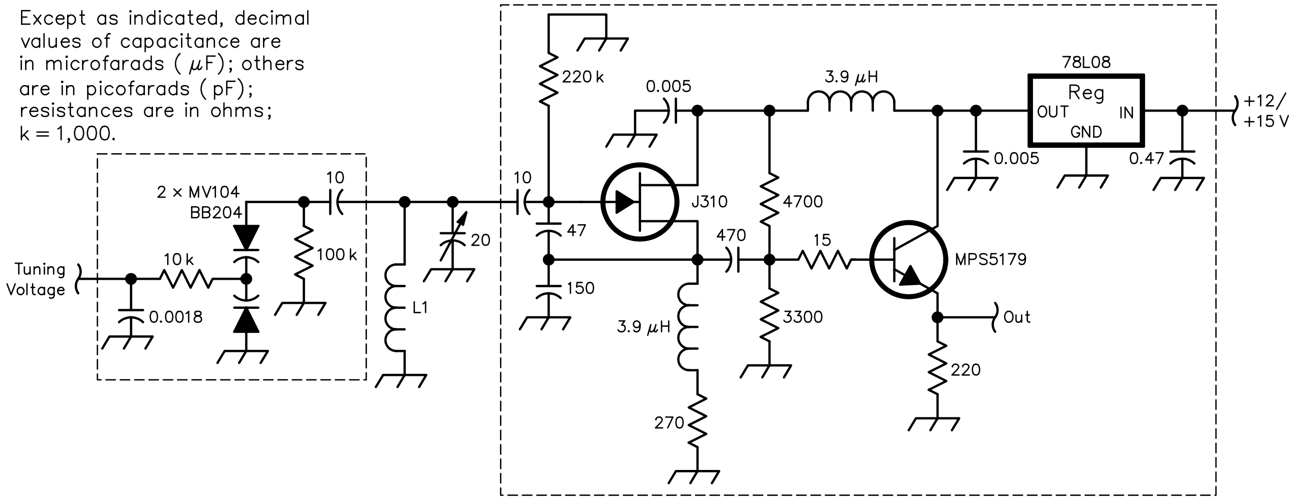

Fig 12—10-meter VCO schematic diagram.

Appendix A

To make meaningful measurements on this receiver, I needed a signal source with a noise floor lower than my Hewlett Packard 8640B signal generator and with lower close-in phase noise. A crystal oscillator was the only possible solution. Crystals at 14.31818-MHz are commonly available for microprocessors, so I constructed the crystal oscillator and filter in Fig A using three of these crystals.

Y1 is placed in an oscillator that generates approximately +10 dBm of output at 14.3185 MHz. A 6-dB attenuator and a crystal filter follow the oscillator. The 300-pF capacitor in series with Y1 tunes the oscillator to the center frequency of the filter. Y2 and Y3, along with the 300-pF capacitor, form the crystal filter. Its +3 dB bandwidth is 1.2 kHz. The –30 dB bandwidth was measured at 8.8 kHz. Insertion loss was measured at 3 dB, so the input to the MSA1104 amplifier is approximately +1 dBm. This should result in a noise floor of –171 dBc/Hz. I can't measure this directly, but given the data provided by HP and the measured signal level at the filter output, it should be correct. The final output level, after amplification and low-pass filtering, was measured to be +13.8 dBm.

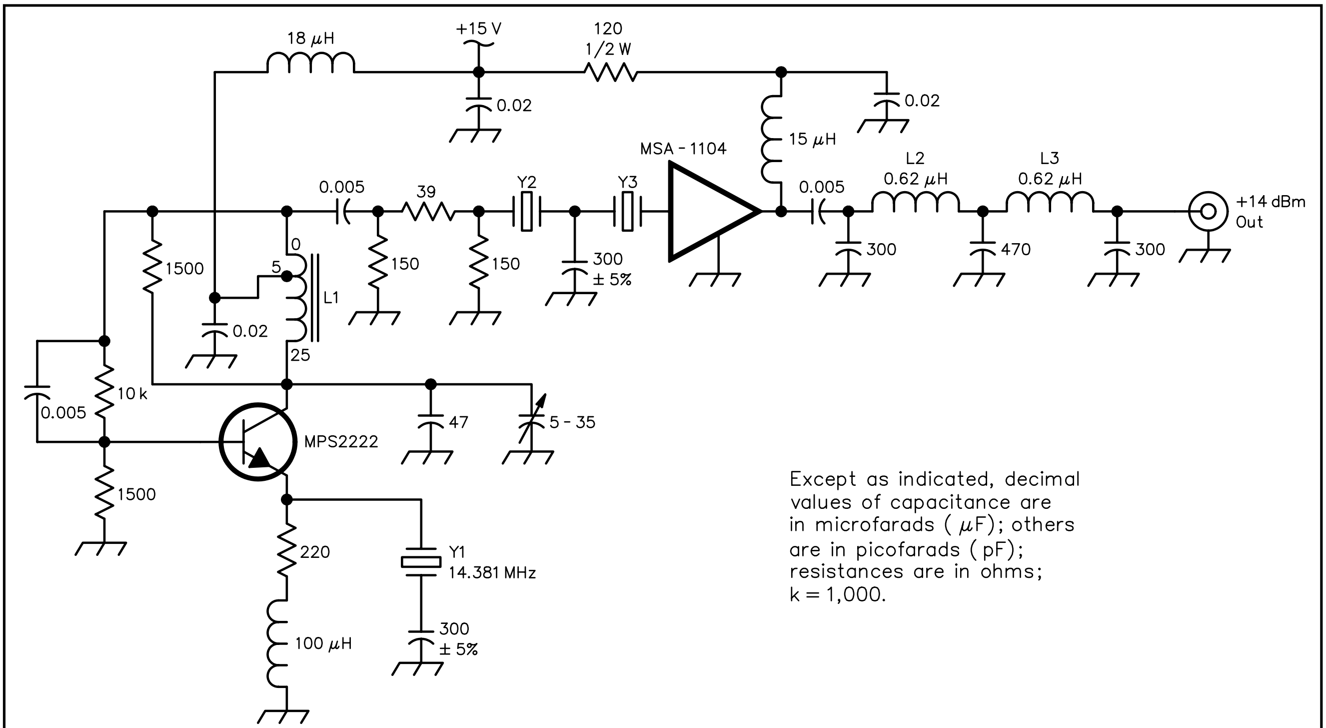

Fig A—Low-noise crystal oscillator schematic diagram.

L1—2.63 µH, 25 turns #26 AWG on T44-6 powdered iron core, tap at 5 turns

L2, L3—0.62 µH, 14 turns #28 AWG on T25-6 powdered iron core

spectrum-analyzer display with an accuracy of ±3 dB.

Measurements were made at offsets of from 1 to 10 kHz, resulting in the LO performance shown in Table 1. Phase noise was slightly lower than expected at the 10-kHz offset and slightly higher than expected at the 1-kHz offset, but the difference between predicted and measured values was still within the error range of this test setup. There may also be some error induced by inadequate attenuation by the filter at low offsets. I will be repeating phase-noise mea-

```
1. L=  0.690uH     2. Q= 100       3. Cpad=  18.1pF     4. Cv=  70.0pF
5. Ev=     4.0V    6. Rs= 1.0      7. Fref=   58.0kHz  8. Fofs=    0.0MHz
9. Epd=    5.0V   10. t1= 200ms   11. NF=  6.0dB       12. Pi=  0.0 dBm
13. Fc= 10.00kHz 14. Cser=   10.0pF 15. Hyperabrupt: 0
t2=  61.31771ms                                   --------- L(Fm) dBc ---------
Vs(V) Fo(MHz) dF(Hz) Wn(r/s) damp. Tacq(s) BW(Hz)  50  500   2K   10K 100K   1M
  4.0   36.98    0.5    56.0  1.72  532.35    33  -76 -106 -123 -141 -163 -171
  6.0   37.12    0.5    50.3  1.54  661.52    27  -76 -106 -123 -142 -163 -171
  8.0   37.24    0.4    46.3  1.42  779.68    23  -76 -106 -123 -142 -163 -171
 10.0   37.34    0.4    43.3  1.33  890.61    21  -76 -106 -123 -142 -163 -171
 12.0   37.43    0.4    41.0  1.26  996.37    19  -76 -106 -123 -142 -163 -171
 14.0   37.52    0.4    39.0  1.20 1098.22    17  -76 -106 -123 -142 -163 -171
 16.0   37.59    0.3    37.4  1.15 1197.00    16  -76 -106 -123 -142 -163 -171
 18.0   37.66    0.3    35.9  1.10 1293.28    15  -76 -106 -123 -142 -163 -171
 20.0   37.73    0.3    34.7  1.06 1387.51    14  -76 -106 -123 -142 -163 -171
 24.0   37.84    0.3    32.6  1.00 1571.03    13  -76 -106 -123 -142 -163 -171
```

Fig 13—Predicted 10-meter VCO phase noise.

Appendix B

I use either a 10-MHz Rubidium oscillator or a 10-MHz oven-controlled crystal oscillator to generate the reference frequency for the transceiver. Both were obtained on the surplus market. The circuit in Fig B generates the 30-MHz reference for the receiver by tripling the output of a 10-MHz frequency standard.

The tripler consists of two 50-Ω resistive-feedback amplifiers, a passive tripler, and a filter. The first amplifier brings the 10-MHz reference up to the 20-mW level to drive two diodes configured as a clipper. The anti-parallel connection suppresses even harmonics to simplify filtering. The input and output are isolated from each other by low-Q series-resonant circuits tuned to the fundamental and third harmonic. These are constructed with ±5%-tolerance silver-mica capacitors and ±10%-tolerance molded RF chokes. They require no tuning. The main filter is a conventional three-pole, top-coupled Butterworth tuned for peak output. The final amplifier generates 50 mW for distribution throughout the station via power splitters.

The output spectrum is fairly clean as shown in Table B. These levels were measured with +7 dBm input at 10 MHz and +16.8 dBm output at 30 MHz.

Table B—Tripler Output Spectrum

Frequency (MHz)	*Output (dBc)*
10	–70
20	< –78
30	0
40	< –78
50	–71
60	–35

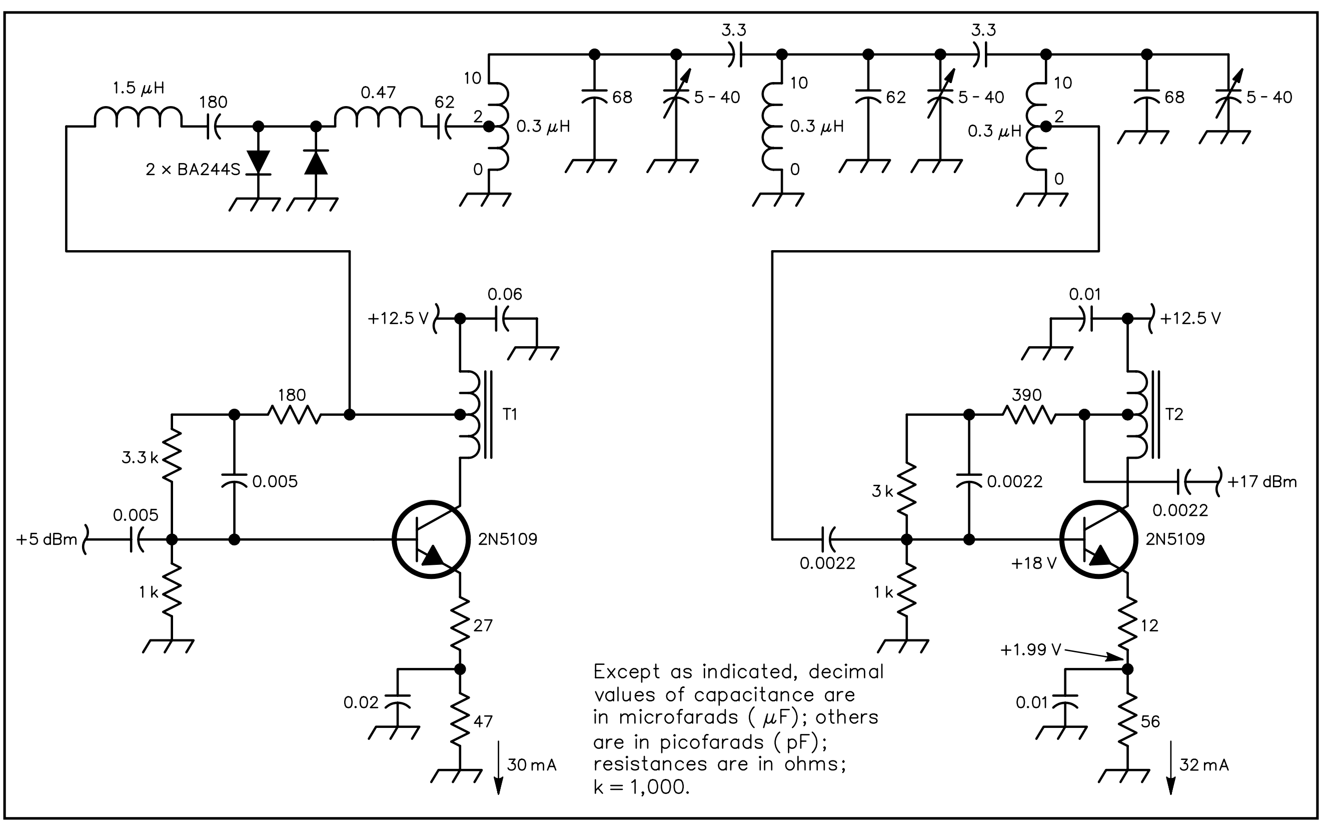

Fig B—10 to 30 MHZ tripler schematic.

surements on the completed receiver with the actual filters and IF amplifier.

Extension to Other Bands

To extend the transceiver to other bands, additional LO- and RF-filter modules must be constructed. Design and construction of RF bandpass filters is straightforward as the Q of the filter need not exceed 25 to achieve adequate image rejection (90 dB). The PLL portion of the LO works up to 225 MHz. The VCO design is more problematic, as we must maintain a resonator bandwidth of 400 kHz or less to meet the phase-noise requirements laid out in the beginning of this article. If the power level of the VCO is lowered, the bandwidth of the resonator must be even smaller. As we go lower in frequency, design and construction of the LO module is simpler, since the required loaded Q decreases as shown in Table 2. As we go to higher frequencies, however, the loaded Q increases proportionally.

The problem is that the unloaded Q of the inductor must be at least three times its loaded Q in the oscillator circuit. For 6 meters, this means that the inductor must have a Q of 338 or more. Toroidal inductors using iron-powder cores cannot achieve these Qs at the frequencies required. In fact, the Q of the inductor in the 23-MHz LO is about the best that we can do today at that frequency.

Luckily, a solution is very easy and inexpensive to employ in home-built equipment. A toroidal coil is actually very inefficient, as its Q is much less than can be achieved with a solenoid coil occupying the same volume.[24] Very high-Q solenoid coils can be constructed that easily meet the requirements above by making them as large as possible with the appropriate length-to-diameter ratio.[25]

The disadvantage of a solenoid coil is its high leakage inductance. Unlike a toroidal coil, it must be shielded to prevent radiation. In addition, the shield must be spaced at least one diameter from the coil so that the Q is not degraded by eddy currents in the shield metal. This may no longer be the best solution for commercial equipment because of labor costs and the desire for small size, but it is a very good solution for home construction.

The maximum Q for a given coil diameter is obtained when the length-to-diameter ratio is about two.[26] The obtainable Q is:

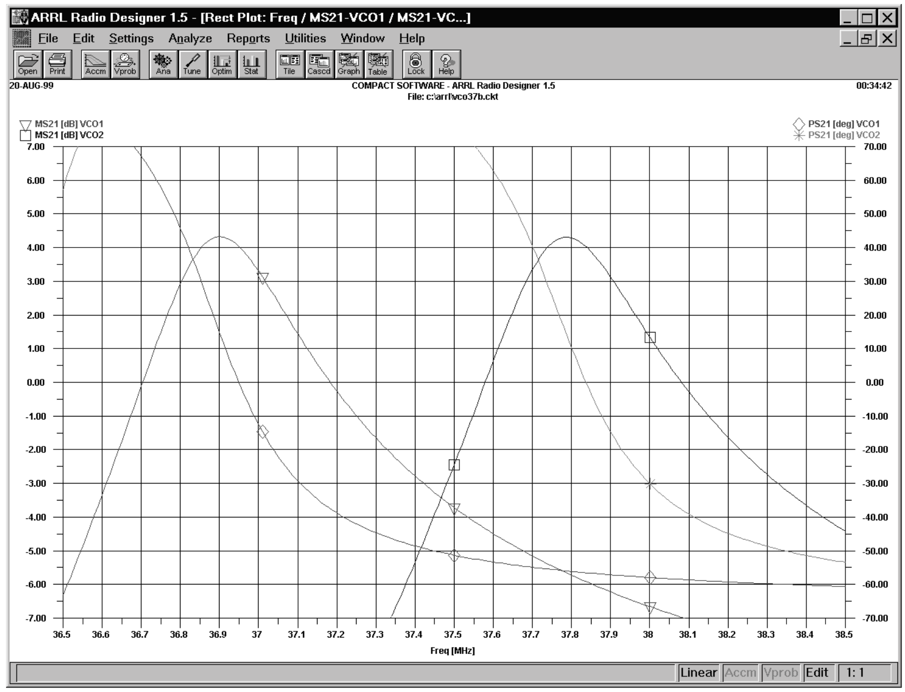

Fig 14—Predicted 10-meter VCO gain and phase.

$$Q_{max} = (120D)F^{-\frac{1}{2}} \quad \text{(Eq 2)}$$

where *D* is the diameter in inches and *F* is the frequency in megahertz. Maximum diameter is limited, however, because its size must be small compared to a wavelength. The maximum practical size is about 30/*F* inches in diameter. The maximum-possible Q is therefore:

$$Q = 3600F^{-\frac{1}{2}} = 537 \quad \text{(Eq 3)}$$

at the highest LO frequency of 45 MHz.

To test this type of resonator, I constructed a VCO for the 10-meter amateur band. The circuit (shown in Fig 12) is identical to the 20-meter LO, except for the inductor and capacitor values. L1 is 10 turns of #14 solid copper wire, 5/8-inch ID and 1¼ inches long. This is smaller than the maximum diameter (0.67 inch) at the highest frequency (45 MHz). The calculated inductance is 0.69 µH and the calculated Q is 456 at the lowest frequency (37 MHz). C1 is a chassis-mount, air-dielectric variable capacitor for maximum Q. A ceramic or Mylar trimmer would probably degrade the resonator Q. A simulation of the VCO shows that phase noise is low enough with a loaded Q of 100. Capacitor values were adjusted to provide the appro-priate Q and tuning range using *ARRL Radio Designer*. The simulation results are shown in Figs 13 and 14.

Note that I made the tuning range wider than for 20 meters, but not wide enough to cover the entire 10-meter band. An 800-kHz tuning range seems to cover almost all terrestrial SSB and digital operation and is consistent with the bandwidth of my 10-meter beam. I have a separate transverter and antenna for the satellite portion of the band and have never used 10-meter FM.

To allow space for the coil, the VCO was constructed in a 2¼×2¼×5-inch aluminum Minibox and coupled to a PLL in a separate box. The tuning range is 1 MHz—slightly wider than predicted—and C1 was only about 25% meshed. This is probably due to stray inductance in the component leads so the inductor turns were spread apart to lower the total inductance. The inductor should be constructed 1½ inches long for better results. Power output across the band is as predicted and constant, so the inductor Q is adequate. Phase noise seems consistent with the 20-meter LO when heterodyned with the output of a HP 8640 signal generator.

I do not currently have a low-noise crystal oscillator in the 10-meter band so I have not been able to measure close-in phase noise or the noise floor. However, there is no reason to expect that they differ significantly from what is predicted given the measured performance of the 20-meter version of the circuit.

Summary

The 20-meter front end was verified to have approximately the required performance after construction and measurements. In addition, extension of the design to other amateur bands within the HF and lower VHF spectrum was tested and shown possible. Part 2 of this series will cover the remaining portion of the interface to the PC as well as the noise gate, IF amplifier, AGC, speech processing and audio circuits. Part 3 will cover the 150-W amplifier, power supply and noise receiver.

Notes

[1]At the beginning of the project I assumed that I would either pass the 13 WPM code test by the fall of 1999 or that new FCC rules would be in effect lowering the code requirement to 5 WPM. Neither happened, so I tested the 20-meter transmitter portion into a dummy load and then constucted a 10-meter front end for on-the-air transmitter testing.

[2]Some may disagree, citing the situation on the 40-meter band where International Broadcast stations coexist with amateur stations from 7100-7300 kHz. However, the minimum noise level on this band is +37 dBN_0, which creates a –103 dBm (S4) noise level in a 2.5 kHz bandwidth. When connected to a dipole or larger antenna, all but the simplest 40-meter receivers should be operated with an attenuator ahead of the mixer to maximize their dynamic range. With a 12-dB noise figure, the attenuation can be 18 dB without degrading weak-signal reception. A receiver with a 101-dB SFDR will then handle –9 dBm (S9 + 64 dB) signals without spurious responses. Unless you are located within 10 km of two 500-kW transmitters, this should provide all the performance needed.

[3]SSB transmitters using audio tones to generate FSK (and possibly PSK) will probably generate spurious signals stronger than this level. Low-price HF amateur radios with synthesized VHF LOs will also transmit noise sidebands higher than this level. Older equipment with crystal-controlled and mechanically-tuned LC oscillators will actually generate the least interference at this spacing.

[4]B. Goodman, "What's Wrong with Our Present Receivers," *QST*, Jan 1957.

[5]W. K. Squires, "A New Approach to Receiver Front End Design," *QST*, Sep 1963.

[6]"Recent Equipment—The Collins KWM-1 Transceiver," *QST*, April 1958.

[7]W. J. Hall, "The ATR-166—A Homemade Transceiver for the 160- to 6-Meter Bands: Parts 1 and 2," *QST*, Feb and Mar 1971.

[8]U. Rohde, "Increasing Receiver Dynamic Range," *QST*, May 1980.

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[10]U. Graf, "Performance Specifications for Amateur Receivers of the Future," *QEX*, May/Jun 1999.

[11]J. Craswell, "Weekend DigiVFO," *QST*, May 1995.

[12]B. Carver, "A High-Performance AGC/IF Subsystem," *QST*, May 1996.

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[14]W. K. Squires, "A Pre-IF Noise Silencer," *QST*, October 1963.

[15]L. E. Campbell, "Recent Equipment—Collins Noise Blanker (KWM-1 and KWM-2)," *QST*, Nov 1959.

[16]M. Mandelkern, "Evasive Noise Blanking," *QEX*, Aug 1993.

[17]Most of the filters used here are from KVG and were distributed in the US by Spectrum International. These are no longer available. Similar filters are now available from International Radio, 13620 Tyee Rd, Umpqua, OR 97486; tel 541-459-5623 9AM-1PM PDT Tues-Sat, fax 541-459-5632; **inrad@rosenet.net**; **http://www.qth.com/inrad/**.

[18]W. Sabin, W0IYH, "RF Clippers for SSB—Observations on Measurements and On-the-Air Performance," *QST*, Jul 1967.

[19]PNP transistors provide a lower noise figure below 1 kHz than NPN transistors at the same current levels.

[20]R. Campbell, "High-Performance Direct-Conversion Receivers," *QST*, Aug 1992.

[21]B. Goodman, "Better AVC for SSB and Code Reception," *QST*, Jan 1957.

[22]H. R. Hyder, "Atmospheric Noise and Receiver Sensitivity," *QST*, Nov 1969.

[23]V. Manassewitsch, *Frequency Synthesizers—Theory and Design* (New York: Wiley, 1987).

[24]N. B. Watson, W6DL, "Relative Merit of Toroidal and Conventional RF Inductors," *QST*, Jun 1968.

[25]R. S. Naslund, W9ISA, "Optimum Q and Impedance of RF Inductors," *QST*, Jul 1941.

[26]R. W. Rhea, *Oscillator Design and Computer Simulation* (Englewood Cliffs, New Jersey: Prentice-Hall, 1990; New York: McGraw-Hill, 1997).

John Stephensen, KD6OZH, has been interested in radio communications since building a crystal radio kit at age 11. He went on to study Electronic Engineering at the University of California and has worked in the computer industry for 26 years. He was a cofounder of Polymorphic Systems, a PC manufacturer, in 1975 and a cofounder of Retix, a communications-software and hardware manufacturer, in 1986. Most recently, he was Vice President of Technology at ISOCOR, which develops messaging and directory software for commercial users and ISPs. John received his Amateur Radio license in 1993 and has been active on the amateur bands from 28 Mhz through 24 GHz. His interests include designing and building Amateur Radio gear, digital and analog amateur satellites, VHF and microwave contesting and 10-meter DX. His home station is almost entirely home-built and supports operation on SSB, PSK31, RTTY and analog and digital satellites in the 28, 50, 144, 222, 420, 1240, 2300, 5650 and 10,000 MHz bands from Grid Square DM04 in Los Angeles. The mobile station includes 10-meter SSB, 144/440-MHz FM and 24-GHz SSB.

A Homebrew Multiband Yagi

This antenna comprises monoband Yagis for the 20- through 10-meter bands on a single boom. Boom length may be 40, 50 or 60 feet. Performance data include comparisons against similarly modeled commercially made antennas.

By Johan Van de Velde, ON4ANT

As most of us know, monoband Yagis are by far the best choice when building an antenna farm. Unfortunately, most hams don't have the room to put up multiple towers with these antennas. The average Amateur relies on a multiband Yagi with traps. This choice allows him to operate on several bands, but has some drawbacks, such as reduced band-width, antenna gain and front-to-back ratio (F/B).

Over the years, several designs have been developed that cover a number of bands—most of them for two bands—using separate elements on the same boom. These so-called *interlaced* Yagis offer a valuable alternative to designs using traps. My desire to cover a fair number of bands with sufficient gain and SWR bandwidth left me no choice but to build one myself. The principle is easy: Just put some monoband antennas on a single boom.

At first, the boom rapidly gets quite long, especially if one wants to cover the 20- through 10-meter bands. In my design, the boom length was kept to a manageable 50 feet, with 40- and 60-foot options. This should allow most to build this design. All bands from 20 through 10 meters are covered. All calculations have been done with *AO*, *YO*, *EZNEC/4*, *STRESS* and *YAGI DESIGN*.[1, 2, 3, 4, 5]

This article is logically divided into three sections. First, there's a description of the electrical design including information on feed-point impedance, gain and bandwidth. Mechanical details such as element lengths and spacing are also given. We cover the standard (50-foot boom) and the two modified designs (40- and 60-foot booms).

Details of feed methods, element tapering and wind-load considerations come next. The element tapering is done as a function of maximum wind-survival speed. The total wind load is calculated, too.

Performance comparisons weigh the modeled performance data for this antenna against that for the most-common commercial Yagis available on today's market. Trapped triband-

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Kester, 1755
Belgium
ON4ANT@hotmail.com

[1]Notes appear on page 19.

ers are not included because most manufacturers give unrealistic gain figures, and no design models are available.

Electrical and Mechanical Design

All values indicated are free-space figures and the gain is referenced to an isotropic radiator (0 dBd = 2.15 dBi). We don't account for the influence of the Earth. The actual gain is easily 4 dB higher if one accounts for ground effects. The actual antenna setup will influence the radiation pattern, as well.

Table 1 gives element lengths for a constant element diameter of 20 mm and element spacing for the 50-foot boom model. The table also gives electrical parameters of interest.

We have gain comparable with three- or four-element monoband beams, excellent bandwidth and F/B. The calculated antenna performance shows almost constant gain over these five bands. SWR is excellent over all the bands with the exception of 10 meters, where the upper frequency limit is 28.8 MHz. Our reference here is a SWR value normalized to the match frequency. It's obvious that things still can be improved, but other parameters might be compromised. On 10 meters, for instance, we could have another 0.5 dB of gain if one adapts the element lengths and spacing. However, the impedance around 29 MHz becomes very low (4 Ω) and the antenna won't be useable there.

Another disadvantage of this higher gain is that the antenna becomes less tolerant of element errors. If you really want more gain with the same bandwidth, the only solution is a longer boom.

Variant 1: The 40-foot Boom

Is 50 feet too long? Go for the 40-foot boom. On 20 meters, it has one less director and gain drops to about 7 dBi. Its performance is still excellent.

Table 2 shows the element lengths, spacing and electrical parameters for the 40-foot-boom design. The gain figures on the higher bands do not change.

Variant 2: The 60-foot Boom

Are you fortunate enough to be able to put up a 60-foot boom? This design is the ultimate! Higher gain figures for the upper three bands are obtained with excellent bandwidth.

Table 3 shows the element lengths, positions and electrical parameters for the 60-foot-boom design. This particular design is mounted at my home location; the calculated values correspond well to those I measured. Initial testing shows that this antenna works. The design has broad bandwidth and permits different feed methods.

Feed Methods, Element Tapering and Wind-Load Considerations

The elements resonate in band. The actual feed-point impedance is high enough to allow different feed methods. I chose not to split the driven element and went for a gamma match. A delta match is also possible.

One can choose isolated or non-isolated elements. The actual influence of the boom is minimal. Non-isolated elements are an advantage if you choose to use your tower as a top-loaded vertical for 160 meters (a la ON4UN). The boom-to-element mounting plate is 200×100 mm. *YAGI DESIGN* is one of the programs capable of calculating the influence of the boom-element plate and the isolated versus non-isolated element construc-

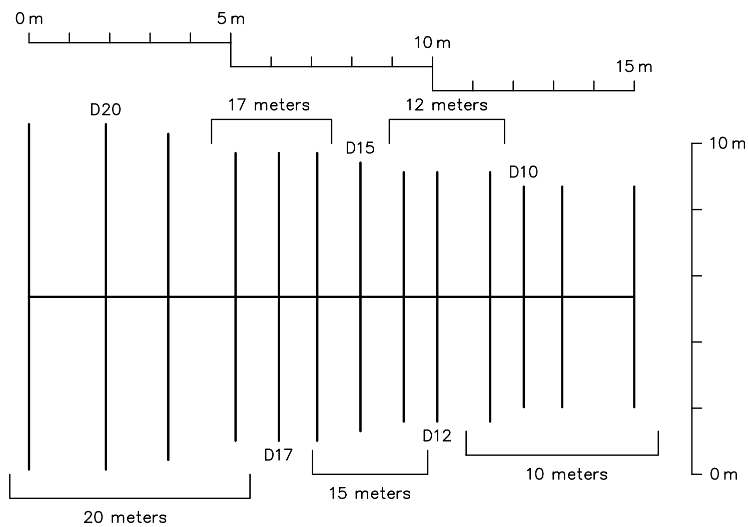


Table 1—Antenna with 50-foot Boom

Half-Element Lengths and Positions

Element Length (m)*	*Description*	*Position (m)*
5.45	Reflector 20	0.00
5.2	Driver 20	2.00
4.9	Director 20	3.60
4.15	Reflector 17, Director 20	5.25
4.02	Driver 17	6.20
3.8	Reflector 15, Director 17	7.20
3.395	Driver 15	8.40
3.02	Director 15, Reflector 12	9.50
2.91	Driver 12	10.30
2.78	Reflector 10, Director 12	11.60
2.55	Driver 10	12.45
2.355	Director 10	13.40
2.265	Director 10	15.00

Electrical Parameters

Frequency (MHz)	*Gain (dBi)*	*Impedance*	*F/B*	*SWR*
14.000	8.1	33.0–*j*4.1	26.8	1.26
14.175	8.2	30.9+*j*3.0	29.1	1.00
14.350	8.3	26.0+*j*12.2	25.9	1.44
18.068	8.1	20.9–*j*3.6	21.5	1.10
18.118	8.6	22.3–*j*2.3	22.3	1.00
18.168	8.6	23.5–*j*1.2	23.2	1.07
21.000	8.4	32.4–*j*7.8	21.1	1.27
21.200	8.5	34.2+*j*0.5	21.0	1.00
21.400	8.6	35.7+*j*8.1	20.9	1.25
24.880	8.5	10.7–*j*3.6	30.6	1.19
24.940	8.5	10.8–*j*1.7	30.6	1.00
24.990	8.5	10.8+*j*0.1	28.0	1.19
28.000	7.9	26.0–*j*7.2	29.7	1.47
28.350	8.1	26.9+*j*3.1	25.7	1.00
28.700	8.2	27.6+*j*13.9	22.6	1.48

*Lengths are for constant-diameter elements

Table 2—Antenna with 40-foot Boom

Half-Element Lengths and Positions

Element Length (m)*	*Description*	*Position (m)*
5.45	Reflector 20	0.00
5.2	Driver 20	2.00
4.15	Reflector 17, Director 20	3.05
4.02	Driver 17	4.00
3.8	Reflector 15, Director 17	5.00
3.395	Driver 15	6.20
3.02	Director 15, Reflector 12	7.30
2.91	Driver 12	8.10
2.78	Reflector 10, Director 12	9.40
2.55	Driver 10	10.25
2.355	Director 10	11.20
2.265	Director 10	12.80

*Lengths are constant-diameter elements

Electrical Parameters

Frequency (MHz)	*Gain (dBi)*	*Impedance*	*F/B*	*SWR*
14.000	7.2	33.5–*j*11.6	16.0	1.40
14.175	7.1	39.8–*j*0.9	29.1	1.00
14.350	7.0	45.3+*j*9.0	14.3	1.30

Other bands match the 50-foot-boom design

Table 4—Wind-Load Parameters

EIA-222-C 30 lb/sq-ft pressure at 86 mph.
Shape factor 0.666
No ice-load
Aluminum 6061-T6 (yield strength 35000)

Table 3 —Antenna with 60-foot Boom

Half-Element Lengths and Positions

Element Length (m)*	*Description*	*Position (m)*
5.45	Reflector 20	0.00
5.2	Driver 20	2.00
4.9	Director 20	3.60
4.15	Reflector 17, Director 20	5.25
4.02	Driver 17	6.40
3.8	Reflector 15, Director 17	7.20
3.395	Driver 15	8.40
3.02	Director 15, Reflector 12	9.50
2.91	Driver 12	10.80
2.68	Reflector 10, Director 12	12.00
2.55	Driver 10	13.014
2.47	Director 10	13.816
2.44	Director 10	15.775
2.31	Director 10	18.25

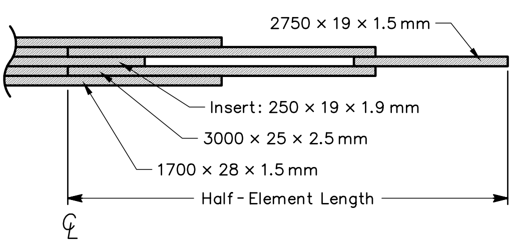
*Lengths are for constant-diameter elements

Electrical Parameters

Frequency (MHz)	*Gain (dBi)*	*F/B*
14.175	8.3	34
18.118	8.3	21
21.200	8.7	23
24.940	9.6	38
28.350	10.0	29

Table 5—Element Physical Characteristics for the 50-foot-Boom Antenna*

Elements 1, 2 and 3

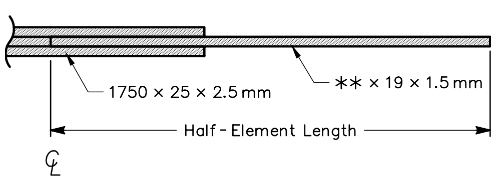


Half-element wind area is 0.13 m^2
Half-element weight is 2.6 kg
Element sag is 20.5 cm

Half-Element Lengths (100 mm overlap)*

	Isolated	*Non-isolated*
Element 1	5567 mm	5570 mm
Element 2	5309 mm	5312 mm
Element 3	4998 mm	5000 mm

Elements 4, 5 and 6

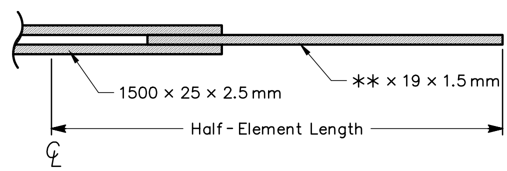


Half-element wind area is 0.084 m^2
Half-element weight is 1.85 kg
Element sag is 8.4 cm

Half-Element Lengths (100 mm overlap)*

	Isolated	*Non isolated*
Element 4	4222 mm	4224 mm
Element 5	4053 mm	4056 mm
Element 6	3852 mm	3854 mm

Elements 7, 8 and 9

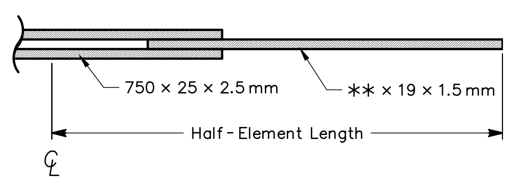


Half-element wind area is 0.074 m^2
Half-element weight is 1.2 kg
Element sag is 6 cm

Half-Element Lengths (100 mm overlap)*

Element 7	3450 mm
Element 8	3082 mm
Element 9	2957 mm

Elements 10, 11, 12 and 13



Half-element wind area is 0.059 m^2
Half-element weight is 1.1 kg
Element sag is 3.5 cm

Half-Element Lengths (100 mm overlap)*

Element 10	2845 mm
Element 11	2583 mm
Element 12	2364 mm
Element 13	2264 mm

Total Antenna Wind Area and Weight
Wind area is 2.20 m^2; Weight is 44.55 kg

Note
*Tubing sizes are given as length × diameter × wall thickness, in millimeters. Lengths are for tapered elements, considering the boom and mounting plates; they differ slightly from those shown in Tables 1 through 3. All tips should overlap larger tubes by 100 mm.

Table 6—A Comparison of Similarly Modeled Commercially Made Monoband Yagis

Model No.	Gain (dBi)	SWR at Band Ends	Description
310-08	7.17	1.31-1.46	3-el 10 m on 2.3-m boom
103BA	7.51	1.53-2.01	3-el 10 m on 2.3-m boom
153BA	7.68	1.45-1.68	3-el 15 m on 3.5-m boom
315-12	7.54	1.49-1.45	3-el 15 m on 3.6-m boom
320-16	7.21	1.27-1.38	3-el 20 m on 4.7-m boom
203BA	7.17	1.22-1.20	3-el 20 m on 4.8-m boom
20-3CD	8.09	2.03-2.90	3-el 20 m on 6.0-m boom
10-4CD	8.58	1.63-1.79	4-el 10 m on 4.8-m boom
412-15	8.40	1.09-1.09	4-el 12 m on 4.4-m boom
415-18	8.24	1.41-1.38	4-el 15 m on 5.4-m boom
417-20	8.52	1.08-1.11	4-el 17 m on 6.0-m boom
204CA	8.25	1.49-1.47	4-el 20 m on 7.8-m boom
420-26	8.60	1.28-1.37	4-el 20 m on 7.8-m boom
20-4CD	8.54	1.78-2.20	4-el 20 m on 9.6-m boom
510-20	9.75	1.49-1.53	5-el 10 m on 6.0-m boom
KLM510	9.21	1.33-1.43	5-el 10 m on 6.1-m boom
105CA	8.38	1.42-1.23	5-el 10 m on 7.2-m boom
155CA	9.70	1.49-1.62	5-el 15 m on 7.7-m boom
205CA	9.23	1.43-1.96	5-el 20 m on 10.4-m boom
KLM520	9.43	1.66-1.25	5-el 20 m on 12.8-m boom

Table 7—Gain versus Elevation Angle of ON4ANT Yagi*

Frequency (MHz)	Gain (dBi)	Elevation Angle
14.150	13.55	12°
18.118	13.64	10°
21.200	13.74	8°
24.940	14.20	7°
28.400	13.77	6°

*Antenna 24 meters above ground level

tion. The actual difference in element lengths between the isolated and non-isolated construction is only a few millimeters for the 20-meter elements; it was ignored.

One needs to use tapered elements to make the strongest element with minimum weight and wind load. Most of the antenna-design programs permit element tapering; but only a few can actually calculate the element strength.

Initially, *STRESS* was used. The former hy-gain/Telex company used this one. I finally used *YAGI-DESIGN* by ON4UN. This powerful package allows one to calculate—in all circumstances—the taper and wind survival for least weight and element sag. The wind survival of this antenna is 160 km/h. The antenna is located on top of a hill.

Table 4 gives wind-load parameters. Table 5 gives element length, dia-meter and wall thickness, as well as half-element wind area and weight. The final element length is adjusted with the smallest element diameter. All 20-meter elements have three diameters.

If one chooses variant 1, deleting a 20-meter element (first-director, 4.9 m) reduces the antenna area by 0.26 square meters and weight by 5.2 kg from that with the 50-foot boom. The total weight of this antenna depends on the boom chosen, as well. My homebrew Yagi uses a four-inch boom; its total weight is a about 60 kg.

Performance Comparisons

Is it actually worthwhile to build an antenna yourself? With respect to cost, yes. The actual cost is less than $1000. The design is non-critical and can be reproduced. An antenna of this size, however, requires a strong tower and a big rotator. A wind area of more than two square meters can't be turned with a small rotator. If you have the tower and rotator, go for it.

A comparison with commercial monoband antennas available gives a good idea of what can be expected. See Table 6. The values shown for these antennas are not those claimed by the manufacturer but rather those calculated with the same software used for this design. Only in this way can we compare performance.

A gain simulation was done over real ground, with the antenna at 80 feet. See Table 7. Note the nearly constant antenna gain at low radiation angles.

Conclusion

Considering gain and SWR bandwidth, this design is a valuable alternative to a four-element monoband antenna. Of course, several improvements are still possible; those would be a function of your specific needs. I want a wide-bandwidth antenna and so did not go for the last half decibel of gain.

Are the 60-50-40-foot booms out of your reach? A possible solution might split the elements onto two antennas with shorter booms and stack them. One boom could carry the 17- and 20-meter beams, while the other would carry those for 10 through 15 meters. The major disadvantage would be changes in radiation pattern. High side lobes will increase; F/B and gain will decrease.

If you need more information about this antenna, e-mail me at the address on the title page. Also, visit the W4RNL Web site (see Note 3). You can find a lot of valuable antenna information there.

Notes

[1]*AO* and *YO* are written by Brian Beezley, K6STI.

[2]*NEC4* is commercialized under different names. It was only recently released for use by US citizens only. The Lawrence Livermore Lab holds the *NEC4* copyright.

[3]Comparison of my design in *EZNEC4* and *AO* shows that the results are very close. L. B. Cebik, W4RNL, checked this. His Web site carries plenty of antenna-related information, all free of charge. The URL is **www.cebik.com**. L. B. has been extremely helpful in checking the design. His valuable tips helped me get the last few things straight. There's a description of work on this antenna there; look for "Three Forward-Stagger 5-Band Yagis from ON4ANT."

[4]*YAGI DESIGN* is a Yagi antenna design program developed by ON4UN. It handles all mechanical issues as well as electrical characteristics. This DOS-based program is extremely easy to use. Please contact the author for full information and support on his program.

[5]*STRESS* was obtained from Mr Cox at hy-gain/Telex. My sincere thanks.

Johan was first licensed in 1985. He is mainly active on VHF and UHF, moonbounce and meteor scatter. Since 6 meters has opened in Europe, his activity has shifted toward 50 MHz. Nowadays it's 50 MHz and HF RTTY, including contesting with the OT9E team.

Having owned commercially made Yagis for years, and the contest crew "burning" them, Johan made the choice to build something inexpensive, but good. Thanks to his brother, ON4GG, this paper design turned into a real antenna. The design has been modified again for the specific needs of some radio-club friends.

Johan is an electronic engineer in telecommunications. He has been in electronics sales and now handles the customer-service department of a UHF network operator in Belgium.

A Giant LF Loopstick

Here is a LF antenna design offering good sensitivity, selectivity and directivity in the frequency range 70-340 kHz. It covers the "lowfer" band at 160-190 kHz, the new European 136-kHz amateur band, European and North African long-wave AM broadcast stations, several time- and frequency-standard stations and much more.

By Richard Q. Marris, G2BZQ

Many readers may have a modern-day radio for reception of medium- and maybe long-wavelength AM broadcast stations. It will probably have a built in ferrite loop-stick antenna a few-inches-long, perhaps 1/4 or 1/3-inch in diameter. The rod is almost certainly a nickel-zinc ferrite mixture. In contrast, my giant loopstick uses a monster ferrite rod to provide maximum sensitivity, selectivity and directivity.

"Yes," you may say, "but where can I buy the specified 12-inch-long by 1.125-inch-diameter nickel-zinc ferrite rod?" Of course, you cannot buy such a monster rod unless you are prepared to pay a phenomenal sum of money to a sympathetic manufacturer who is prepared to make a "one-off." On the other hand, why not make your own 12×1.125-inch-diameter ferrite rod? It is relatively simple!

35 Kingswood House, Farnham Rd
Slough, Berks SL2 1DA
England

Schematic Diagram

Refer to Fig 1. The Giant consists of a 11-inch-long winding on the fabricated ferrite core. In a balanced circuit, it is tuned by a double-gang, air-dielectric variable capacitor of good quality (C1A + C1B). Each gang is 500 pF. After much experimenting, a simple 50-Ω coaxial coupling was evolved via C2, which is a 470-pF silver-mica capacitor. The whole unit is constructed very simply (see Figs 2, 3, 4, 5, and 6).

Why use such a large ferrite rod? Well, experiments done over the years have shown that a large-diameter long rod provides substantially greater selectivity, sensitivity and directivity than a small-diameter short rod, as those used in transistor radios. With a sensitive LF receiver at my location, a preamplifier is not necessary and only produces unwelcome intermodulation distortion. More on this later.

Ferrite Rod Construction

The 12×1.125-inch ferrite rod consists of six 12×0.375-inch rods cemented in a circle as shown in Fig 3. As shown in Fig 2, a single 12×0.375-inch rod can be made from two or more shorter rods, secured end-to-end with Super Glue or a similar cyanoacrylate adhesive. Before gluing, lightly clean the rod ends with abrasive paper to remove any grease and roughen the surface slightly. The long rods can be made from four three-inch rods of type-61 material (available from Amidon) or from two 160×9-mm MMG grade-F14 rods.[1] Ferrite rods can be cut with a small hacksaw—using a new blade—if other lengths are available.

[1]Amidon Associates, Inc, 240 & 250 Briggs Ave, Costa Mesa, CA 92626; tel 714-850-4660, fax 714-850-1163; sales@amidoncorp.com; www.amidoncorp.com/. The rods are #R61-037-300. MMG-North America, 126 Pennsylvania Ave, Paterson, NJ 07503; tel 973-345-8900, fax 973-345-1172; sales@mmgna.com; www.mmgna.com. MMG makes rods in metric sizes; their closest product is a 9×160 mm rod, #37-0916-59 (59 is *not* their mix designator; order mix F14).

Wear rubber kitchen gloves for the whole assembly process. Six 12-inch rods are assembled with a center rod of either plastic or wood, as shown in Fig 3. (Tests indicate that using a ferrite center rod provides no noticeable performance advantage.) Hold the rods together temporarily with strong elastic bands. Next, flow the rapid-setting adhesive into and along the "valleys" between the rods. Within minutes, a solid 1.125-inch-diameter rod has been produced. Allow the glue to cure for 24 hours to provide maximum strength; then cut the rubber bands.

Fabricating and Mounting the Coil Assembly

A 12-inch length of 1.25-inch-OD, thin-wall cardboard tubing is required (see Fig 4). This diameter is often used in rolls of various kitchen foils or plastic films. Make certain that the ferrite core assembly will slide into this tubing.

For the coil, #24 (AWG or SWG) enameled copper wire was used. Winding the coil is quite a long, tedious job. Starting 1/2-inch from one end of the cardboard tubing, close-wind an 11-inch coil, finishing 1/2-inch from the other end of the tubing. Tape the ends of the winding in place, then solder a lead of insulated hook-up wire to each end of the coil. Next, insert the ferrite core into the coil. Wind a few turns of masking tape on the two ends and around the center of the core to make this a close fit.

Cut a wooden coil-mounting base, 12×1¼×½-inches, using well seasoned timber. At the extreme ends, mount narrow, plastic-coated "Terry" clips, as shown in Fig 6. Mount the coil assembly in the clips, ensuring that the coil is not trapped or disturbed. No doubt, plastic **P** clips could be used in lieu of those shown.

Final Assembly

The mounted coil assembly is fitted onto a plastic box that houses the variable capacitor and the coaxial connector, as shown in Fig 6. The box size is not critical; an 8×5×3-inch freezer/microwave box was used on the prototype. Any convenient plastic box of similar dimensions will suffice. The box lid forms the base of the unit.

The variable capacitor should have two 500-pF assemblies. Large, rigid air-dielectric capacitors are best. The one I used was salvaged from an old vacuum-tube receiver. It should be mounted at the center of one box side, as shown, and fitted with a shaft coupler, six-inch insulated shaft and a large instrument knob of 2½- or 3-inch diameter.

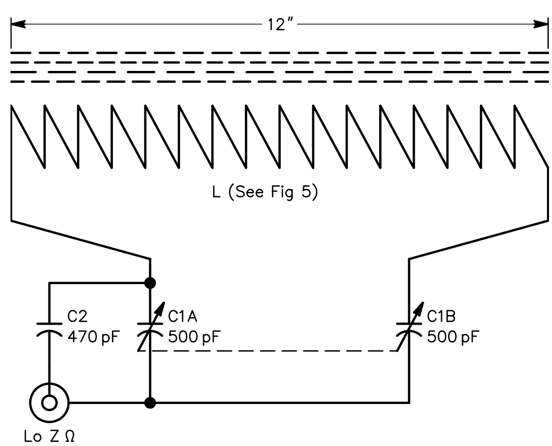


Fig 1—A schematic of the antenna and circuitry.

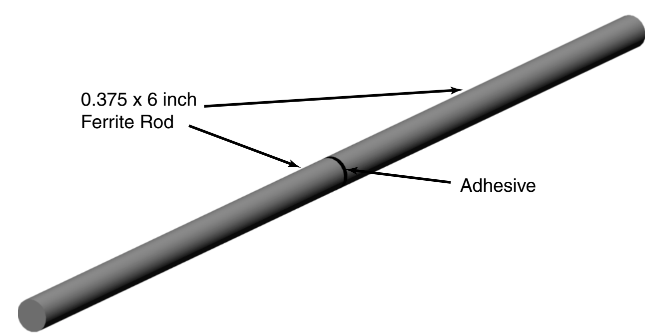


Fig 2—Fabrication of a 0.375×12-inch ferrite dowel from shorter dowels.

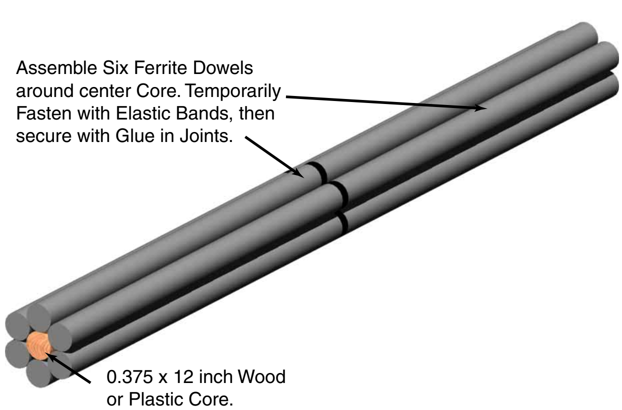


Fig 3—Fabrication of a 1.125-inch-diameter core bundle from a non-ferrous core and six ferrite dowels.

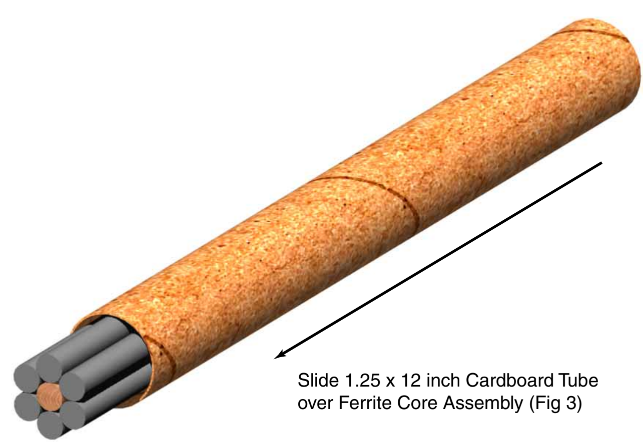


Fig 4—A 1.25-inch thin-wall cardboard tube from a box of kitchen wrap is used as a sleeve over the core bundle.

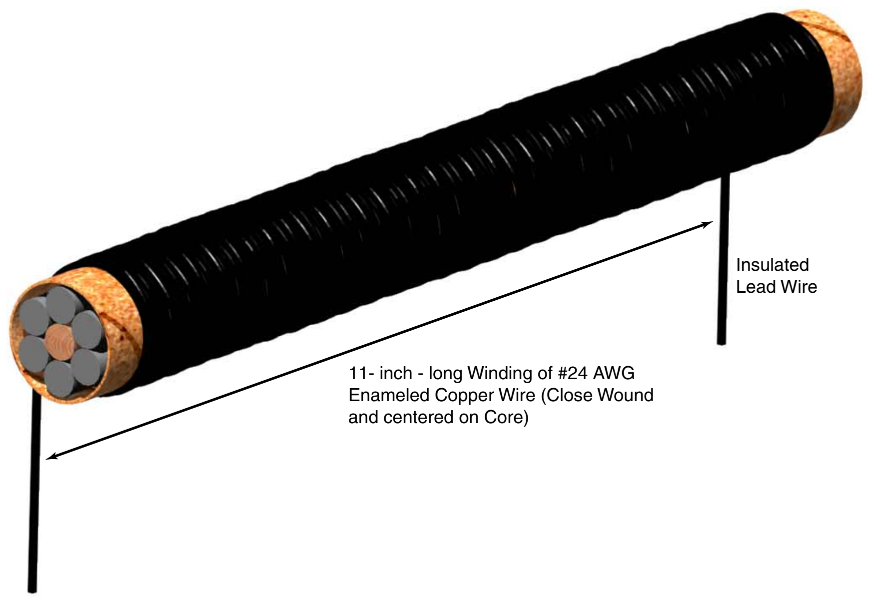


Fig 5—L is an 11-inch-long, close-wound coil of #24 enameled wire on the core bundle. Attach insulated wire leads to the coil ends.

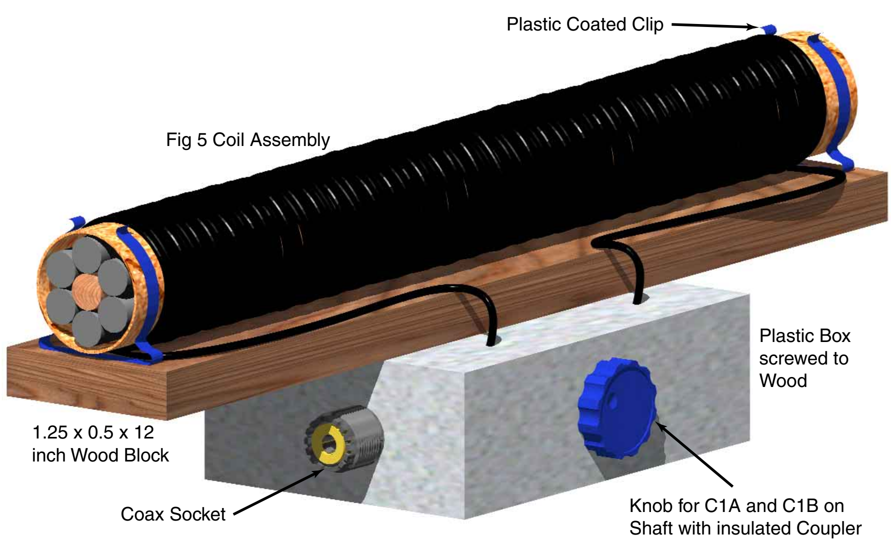


Fig 6—The coil is clipped to a wooden base that is mounted on a plastic box. The box houses C1 (A and B); the box size is not important, so long as it can hold the capacitor (see text).

Fig 7—G2BZQ's completed Giant LF loopstick antenna.

The coil assembly is fastened to the plastic box with two wood screws. The coil leads pass through two holes in the box to C1. The small amount of wiring is best done with #18 or #16 tinned copper wire to maintain absolute rigidity.

Testing and Operation

Connect the loopstick to the receiver with a short length of RG-58 coax. Initially, the receiver should be tuned to a signal around 150 kHz. The loopstick's **TUNE** knob should be rotated for maximum signal. Then rotate the loopstick slowly and a distinct peak will be noted when the antenna is on a direct bearing to the transmitter. If you rotate the loopstick another 90°, the signal should no longer be heard, unless it is a nearby, multi-kilowatt station. Repeat the process at both ends of the frequency range. For my prototype, that is 70 to 340 kHz.

I use a Palomar VLF converter and a "souped-up" ham-band receiver that covers the 80-meter band. With this arrangement, no preamplifier is needed between the loopstick and receiver. In fact, a preamplifier con-stitutes a liability, as it introduces intermodulation distortion by over-loading the converter. A less-sensitive receiver system might require a preamplifier.

The loopstick is very sensitive. With its narrow bandwidth and excellent directivity, it eliminates or greatly reduces interference and noise. It pulls in signals that cannot be heard or are lost in noise with other antennas. It gives excellent service. It even outperforms a 48×48-inch box loop (multiturn), which is also excellent. The signal level is nearly equal to that with the box loop, but the noise level is much less. I believe that the box loop can inductively couple to the house electrical wiring when pointed in certain directions.

Why Use a Giant Loopstick?

Why not use a dipole, or some other wire antenna? Well, a dipole for 340 kHz would be about 1377 feet long, 3441 feet at 136 kHz and a mighty 6686 feet at 70 kHz! Even if such an antenna were practical, the noise would be horrendous! A box loop of 36×36 to 48×48 inches will give excellent results, but can introduce inductive-coupling problems. A ferrite loopstick with, say, a single 8× 0.375-inch ferrite rod will work and may be okay for general listening. The Giant loopstick, however, with its massive ferrite core, seems to outperform all others. In addition, it does not take up very much space. It can be used indoors, as a portable receiver antenna, or on holiday.

Richard Marris became 2BZQ in 1936. He is now retired after a "life-time" in the communications and electronics business.

A High-Performance Homebrew Transceiver: Part 5

A full-featured radio has many circuits associated with operator controls. This logic board uses a simple, direct method to execute those controls. This segment also covers the PTOs, frequency counter, power supply and some construction details.

By Mark Mandelkern, K5AM

Part 1 gave a general description of this transceiver, built for serious DX work and contest operating.[1] Parts 2 through 4 covered the main signal boards.[2, 3, 4] This article gives circuit details for the logic board, PTOs, frequency counter and power supply. It concludes the series describing the 40-MHz main panel. Subsequent articles will describe the three front-end panels for HF, 50 MHz and 144 MHz and will include performance measurements.

Logic Board

The K5AM homebrew transceiver was designed on a no-compromise basis with regard to performance and operating features. The result is that numerous circuits require operator adjustment and many of them interact. A general description of the logic board was given in Part 1. The board is shown here in Fig 1. The rather large circuit diagram need not be given here in its entirety; it will suffice to specify all the logical relations and to give examples of the methods used.

Requirements

Here are some transceiver features that require control by the logic board:

1. Instant, one-button PTO switching.
2. Six modes—a mode switch with only six leads.
3. Relay-switched crystal filters.
4. One-button second-PTO monitoring.
5. PTO B frequency spotting while listening to PTO A's frequency.
6. CW offset panel control with an audio monitor to hear the amount of offset.
7. Full break-in (QSK) at 50 WPM.
8. **TUNE** switch; pulse tuning or steady carrier.
9. Dual receive.
10. Secondary band selection.

In addition to the control circuitry for the features listed above, the logic board contains the T/R, R/T, PTT, key-line, standby, QSK, semi-QSK and automatic keyer circuits.

TTL Control: Is It Obsolete?

The motivation for using TTL to control this radio was discussed in Part 1 (pages 22-23). TTL stands for "transistor-transistor logic." This refers to a series of low-cost logic ICs

[1]Notes appear on page 36.

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that became available in 1972. The TTL system allows very simple and direct generation of the logic functions required to control radio circuits. The effectiveness of this system is shown by its continued use over three decades.

Modern alternatives to TTL exist. The currently very popular "PIC" system is a compact solution to control problems. However, TTL and other discrete-logic systems have not been completely superseded by PIC or other microprocessor systems—only supplemented. PICs operate with the same voltage levels as TTL, so the systems may be used in conjunction. PICs were not available in 1990 when this logic board was built. Even now, however, PICs may not be the preferred method for controlling a radio of this sort. Along with their many advantages, PICs have some disadvantages compared to TTL: PICs are more difficult to use, take longer to design, require facility in a programming language and require special equipment to physically install the program. In addition, to use PICs efficiently requires expertise in proper software-design processes.

This discussion assumes a single ham building a single radio. For mass production and team-designed equipment, or for a widely distributed amateur project with circuit boards, parts kits and preprogrammed devices, the situation is very different. The investment of time and money in a PIC controller is justified by the savings in production. In contrast, for one ham working alone in a garage workshop, using a PIC for a medium-sized project may be too involved.[5]

Along with the original 7400-series TTL, there are also many high-speed and low-power spin-offs. Thus for high-frequency and battery-operated devices, the 7400 series is indeed obsolete. For control of a line-powered radio in a home ham shack running usual amateur power levels, however, a few milliamperes won't matter, so the plain 7400 series is a good choice. The higher currents and lower impedances of this original series result in less susceptibility to RFI than some faster and low-power systems.

In summary, TTL is still a reasonable choice for controlling a homebrew transceiver, especially for someone who does not want to put down the soldering iron long enough to take a comprehensive course in programming.

Using TTL

Information on TTL and Boolean algebra can be found in various places.[6, 7] However, no extended study program is required. It is not necessary to understand much of what is inside the logic chips in order to use them, just as one need not explore the innards of a Pentium chip in order to send e-mail. Only the *function* of each chip need be known. Only four different types of TTL gates are used here and one type of data selector. If an inverter may be thought of as a 1-input **NAND** gate, then the four gate types used are all **NAND** gates: 1-input, 2-input, 3-input and 4-input. They are the 7404, 7400, 7410 and 7420, respectively. The data selector is discussed below. The function and pin-outs for each of these ICs is given in the data book.[8]

Including some **OR** gates or other types could simplify some parts of the circuit. We must consider the trade-off, though, between achieving circuit simplicity and employing a small number of different gate types. Too many different gate types in a circuit design may result in some TTL packages being 75% unused. This is especially true for this board, since it is built in eight small, self-contained sections.

It *is* necessary to acquire a working knowledge of the logical connectives: **AND**, **OR** and **NOT**; these are denoted in the formulas used here by •, + and (overline, read as "bar"). DeMorgan's Theorem is needed to manipulate the expressions —it is just common sense. For a discussion of logical rules, notation and calculations see Chapter 7 of recent *ARRL Handbooks* (Note 6).

Input Lines

The behavior of logic circuits may be contemplated in terms of inputs and outputs. The *inputs* are the settings of all the various panel controls, plus the PTT and key lines. The *outputs* are the control lines leading to all the circuits in the radio. The inputs may be whatever the operator chooses; the outputs depend on the inputs. Thus, the inputs may be thought of as independent variables and the outputs as dependent variables. This enables us to think of the logic circuits as functions—in this case, Boolean functions. Boolean algebra is used to calculate the outputs as functions of the inputs. All the rules for Boolean calculations are in the referent of Note 6. In this design, *intermediate logic lines* are specified; interface circuits convert these lines to output lines.

Some notational conventions are used here to make the formulas easier to read. Capital Latin letters are used for TTL lines. The letters are chosen as mnemonics. For clarity, a letter alone indicates a high logic state, or an *enabled* function. The panel switches, however, are *grounded-to-operate* (that is, active low). For example, dual-receive is enabled by a switch that grounds a TTL line. Thus the line to the panel is labeled $\overline{D}$, so that dual receive is enabled when line $\overline{D}$ is low. An inverter on the logic board generates line D, which is high when dual receive is enabled. All TTL inputs are listed, in enabled form, in Table 1. In

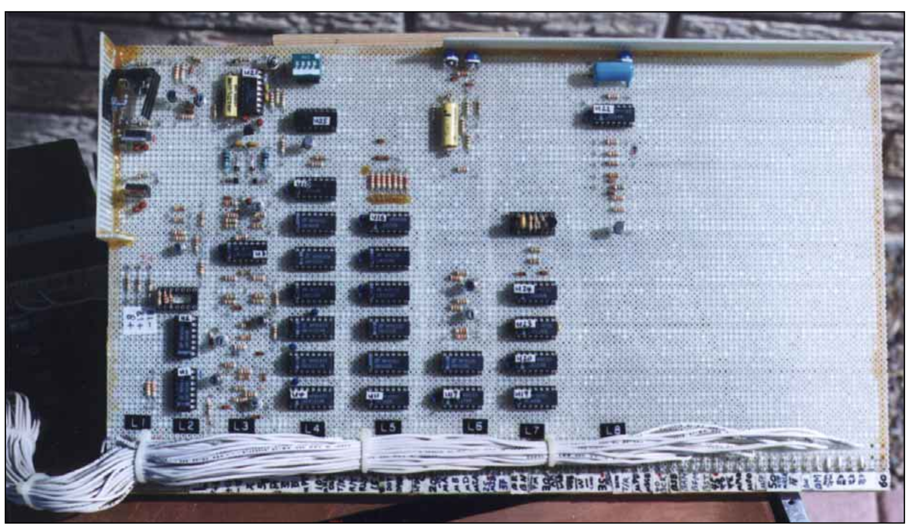


Fig 1—Top view of the logic board in the K5AM homebrew transceiver. The DIP test switch is at the top toward the left. At top center are the pulse-tuning adjustments for on/off timing. Toward the right, under the protective edging, is the QSK-delay adjustment. It keeps the transceiver in the transmit mode during the decay (break) portion of each CW element. This adjustment ensures that the CW element will not be chopped and that this delay is not excessive, so the receive circuits may recover quickly. This transceiver "hears" breaking stations while transmitting at 50 WPM.

Table 1

Logic-board input lines: This table lists TTL lines actuated by front-panel controls. The lines are denoted in *enabled* form (see text). There is no designation for the center position A/B of the PTO switch for split operation. No contact is made here—the split-operation command is deduced by the logic board. Thus, only two wires are sufficient for selecting three PTO modes.

A	PTO A; transceive
B	PTO B; transceive
D	Dual receive
M	Monitor frequency B
N	Noise blanker
S	Spot PTO B
PT	From PTT line
CWW	CW; 2-kHz bandwidth
CWN	CW; 200-Hz bandwidth
USB	Upper Sideband
LSB	Lower Sideband
AM	Carrier with two sidebands
FM	NBFM
SO	Spot CW offset
TD	Tune, pulse
TK	Tune, steady carrier

Table 2

Logic-board output lines: These lines lead to all parts of the radio, controlling all the circuits. Table 1 in Part 2 lists the Greek-letter prefixes that indicate the functionality of each type of line. For example, a µ line enables a circuit when at ground level and disables it when at –15 V. This table indicates the circuits controlled by each line.

βR	Counter (readout)
βN	Noise blanker
βM	AM/FM relay and circuits
βX	Transmit relays and circuits
µPD	Product detector gate
βAM	AM detector gate
βFM	FM detector gate
βCR	Carrier gate for CW and Tune
βSB	Sideband selection, LSB/USB
βST	Sidetone oscillator
βSSB	Transmit SSB circuits
µA	PTO A
µB	PTO B
µD	Dual receive
µMO	Master oscillator
µOO	Offset oscillator
µSO	Spot CW offset
µIA	LO injection buffer for PTO A
µIB	LO injection buffer for PTO B
T/R	–15 V receive, 0 transmit
R/T	0 receive, –15 V transmit
XMIT	Transmit order to front-end panel

Table 3

Intermediate TTL lines: These lines operate in the logical realm between the input and output lines. They are formed as Boolean functions of the inputs and applied to the interface circuits to generate the outputs.

P	PTO A
Q	PTO B
R	Counter readout; PTO B
J	Receiver-mixer LO injection; PTO A only
K	Receiver-mixer LO injection; PTO B only
L	Receiver-mixer LO injection; dual receive
X	Transmit
T	Tune, pulse or steady
OP	Operate
KL	Key-line control
CW	CW, either bandwidth
PD	Product detector
CR	Carrier
ST	Sidetone
NB	Noise blanker
MO	Master oscillator
OO	Offset oscillator
SSB	LSB/USB, transmit

Table 4

Boolean Functions: These determine the operation of all circuits in the radio.

$$CR = (CW + T) \bullet X$$
$$CW = CWW + CWN$$
$$J = P \bullet \overline{L}$$
$$K = \overline{P} \bullet \overline{L}$$
$$KL = TK + \left(X \bullet \overline{CW} \bullet \overline{TD}\right)$$
$$L = \overline{X} \bullet D \bullet \overline{M} \bullet \overline{B}$$
$$MO = \overline{(CW + T)} + \overline{X}$$
$$NB = N \bullet \overline{X}$$
$$OO = \overline{MO} + \left(SO \bullet \overline{X}\right)$$
$$P = \left(A \bullet \overline{M}\right) + (A \bullet X) + \left(\overline{A} \bullet \overline{B} \bullet \overline{M} \bullet \overline{X}\right)$$
$$PD = \overline{(AM + FM + SO)}$$
$$Q = \overline{P} + \left(\overline{X} \bullet (S + D)\right)$$
$$R = \overline{P} + \left(\overline{X} \bullet S\right)$$
$$SSB = (LSB + USB) \bullet X \bullet \overline{T}$$
$$ST = CW \bullet \overline{T}$$
$$T = TD + TK$$
$$X = (PT + T) \bullet OP$$

the discussion, we will refer only to the lines representing enabled functions, without reference to the panel lines and the inverters. Using two hex-inverter TTL packages avoids having to work with all the input lines in reversed logic. The PTT and key lines are *external* inputs, labeled αPT and αKL, respectively. The Greek-letter prefix conventions, also used in previous series segments, are listed in Table 1 of Part 2. The logic board outputs also use Greek-letter prefixes. They are listed in Table 2 here.

Intermediate Logic Lines

Generation of these signals is the central function of the logic board. Various inputs, corresponding to operator controls, are combined to form logic lines that represent command signals for the circuits. At the intermediate point, they are still TTL-level signals; additional interfacing circuits produce the actual output lines. The TTL designations of the intermediate logic lines are given in Table 3. The circuit decisions involve Boolean functions; these are listed in Table 4.

For example, the function NB = N • $\overline{X}$ (NB equals N **AND** not X) enables (high state) the noise blanker when the blanker control knob is pulled out, sending input line N high, but only when receiving. This limitation is necessary because the tunable noise-channel LO could cause a spurious output if allowed to run while transmitting. An interface circuit converts line NB to the output line βN.

As another example, pulling the dual-receive knob out sends input line D high. This turns on both PTOs—but not while transmitting! Thus the transmit line X must be combined with D. When transmitting, the PTO switch position must also be considered. When receiving, pressing the **MON** button must momentarily terminate the dual-receive function, enabling only PTO B. During split-frequency operation, the **SPOT B** switch turns on PTO B while receiving with PTO A; PTO B reads on the counter. The result is that the intermediate line Q, enabling PTO B, depends ultimately on six input lines.

Generating Boolean Functions

Only a few examples will be shown here. These may suggest ideas for using TTL on any small control problem that might arise. In Fig 2, the simple intermediate TTL lines CW and SSB are obtained with only a few gates. Somewhat more complicated are the intermediate lines CR, MO, and OO. Their circuit is shown in Fig 3. For Boolean functions with even more variables, data selectors are used. These are discussed below.

Output Lines

The intermediate logic lines must be converted to control signals that can be used by the circuits in the radio. Some of the circuits simply require a supply voltage to be turned on or off at the right times. Some circuits use dual-gate VHF MOSFETs, which require a positive voltage at the control gate to activate them, or a negative voltage to turn them off. There are also relays to control. The logical relation between an intermediate line and an output signal is simple and direct. For example, intermediate TTL line Q is high when PTO B is enabled and low otherwise. Output control line µB, derived from Q, is accordingly either 0 or –15 V. PTO B will run with 0 V on the control line, while –15 V applies cutoff bias to the oscillator and buffers.

Op amps are used to convert intermediate TTL lines to control signals. Two basic types are used here: ordinary op amps and comparators (which is, strictly speaking, not really an op amp). A regular op amp, such as LM324N, will generate a β-type control line, switching from about +15 V to –15 V. In practice, with ±15 V dc rails, only about ±13 to ±14 V is obtained. This is taken into account in the design of the circuits controlled and of the diode switches, although the full voltages are used in the discussions.

A comparator, such as the LM339N

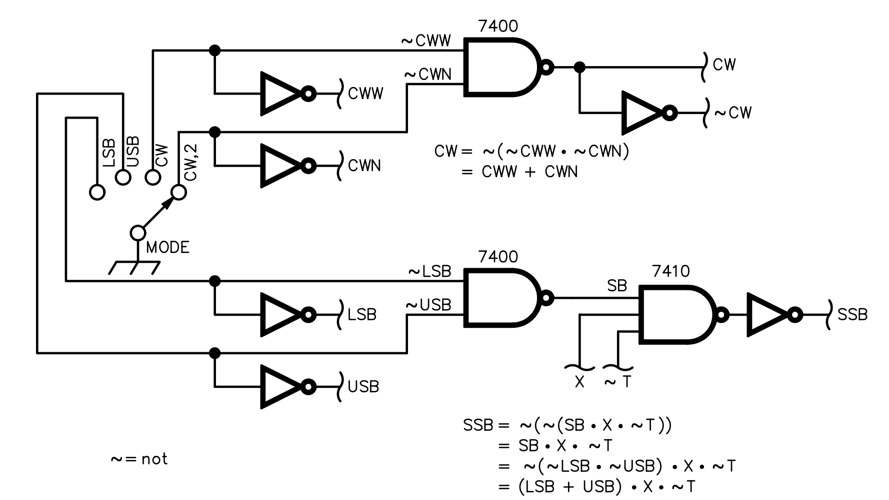


Fig 2—Intermediate TTL lines. This circuit generates the lines, CW and SSB, that control the transmit CW and SSB circuits. When using the TUNE function in SSB mode, the transmit SSB circuits are disabled. This avoids noise on the carrier while tuning. The calculations given here show how DeMorgan's Theorem and the other logic laws are used to design the circuits. The labels shown on the MODE switch are panel labels, not TTL designations. Not shown are the 2.2-kΩ pull-up resistors from the +5 V dc rail to each panel input line.

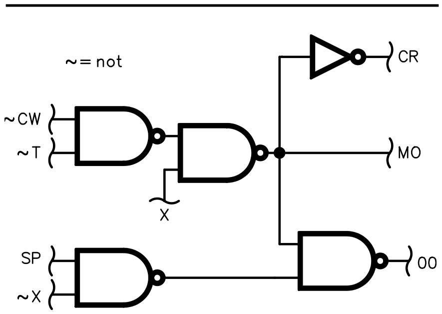


Fig 3—Intermediate TTL lines. This circuit generates the lines CR, MO and OO, for controlling the carrier, the master oscillator and the offset oscillator, respectively.

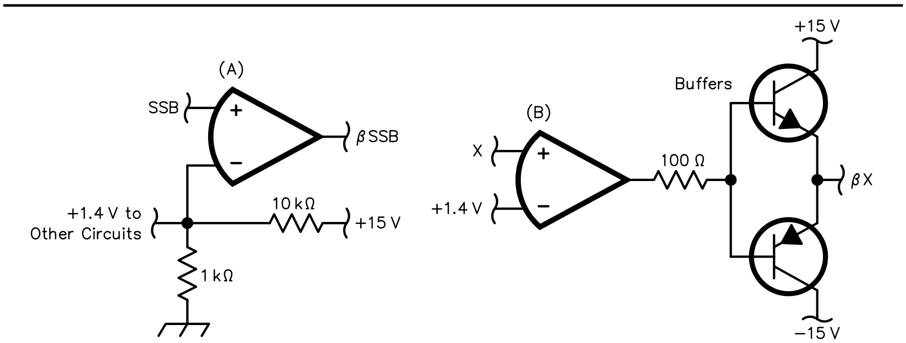


Fig 4—Interface circuits for obtaining control lines. In (A), the TTL line SSB is interfaced to obtain control line βSSB, used for controlling transmit SSB circuits. The circuit in (B) derives the control line βX from the TTL line X. Because line βX drives a number of circuits, the required current drain exceeds the capacity of the op amp, which has a maximum output of about 20 mA. With the buffer transistors added to the circuit as shown, we can draw well over 200 mA from line βX.

The schematics in this article use the following conventions: Except as noted, the op amps are LM324Ns, powered from the ±15 V dc rails. The transistors are small-signal types, such as 2N4401 (NPN) or 2N4403 (PNP); the diodes are small-signal silicon switching types such as 1N4148. Resistors are 1/4-W, carbon-film units. Trimmer potentiometers are one-turn miniature parts, such as Bourns 3386 (Digi-Key #3386F-nnn, see Note 22). Capacitors labeled "s.m." are silver micas, with values given in pF. Values of RF chokes (RFC) are given in µH. The 100-nF monolithic ceramic bypass capacitors at the power terminals of each TTL and op-amp package and 10-nF disc ceramic bypass capacitors at each board terminal are not shown. Unmarked coupling and bypass capacitors are 10-nF disc ceramics. Potentiometers labeled in all capital letters are front-panel controls; others are circuit-board trimmers for internal adjustments.

used here, has an open-collector output. With ±15 V dc rails, it provides either a –15 V output or an open circuit. This is useful for generating a μ-type control line. In later designs, however, only a regular op amp is used. This simplifies things, as a μ-type line can be generated from a β-type line merely with the addition of a diode. Thus, only the generation of β-type lines will be shown here.

One valuable characteristic of an op amp is its ability to supply moderate output current while drawing virtually no current from the inputs. Fig 4 shows the way TTL lines are converted to β-type control lines. Although we use 0 and 1 as logic symbols, the TTL circuits do not use 0 and 1 V for **FALSE** and **TRUE**, they use roughly 0 and 5 V. These voltages are only nominal; typical values might be 0.5 and 3.5 V. The exact range of permitted values is given in TTL reference books. A good intermediate value is 1.4 V. A voltage divider establishes this with only two resistors, and the reference voltage thus obtained is used throughout the logic board. We apply 1.4 V to the inverting input of the op amp and the TTL line to the noninverting input. When the TTL line shifts from about 0 V to about 4 V, the op-amp output shifts (nominally) from –15 V to +15 V. For example, if we apply the TTL line X (transmit), we obtain the output control line βX, which shifts from –15 V in receive to +15 V in transmit. One use of this βX line is to control MOSFETs in the transmitter section of the radio, with a voltage divider in the MOSFET circuit to obtain ±4 V for the control gate.

PTT and T/R Circuits

The STBY/OPER function is incorporated into the PTT circuit. Several output lines are provided for the transmit/receive function: T/R, R/T, XMIT and βX. This extra bit of circuitry simplifies design of other circuits throughout the radio. A PTT circuit is shown in Fig 5.[9]

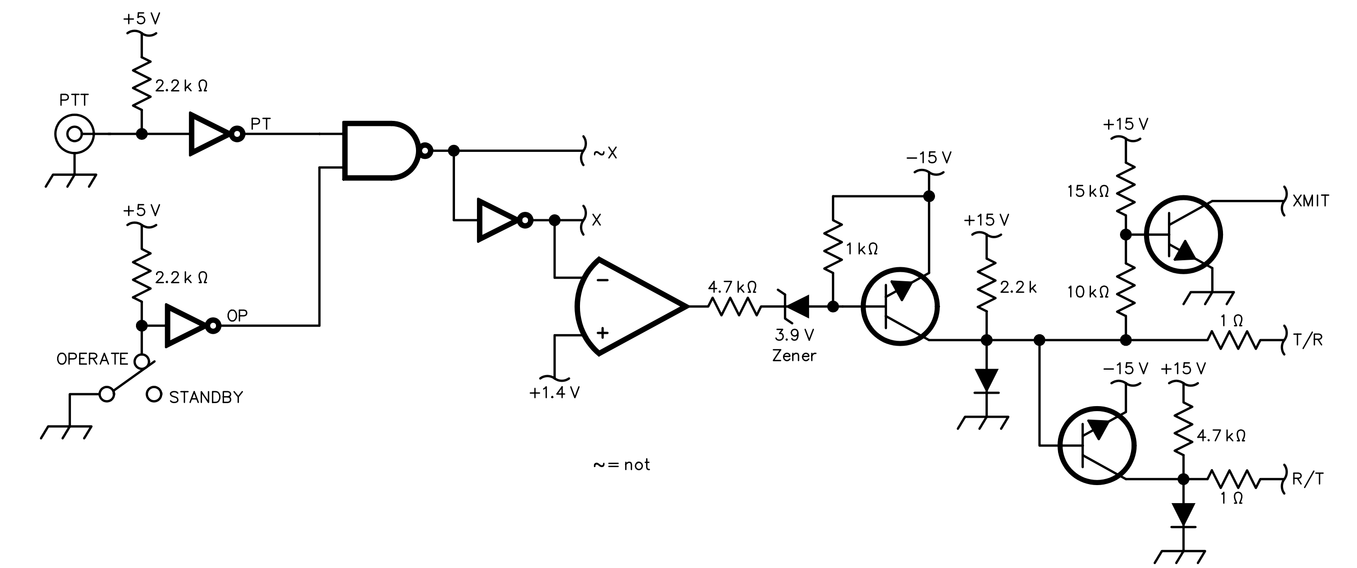


Fig 5—PTT and standby circuit schematic (see Note 9). In this circuit, the external PTT line, usually operated by a foot switch, is transformed into the TTL line X, for transmit. The STBY/OPER switch on the front panel enables the circuit. The output line XMIT conveys a transmit order to the selected front-end panel.

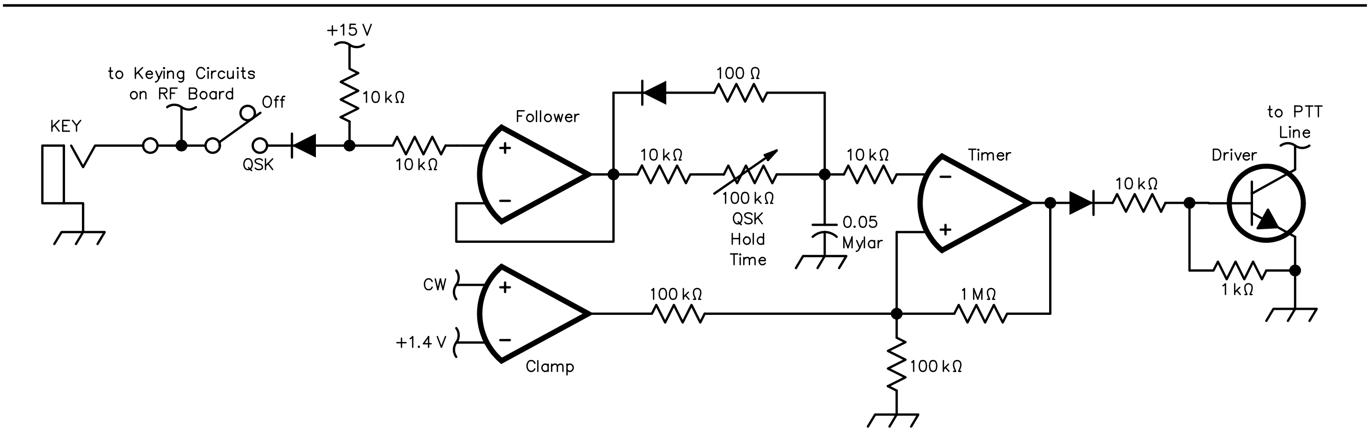


Fig 6—CW break-in (QSK) circuit (see Note 9). The timing adjustment is set for a PTT hold-in delay of about 4 ms after the end of each code element. This prevents any chopping of the decaying CW waveform. Other methods that introduce delay into the keying circuit may distort the CW waveform and cause excessive delay that degrades the QSK performance. This radio can hear breaking stations while sending at 50 WPM.

Key Line and Break-in Circuits

The interface of an external input line with internal circuits becomes critical for the keying line, since timing is important. We must have circuit isolation, otherwise capacity on the key line or from an external keyer might alter the timing and distort the keying waveform. A keying circuit is shown in Fig 6 (see Note 9). The keying waveform is developed at the 40-MHz buffers on the RF board, as shown in Fig 9 of Part 3. A Curtis 8044 keyer IC is included. It is valuable in case the external memory keyer fails.

There are two CW break-in modes. *Full break-in* (QSK) requires full receiver recovery between CW elements. This radio can hear breaking stations while sending at 50 WPM. *Semi-break-in* (SQSK) is a misnomer: It only implies an automatic PTT function. Stations cannot actually break in. SQSK is useful with an amplifier that lacks full break-in capability.

The QSK delay is set using a scope to eliminate chopping of the decay (break) portion of the CW waveform, while monitoring with a receiver to check for key clicks. Too much delay will inhibit quick receiver recovery and limit the ability of the radio to hear breaking stations between dits.

The keying section of the logic board includes the pulse-tuning circuit. The advantages of pulse tuning and the circuit details were described in a previous article.[10] To reiterate: The chief advantage is greatly reduced anode dissipation in the kilowatt-amplifier tubes. The circuit has on/off timing adjustments, so the pulse width and duty cycle may be adjusted. Settings for 13-ms on and 27-ms off provide a 33% duty cycle with a pulse rate corresponding to shortened CW dits at 60 WPM.

PTO Control

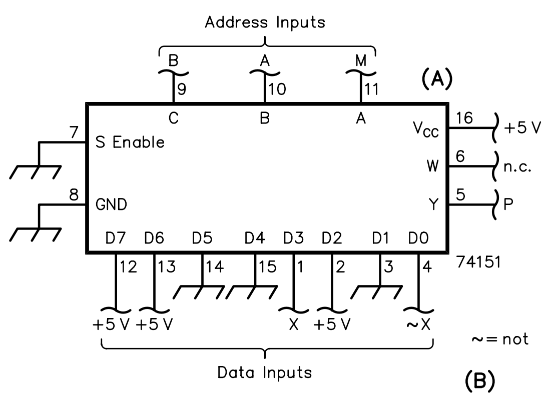
For the utmost in operating flexibility, PTO control is of special importance. This radio has two PTOs and dual-receive capability. In addition, it provides instant, one-button monitoring of the second frequency for split-frequency DX operations. This feature is useful because dual receive is not feasible with extremely weak, barely readable DX signals. There is also provision for spotting the transmit frequency while receiving, as for 40-meter DX SSB work. In any given situation, the logic board must make several decisions. Which PTOs should run? Which PTO should determine the receiver's mixer injection? Which PTO should be displayed?

Data Selectors

Simple gates are used for most of the intermediate TTL functions. For the more-complicated functions, such as PTO control, 74151 *data selectors* are used as Boolean function generators. Complicated Boolean functions would require a large number of simple gates; a single data selector can often do the same job. The term *data selector* refers to the primary use of the device: to choose among several data inputs as determined by the address inputs. Here we employ its secondary use: as a *Boolean-function generator*. We may think of the selected data as the desired result, based on the operator's orders as applied to the address inputs. Three 74151 data selectors are used here to control PTO A, PTO B and counter display selection. Each 74151 is a 16-pin DIP device with three address inputs and eight data inputs.

The Boolean functions for the PTOs must have inputs involving manual PTO selection, dual receive, monitoring of the second PTO, spotting a frequency on the second PTO and the transmit/receive condition. As an example, the circuit for TTL line P, which determines power to PTO A, is shown in Fig 7. We have four variables, A, B, M and X, corresponding to the panel switches for **PTO A**, **PTO B**, **MON** and the transmit condition, respectively. Since the data selector has only three address inputs, a fourth input is effected by using the data inputs. These four variables together result in $2^4 = 16$ logic states. The desired output for each of these states is selected by configuring the appropriate data-input pin.

The three address inputs result in eight possible states, represented by numbers in binary form: 000_2 to 111_2, or 0_{10} to 7_{10}. For each of these states, output signal Y at pin 5 is determined by the level of the corresponding data-input pin. The data input pins are labeled D0 through D7. To generate P, the address inputs are M, A and B; these correspond to the **MON**, **PTO A** and **PTO B** switches. Note that three different labeling systems are employed! The circuit lines are M, A and B; the



(B)

Panel Switch	Address Input Lines			Input State		X	P	Data Input Line	Pin Number	Selected Connection
	B	A	M	Binary	Base 10					
A/B	0	0	0	000	Zero	0 / 1	1 / 0	D0	4	~X
	0	0	1	001	One		0	D1	3	⏚
A	0	1	0	010	Two		1	D2	2	+5
	0	1	1	011	Three	0 / 1	0 / 1	D3	1	X
B	1	0	0	100	Four		0	D4	15	⏚
	1	0	1	101	Five		0	D5	14	⏚

Fig 7—Data-selector logic diagram. The schematic at (A) shows the data selector used to generate the line P, which determines when PTO A runs. The 74151 data selector is used as a Boolean function generator. This method is simpler than using a large number of separate gates. The truth table for this Boolean function is shown at (B). The table is incomplete; there are no entries for the input states 110_2 and 111_2. These states occur when the operator throws the PTO switch in both directions at once!

chip manufacturer denotes the address inputs at the chip by A, B and C; the socket pins are numbered 11, 10 and 9. There is also a fourth variable in the Boolean function—namely X—and thus there are 16 possible input states. By applying X or $\overline{X}$ as data input when needed, we effectively obtain a 16-input device.

For example, in the state $\overline{B}$•A•M, we are in PTO-A mode while monitoring the B frequency. This state is represented by the binary number 011_2, or three; the corresponding data input terminal is labeled D3. Should PTO A run? If we are receiving, the answer is "no," because we are listening to the B frequency. However, "PTO A transceive" is selected on the panel, so when we begin transmitting, we should revert to PTO A, and the answer is "yes." We introduce this fourth variable by applying the transmit line X at data input terminal D3 (pin 5). Thus, during the input state 011_2, the output line, P, will follow the transmit line, X, as required. In this way, the eight-input data selector, which is a Boolean function generator for only three variables, handles four variables. Using the *enable* signal (pin 7), application of the 74151 may be extended even further in some cases.

Each of the 16 possible states determined by the four input variables is handled similarly. The first three variables determine eight states. For each of these states, the desired output is obtained by connecting one of four lines to the corresponding data input pin: +5 V dc, ground, X, or $\overline{X}$.

Note the absence of an input line for split operation. When the panel switch is in the center, A/B, position, both A and B lines are low; the logic board easily infers the split order from this situation. Thus at the data selector for line P, split operation is ordered in either of the address-input states 000_2 or 001_2: zero or one.

Mode-Switching Circuits

The use of logic circuits for mode switching yields many advantages, as discussed in Part 1, pp 22-23. A portion of the mode-switching circuit is shown in Fig 8. This circuit enables the sharp (200-Hz) CW filter when requested by the mode switch. Most of the power-supply load is on the +15 V dc rail, so certain circuits are run from the –15 V dc rail in an effort to distribute the load more evenly. The special sharp CW-filter preamplifier and the CWN relays present good opportunities for this. The special –15CWN line is at once a control line and a power source. The buffer circuit used to obtain more current for the βX control line in Fig 4 drops the available op-amp output voltage by an additional V_{be}, and so is not appropriate here. The circuit used for –15CWN, shown in Fig 8, provides nearly the full –15 V.

Secondary Band Selection

This radio has a special band-switching feature. Although quite meager compared to the many memories and flexibility of a computer controlled radio, it is very useful in many situations. The panel switch labeled **BAND** controls a line leading to the external front-end switch box, which has switches for choosing a primary and secondary band. The **BAND** switch on the main transceiver panel has the following functions: In the center position, no secondary band is selected; in the up position, whenever the radio switches to PTO B the secondary front-end panel is selected; in the down position, the secondary front-end is selected in all PTO modes.

The **BAND** switch has many uses. For example, one can be busy working an HF contest and set the thing up so that one touch of the **MON** button checks the 50.110-MHz DX calling frequency. Or, you can be working 6-meter DX and set

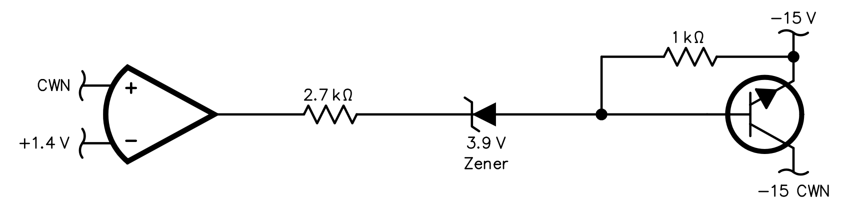


Fig 8—Circuit for controlling the sharp (200-Hz) CW filter. A simple NPN switch is used to provide current for the special sharp-CW preamplifier and the crystal-filter relays.

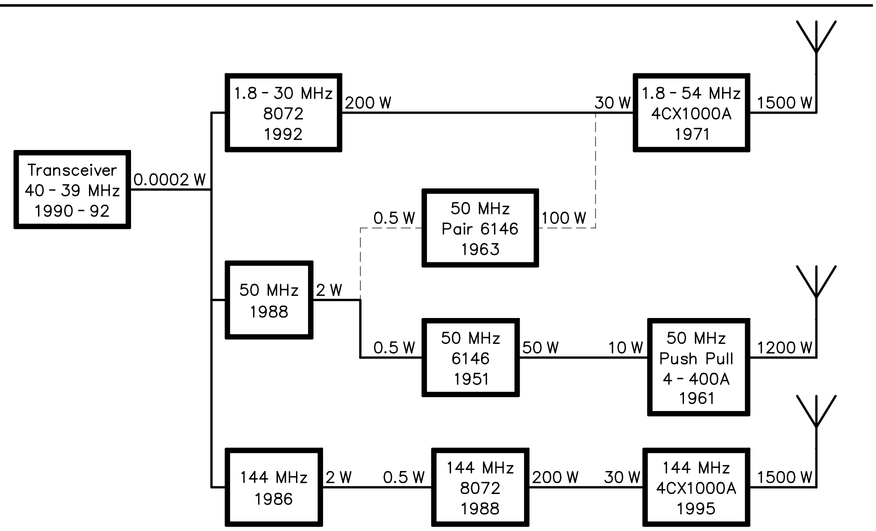


Fig 9—K5AM homebrew station layout. This shows how the main transceiver panel and the three front-end panels fit into the station. To obtain very high dynamic range and minimize spurious responses, the station uses only two mixing conversions on each band from 1.8 to 144 MHz.

To obtain the cleanest signal possible, the station uses class-A transistor stages up to 2 W, and then only tubes—class-AB_1 tetrodes. The diagram indicates the power (in watts) available from each unit and the power required by the next. The resulting headroom yields the best IMD performance. Gain is controlled at the milliwatt level in each front-end. To prevent splatter, ALC runs from each driver and each high-power amplifier back to the corresponding front-end panel, with ALC metering at the transceiver. There are no diodes in the signal path at any point in the station.

The 8072 is a conduction-cooled tetrode—identical to the well-known 8122 except that the 8072, with no air-cooling fins, clamps to a heat sink. The neutralized 8072 driver amplifier for 2 meters has 26 dB gain. The Eimac 4-400A bottles used in push-pull on 6 meters are the originals—only 39 years old and still running at full output!

up so that switching to PTO B immediately puts you on 28.885, the 6-meter DX liaison frequency. With some 6-meter DX openings lasting only 10 seconds, every second counts!

Rear-Panel Connections

Years of trouble with phono plugs and jacks led me to look for a better way to connect the radio to the rest of the station. Hams commonly use BNC connectors only for RF. Where reliability is important, however, it is common to use BNCs for control lines also. For contest work, I also want the highest level of reliability. Thus, I use BNCs for all rear-panel jacks that would normally use phono jacks.[11] All control connections at the rear panel are filtered—mostly with pi-sections using a 1-mH RFC and two 10-nF disc ceramic capacitors.

To facilitate interconnections at the operating bench, a dual-connector system is employed. In addition to the BNC and key jacks, a seven-pin DIN connector connects the PTT line, key line, keyer dit/dah lines and transmit control line to the separate front-end panels with one quick push. It also connects the ALC-meter lines from the front-end panels back to the main transceiver panel. This cable leads to a station-hub switch box on the operating bench that selects the front-end panel: HF, 50 MHz or 144 MHz. Each of the three panels also connects to the station hub with a single control cable and a coaxial cable for 40 MHz. Pulling the radio for a quick trip to the workbench to add a new feature (between contests) is an easy task. There is also a DIN jack on the radio for connecting a front-end panel directly to the radio, so the radio can be used independently of the hub or tested with one front-end panel at the workbench.[12]

Each front-end panel has a similar dual-connection arrangement. All this redundancy in connections required only a few extra hours of work, repaid many times since in convenience. The transceiver and front-end panels fit into the complete station layout as shown in Fig 9.

DIP Test Switch

Many test and alignment procedures require certain oscillators to be turned off. For example, sweep alignment of a filter following a mixer is impossible with the LO running and reacting with the sweep generator. For convenience, a four-gang DIP test switch is installed on the board. It can defeat MO, OO, PTO A or PTO B. The switch is positioned at the top of the board so it is easily reached with the logic board in place.[13]

Logic Board Construction

The board itself is shown in Fig 1; its general method of construction was described in Part 1. Rather than copper circuit board, the logic board is constructed on perf board. Wiring on the board is done mostly with wire-wrap, with some point-to-point hand wiring. The board's bottom surface is shown in Fig 10. A long copper strip along the edge with 59 terminals provides a common ground and a return for bypass capacitors on all the input and output lines.

Permeability-Tuned Oscillators

In a traditional homebrew radio, the VFOs may be the most difficult problem (leaving aside modern digital methods). The VFOs are most demanding of perfection. For the ham more comfortable with soldering irons and transistors than lathes or bearings, the mechanical demands may cause the most headaches; I sidestepped the mechanical obstacles. Although I built the circuit, filter and shielded box, these details are trivial compared to the powdered-iron tuning slug, coil, precision-rolled left-hand-drive screw, bearings, tuning rail, anti-backlash mechanism and the solid frame. All these mechanical parts were stolen from a junked Signal/One CX7.[14] Even after this grand larceny, there is still much work to be done for mechanical overhaul and adjustment of these old PTOs.[15] One shudders to think what it would cost to manufacture such a precision PTO mechanism today.

PTO Circuit

The PTO schematic is shown in Fig 11. The circuit is very similar to that used in the Signal/One, except that the Signal/One included no RIT. The circuit may also be used for a variable-capacitor-tuned VFO. The capacitor's reduction-gear tuning mechanism will be arranged by the individual builder. Because a frequency counter is used, the difficult problem of dial calibration is eliminated and linear tuning is not essential.

This PTO tunes about 25 kHz per revolution—no problem for old-timers, although faster than modern tastes would demand. There are several possible ways to reduce the tuning rate. One is to rewind the coil with a coarser pitch at the low end for the CW segments.[16] Another is to rewind the coil to reduce the range to 500 kHz, adding extra 10-meter segments to the HF front-end panel as needed.

The Case for RIT

An RIT function is essential in most operating situations. Arguments against it—while logically perfect—apply only to a perfect world. If every operator tuned his radio perfectly, RIT would not be necessary. While RIT is the solution for the listening operator, ironically it is RIT itself that often causes mistuning by the transmitting

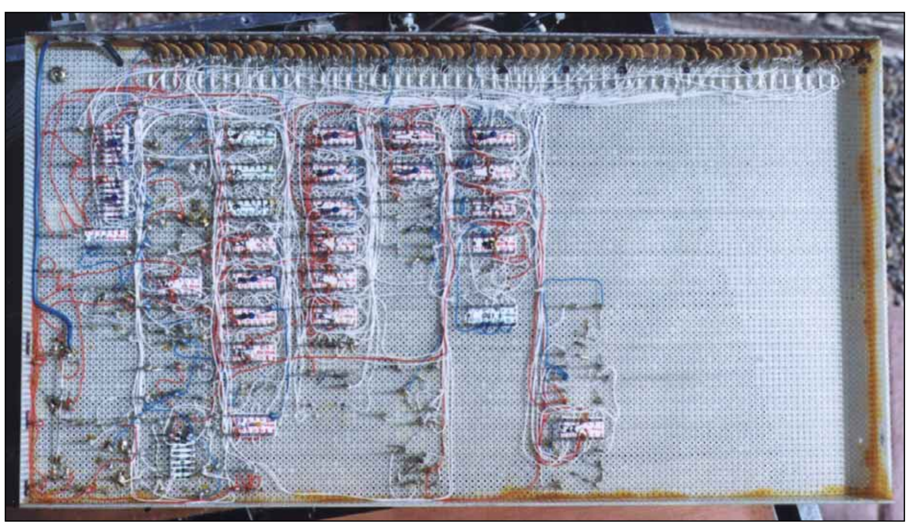


Fig 10—Bottom view of the logic board. Wire-wrap construction is used for the TTL and op-amp circuits and point-to-point wiring for the circuits with discrete components. To minimize connector troubles, the board is hard-wired to the radio. A 12-inch-long bundle of wires allows the board to be easily lifted and serviced.

operator. The problem is a general lack of awareness of the relation between the CQ-calling station and the replying station; that is, the requirement that the replying station tune with the RIT *off*. New operators might be advised simply: "Don't use the RIT until you acquire more experience."

In any event, a contest operator must have RIT to hear the offenders. There are also other valid uses for RIT. During a CW QSO, one might wish to change the received pitch to improve readability in QRN or avoid QRM. RIT is especially useful for moon-bounce work. After acquiring an EME signal and making the first call, I pull out the RIT knob. This fixes my transmit frequency while allowing me to adjust tuning as desired at a slower rate. The RIT is also used to adjust for Doppler shift. As always, the rule is: Never change the transmit frequency after a QSO has begun, or even after a first call.

RIT Circuit

The RIT system uses a simple transistor switch. Let's say the RIT is on. While receiving, the T/R line is at –15 V and keeps the transistor cut off, so the *Xmit-Set* trimmer has no effect; the varactor diode bias is varied by the RIT panel control. When transmitting, the T/R line is near ground; there is very little voltage at the RIT control, and it has no effect. Now, the transistor is turned on, so the *Xmit-Set* trimmer determines the varactor diode bias, setting the PTO frequency to the center of the RIT range. When the RIT is switched off, the transistor is always on, so the frequency is centered.

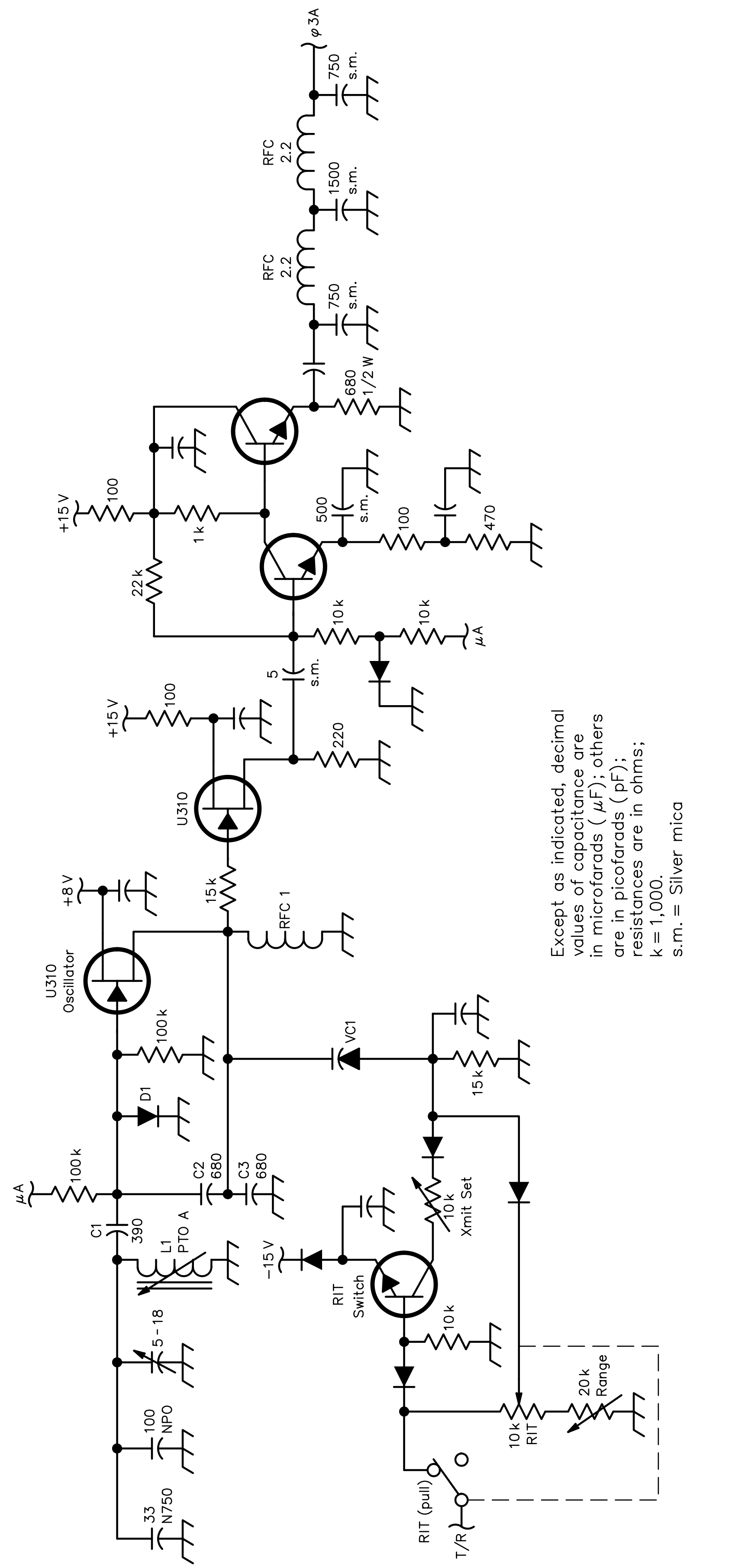


Fig 11—PTO schematic diagram. The 78L08 regulator that powers the oscillator is not shown.
C1—Temperature-stable monolithic ceramic capacitor, type C0G, 390 pF; Panasonic #ECU-S1H391JCA, Digi-Key #P4932 (see Note 22). From 0 to 60°C, these capacitors have a tolerance range of 0.1%.
C2-C3—Temperature-stable monolithic ceramic capacitor, type C0G, 680 pF; Panasonic #ECU-S1H681JCB, Digi-Key #P4935 (see Note 22 and comment for C1).
D1—Hot-carrier diode, HP-2800. A small-signal germanium diode, such as type 1N270, may also be used.
L1—Permeability-tuned oscillator coil, salvaged from a Signal/One CX7.
RFC 1—RF choke, 100 µH. A 1-mH choke is often seen in this sort of oscillator circuit; that is larger than needed and may cause pick-up of AC hum.
VC1—Varactor diode, nominal 33 pF. Motorola type MV2109, NTE type 614 (see Note 21).

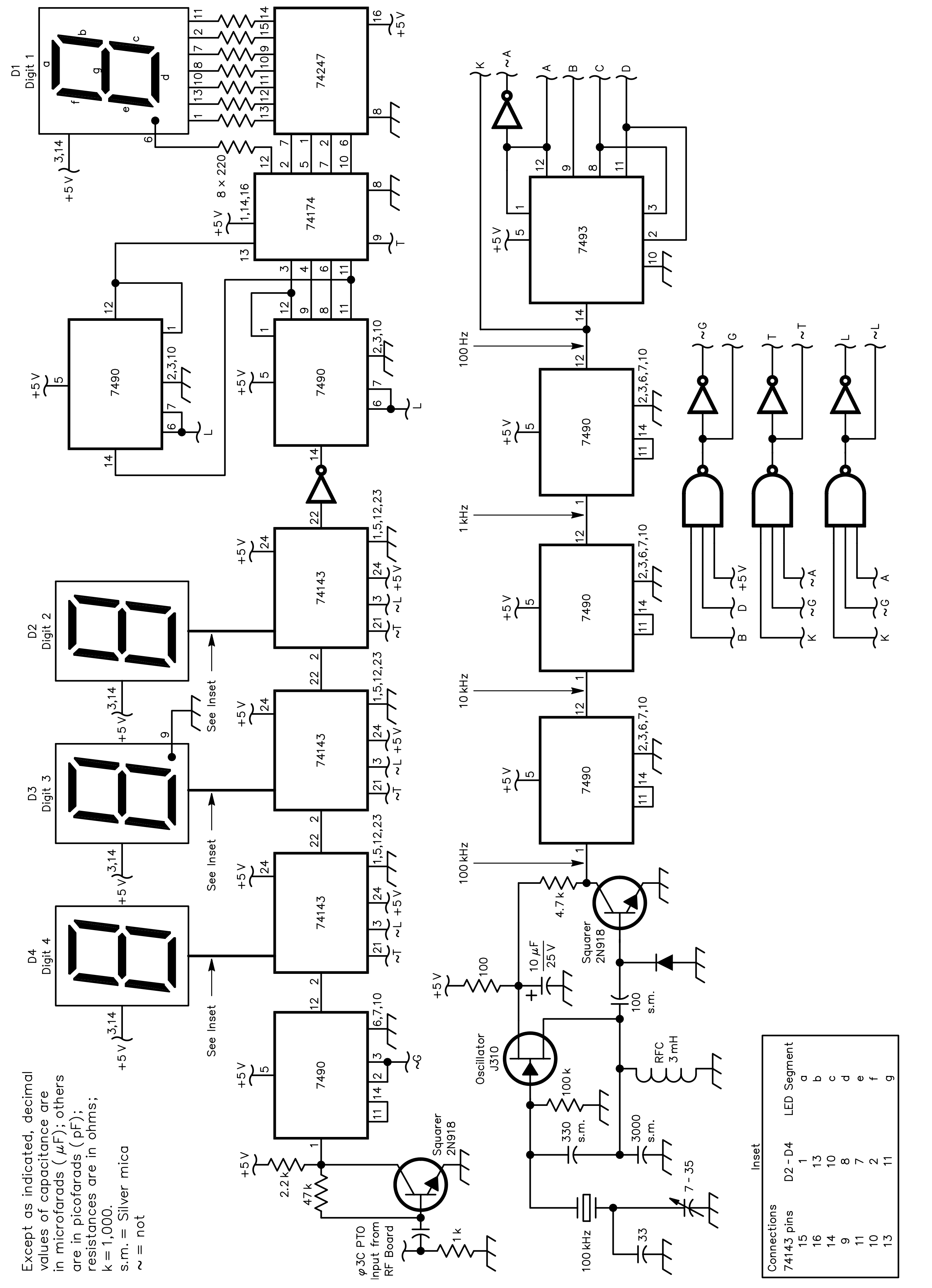

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000. s.m. = Silver mica ~ = not
D1 Digit 1
D2 Digit 2
D3 Digit 3
D4 Digit 4
See Inset
8 × 220
74247
74174
7490
74143
7493
Squarer 2N918
Oscillator J310
φ3C PTO Input from RF Board
100 kHz
10 kHz
1 kHz
100 Hz
RFC 3 mH
Inset
Connections
74143 pins / D2 - D4 / LED Segment
15 / 1 / a
16 / 13 / b
14 / 10 / c
9 / 8 / d
11 / 7 / e
10 / 2 / f
13 / 11 / g

Fig 12

RIT Adjustment

The *Range-Set* trimmer is adjusted to obtain a range of ±1 kHz, which is most convenient. Varactor diodes of the same type may vary greatly in characteristics. If the RIT control does not yield equal shift in both directions from center, resistors are selected and installed between the arm and either side of the control to center the curve. The range obtained may also vary, hence requiring adjustment of the 15-kΩ resistor. Thus, the RIT circuits are individually fine-tuned.

PTO Construction

The original Signal/One PTO cover design has several drawbacks. The cover is held by only two screws and does not fit tightly, which results in an RF-leaky enclosure. Connections to the PTO circuit are made through plain insulated terminals, not feed-through capacitors, resulting in further leakage. For these reasons, new PTO enclosures were fabricated using copper-clad circuit board. Feed-through bypass capacitors are used. All the usual CX7 spurs caused by PTO leakage are eliminated. Building the HF front-end in a separate enclosure is also a major factor in this result.

Frequency Counter

The main transceiver panel covers 40-39 MHz, driving three front-end panels for HF, 50 MHz and 144 MHz. An external switch box with push buttons chooses one of these three. The HF panel (200 W) covers the ham bands up to 30 MHz in 10 1-MHz bands. The 50-54 MHz panel (2 W) has four 1-MHz bands. It was built during solar cycle 22, when ZLs and VKs still used the 51 and 52 MHz segments. The 144 MHz panel (2 W), used only for CW/SSB DX and for moonbounce, covers only 144-145 MHz. The frequency counter on the main panel counts only kilohertz. The operator must read the band from the switch box and the band switches on the front-end panels—an archaic method, although not a problem for experienced operators. The transceiver and front-end panels fit into the complete station layout as shown in Fig 9.

Counter Circuit

The counter is that section of the radio that is closest to a direct copy of an established circuit. With only slight changes, I combined circuits from several versions of Signal/One counters and from an independent supplier.[17] The design is straightforward; its circuit is shown in Fig 12. The chief advantage of this circuit is the absence of multiplexing, which can cause noise problems. The counter reads a PTO frequency directly, displaying only the kilohertz digits. Since the PTO range is 3.1 to 4.1 MHz (as explained in Part 1, p 20), the counter is configured to start at 090000_{10}. The resulting count, over 100-ms intervals, is 400000_{10} to 500000_{10}. Leading and trailing digits are not displayed. The PTO covers 1 MHz, with about 50 kHz over-range coverage at each end. The normally displayed readings are from 000.0 to 999.9. Beyond that, an overflow dot lights on the left of the display, warning the operator. With the band switch on 7 MHz, for example, a reading of 005.0 means 7005.0 kHz. However, a reading of .005.0 means 8005.0 kHz. Use caution!

The counter has three main sections. The portion from the crystal oscillator to the 7493 generates a 100-Hz clock, line K, with a period of 10 ms. The second section, consisting of the 7493 and associated TTL gates, divides the clock by 12; this yields a total counter period of 120 ms and

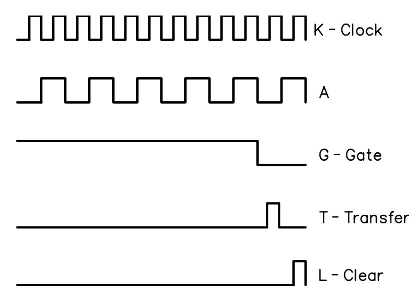


Fig 13—Timing chart for the frequency counter. Each horizontal division represents 5 ms. The vertical scales indicate the high or low states of each line. After the total span of 120 ms, the circuit is reset by the 7473, configured for a count of 12. The gate line, G, limits the actual counting interval to 100 ms; B and D are high when the 7473 reaches a binary count of 1010, or 10 counts of the 10 ms clock, K. Line A is used to divide the remaining 20 ms into two intervals, for T and L. The three-input gate for the transfer line, T, is wired so that T is high only during the 5 ms during which K is high, while G and A are low. The clear line, L, is obtained similarly.

Fig 12—Frequency counter schematic diagram (opposite). Commonly available TTL ICs are used. The simple gates are all included in one 7404 hex-inverter package and one 7410 triple three-input NAND gate package. Type 74143 is a combined BCD counter, storage latch, decoder and seven-segment output driver. It does not have a "start at nine" feature; this complicates the circuit by requiring several other ICs to drive the first digit.

D1-D4—LED panel read-out, seven-segment, common anode; HP type 7661, or 7660 for left-hand decimal point at the first digit for the over-range indicator. Panasonic types LN514OA and LN514GA (orange and green) are available at Digi-Key, #P327 and #P329 (also available in red and amber; see Note 22). The Panasonic LEDs are not available with left-hand decimal point. The over-range indicator may be placed to the right of the fourth digit.

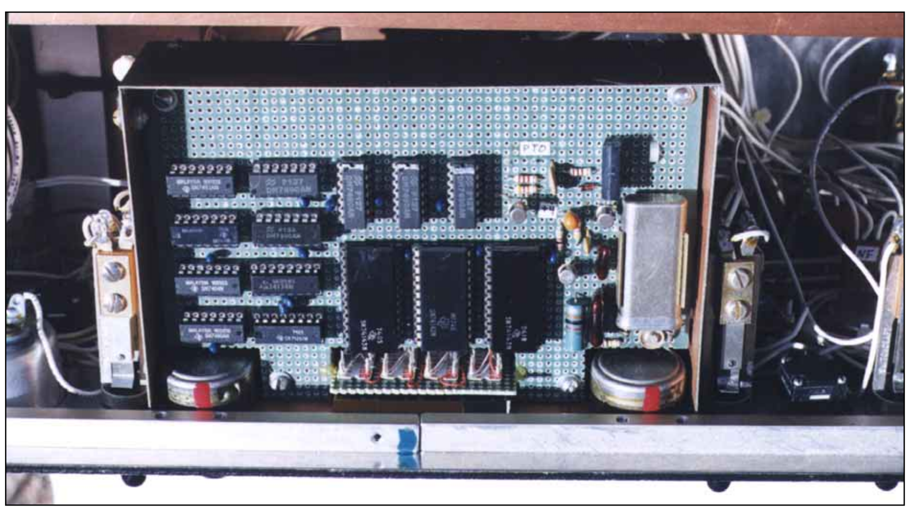


Fig 14—Top view of the frequency counter. The socketed DIP ICs are wire-wrapped; discrete components are hand wired point-to-point. The LED readout sub-board is also wire-wrapped and is cemented at right angles into the cutout on the main counter board.

produces the various control lines. The 7493 is a divide-by-16 binary counter; strapping the C and D output lines to the reset pins reduces it to a divide-by-12 counter. A count of 12 has binary output 1100_2; ie, C and D are high. The gate line, G, determines the total counting time of 100 ms. The remaining 20 ms of the total counter period are for the transfer (line T) and clear (line L) functions. A timing chart is shown in Fig 13.

The third portion of the counter, from the PTO input to the display digits, performs the actual counting. The more complicated circuit for the first digit implements the start-at-nine and over-range functions.

Powering the Counter

The LED drivers in the counter can become quite hot. The TTL ICs are specified for operation at 4.75 to 5.25 V. I found, however, that this counter would operate perfectly at voltages down to 3.5 V dc. An LM317 adjustable regulator, with input from the 7808 primary regulator, supplies the counter power. Tests were run over many months while work continued on other portions of the radio. A setting of 3.9 V has powered the counter adequately for 10 years, with greatly reduced IC temperatures and a much longer expected life. The LEDs also benefit from this technique, giving a softer, yet sharper display.

Counter Construction

The counter is seen in Fig 14. It is built on plain perfboard using wire-wrap construction. Four LED digits are used for the display: three in yellow for kilohertz and a green one for tenths of kilohertz. An escutcheon, seen in Fig 3 of Part 1, covers the ragged edges of the panel cutout; it was recycled from a CX7.

Power Supply

Design considerations and a general description of the power supply were given in Part 1, pp 23-24. The circuit uses standard IC regulators; the schematic need not be given here.[18] Four small transformers are used. A heat sink, spaced away from the rear panel, holds the four primary regulators: two 7818s, a 7918 and a 7808—all have TO-3 cases. Each of the four main boards employs TO-220 secondary regulators as needed: 7815, 7915 or 7805. One adjustable LM317 is used for the counter, as mentioned above. This double regulation avoids all transient problems. In addition, the IF board has a second 7815 regulator that powers only the AGC detector. This keeps AGC noise out of other circuits. In some radios, what sounds like an AGC click from inadequate AGC attack response is only noise conducted by a common power supply.

Cabinet Construction

Packaging for a full-featured radio can be a problem. The many circuits require considerable space. Modern mass-production methods are not available. A homebrew radio is more like a prototype; simple methods are required. Space must be provided for innumerable circuit modifications in the eternal quest for perfection.

The method used to build the main boards was described in Part 1 and is shown in the photos accompanying previous segments. The four main boards are secured inside the cabinet using pins. There are no removable screws or nuts to deal with. Each board has two small holes on the bottom wall and sits on pins fashioned from nylon screws attached to the cabinet bottom.

The IF board is pinned firmly in place by captive thumbscrews secured to the cabinet sides, which enter holes in the sidewalls of the board. This makes board removal very quick and easy. The plan was the same for the three other main boards. Because wire-harness overload developed during construction, however, the other boards had to be shortened; they have side pins only on the right. (I omit gruesome details, including the need to saw off the left end of the RF board after it was built and installed.) A removable top bracket with four nylon pins (not shown) secures the other boards using the IF board as an anchor. The four holes in the top walls may be seen in Fig 5 of Part 1.

Table 5 lists materials needed to assemble the cabinet. The LMB Omni Chassis series provides numerous prepunched holes for fastening and results in an excellently shielded enclosure. The cabinet's bottom plate is permanently attached—all work is done from the top of the radio. The cabinet itself has no front; this avoids the troublesome problem of drilling and mounting controls on a double wall. The front panel, for a standard 19-inch rack, has a plain, unpainted rear surface. This ensures good bonding and shielding. A ridge of solid square aluminum is permanently fastened to the rear of the panel, positioned to fit inside the cabinet. The bar is fastened to the panel with twelve permanent screws threaded into the bar from the front of the panel. These permanent front screws may be seen in Fig 3 of Part 1. Part of the bar may be seen here in Fig 14, and an outline of the bar is shown in Fig 15. The panel is released quickly and easily by removing hidden screws that thread into the bars from the sides, top and bottom of the cabinet. This method avoids the need to remove screws from the front of the panel and so prevents marring or scratching of the paint. All panel-mounted controls and assemblies swing down with the panel; there are no knobs to be removed or shafts to be disconnected. Similarly, the rear panel is easily lowered.

The best size for the main boards is 7.5×15×2 inches. Details were given in Part 1. The front panel is 8¾ inches high. The next standard size, 10½ inches, would allow space for another row of knobs along the top of the panel—and still more special features!

To avoid wire-harness overload, wiring between boards is best done with #24 stranded wire with thin (10-mil) insulation. Irradiated hook-up wire has insulation that will not melt from

Table 5

Material list for Cabinet Construction: The components used are intended for constructing a 4×17×17-inch chassis. Two sets of components are used, bolted together to form a cabinet 8×17×17-inches. LMB stock numbers are listed (see Note 23). The sides are 40-mil (1 mil = 0.001 inch) aluminum; the covers are 63-mil; the front panel is 125-mil. The cabinet is perfectly rigid when fully assembled. Because the bottom cover is permanently attached and all work is done from the top, the slight flexibility with the top cover off causes no trouble.

3	S417	Chassis sides (pair); 4×17 inches
2	C1717	Top and bottom covers; 17×17 inches
1	875	Front panel; 8¾×19 inches (specify black, texture finish)

the heat of a soldering iron. Alpha 7054 irradiated #24 hook-up wire is available as Mouser #602-7054-100-01.[19] Coaxial-cable interconnections were discussed in Part 2, p 5.

There are several advantages to building the front-end sections on separate panels. First, the main transceiver panel—which boasts an output power of 200 µW—is made to generate a minimal amount of heat. This reduces drift problems to a negligible level. The CX7 powdered-iron sliding PTO cores used in this radio have a poor temperature characteristic, making compensation over the entire 1-MHz range impossible.[20] Keeping the main transceiver cabinet cool circumvents this problem. Second, the cabinet needs no vent holes and is completely dustproof.

Front-Panel Layout

This radio was designed and built with the firm belief that it would never be published. The inside photos show neglect for appearance, putting a whole new dimension on the term "ugly construction." Eight years of modifications have not improved the appearance!

A front panel is different: The operator will confront the panel for years to come and that is all most visitors to the shack will see. Every builder wants the front panel of his project to look special. Arrangement of controls is a serious matter, especially for contest work or all-night DXing. The goal is to promote convenient and efficient operating and, at the same time, a pleasing appearance.

An old Mac 512 computer with the *MacDraw* program was used to lay out the panel, although any drawing program that produces full-size output could be used. Two panel versions, with and without center-lines, were kept on a floppy; this was easily managed by keeping the lines in an overlay file. A first-draft plain version was printed. Then all the knobs were set on top of the printout and the array studied for balance and general appearance. Changes were made, a new copy printed and the knobs again put in place. This process was repeated every evening for 30

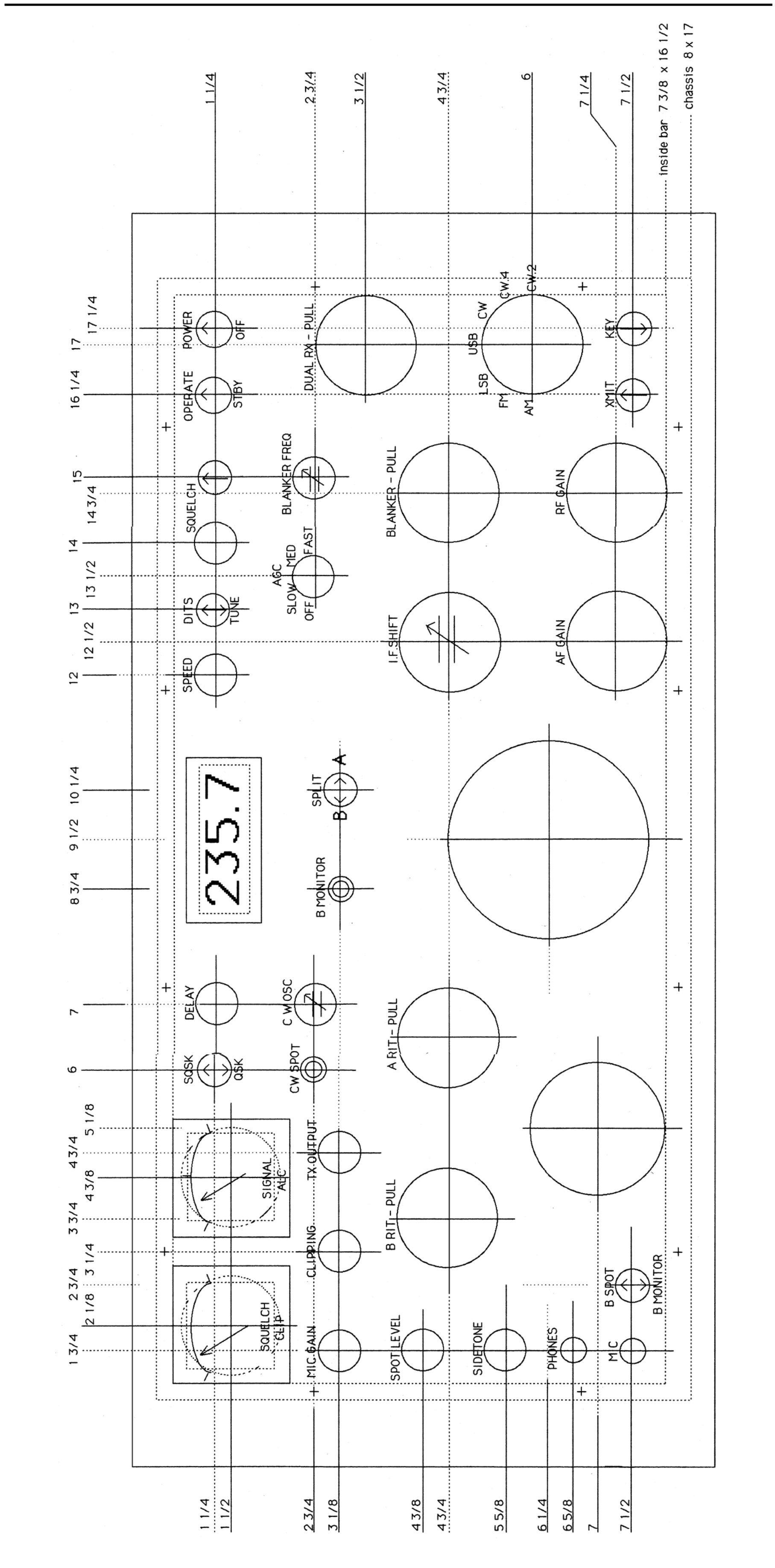


Fig 15—Original template for the front panel. Drawn on a Mac 512 computer, the template is placed over the aluminum panel and each hole is center-punched right through the paper. A few of the panel functions and labels have been changed since the panel was punched.

days. The final panel template is shown in Fig 15. A photo of the panel is shown in Fig 3 of Part 1. Fig 4 of Part 1 shows the control labels clearly, but it is not drawn to scale; the Mac computer printed the template exactly to scale. The printout was taped in place over an aluminum panel and each hole was center-punched through the paper. Many of the other panels in the shack have at least one hole drilled in the wrong place. This transceiver panel, drilled using the template, is the exception.

Performance

It would be well for any construction article to present performance measurements right at the start. With apologies, detailed measurements for this radio must be postponed to a subsequent article, in which complete test set-up details will also be included. Measurements with the available equipment indicate performance at or above the level of factory radios recently reviewed, and eight years of operating have confirmed the results. Nonetheless, more test equipment must be acquired and techniques must be refined before specific measurements can be submitted for scrutiny by *QEX* readers. Unforeseen circumstances have caused delays.

Some articles present proof of performance by listing a few DX stations that have been worked with the equipment. This may not be convincing. On the other hand, mere bench measurements are no substitute for actual operating under severe conditions. The discrepancy between lab techniques and actual ham-band conditions is one problem.

The 20-kHz-separation IMD test is a case in point. Try to find a clear 20-kHz segment on the 20-meter band during a DX contest or Sweepstakes! While published reports of factory radios often indicate IMD performance only for 20-kHz spacing, there are expanded reports available that clearly show a deterioration of performance as the spacing is decreased. Preliminary IMD tests indicate that the homebrew radio has the same excellent dynamic range at 3-kHz spacing as for 20-kHz. This is because the radio does not rely on a first-IF filter to shield the second mixer. See the discussions on “Tuning the First IF” and “Receiver Gain Distribution” in Part 1, pp 18-19. On-the-air tests can and must supplement lab measurements, with respect to actual band conditions, operating features and convenience.

Results

The following statements concern-ing on-the-air tests are not given in lieu of precise measurements to be presented later, but only in reply to some inquires.

During the eight years this homebrew radio has been in use, about 40 contest certificates have been received for events from 160 to 2 meters. Most of these are minor section awards; there is only one top-10 plaque on the wall.

However, a contest is a severe test for a radio not only for a big-gun station, but also for a little pistol.

DX work also subjects equipment to severe tests. With no Beverage antenna, the radio (with an amplifier) has worked 110 countries on top band (1.8 MHz). With only a single Yagi on the horizon for moon-rise/moon-set (and again an amplifier), seven countries have been worked on moonbounce (144 MHz). All the ham frequencies in between have also been used extensively with excellent results.

Summary

This article concludes a series that describes the 40-MHz main panel of the K5AM homebrew transceiver. Although the radio was completed eight years ago and has been in constant use since, modifications and improvements are still in progress. Any suggestions are most welcome! Subsequent articles will describe the three front-end panels for HF, 50 MHz and 144 MHz.

Notes

[1]M. Mandelkern, K5AM, “A High-Performance Homebrew Transceiver: Part 1,” *QEX*, Mar/Apr 1999, pp 16-24.

[2]M. Mandelkern, K5AM, “A High-Performance Homebrew Transceiver: Part 2,” *QEX*, Sept/Oct 1999, pp 3-8.

[3]M. Mandelkern, K5AM, “A High-Performance Homebrew Transceiver: Part 3,” *QEX*, Nov/Dec 1999, pp 41-51.

[4]M. Mandelkern, K5AM, “A High-Performance Homebrew Transceiver: Part 4,” *QEX*, Jan/Feb 2000; pp 47-56.

[5]Thanks to Bill Carver, W7AAZ, for help with this discussion of TTL and PIC methods.

[6]R. Dean Straw, N6BV, Ed., *The ARRL Handbook for Radio Amateurs* (Newington: ARRL, 1999, Order #1832). ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line at **http://www.arrl.org/catalog**.

[7]D. Lancaster, *TTL Cookbook* (Indianapolis: Howard W. Sams & Co, 1974).

[8]D. DeMaw, W1FB, *The ARRL Electronics Data Book* (Newington: ARRL, 1988, Order #2197).

[9]By some historical accident, this station uses negative external control lines for PTT and key lines. Former Signal/One owners will understand. The circuits shown in Figs 5 and 6 are modified and untested circuits that are suggested for use with the positive external lines more commonly used.

[10]M. Mandelkern, K5AM, “Design Notes for ‘A Luxury Linear’ Amplifier,” *QEX*, Nov 1996, pp 13-20.

[11]Thanks to Dan Hunt, K5WXN, for this suggestion.

[12]One of the most useful items of workbench test equipment is an old CX7, used as a test receiver or a test transmitter. With its bidirectional 40-MHz jack, it also serves as a front end for testing the basic transceiver or as an IF for testing a front-end panel. Added panel switches permit switching off the PA and the first LO as needed for various tests. All CX7 rear-panel connections are brought out to a test panel (on a shelf above the workbench) for quick connection to any device under test.

[13]The four switches disable the oscillators by applying –15 V to the lines µMO, µOO, µA, and µB.

[14]These radios can indeed be repaired and used. I have six in perfect working condition, including two converted to 50 MHz for mountaintop contesting. However, CX7s from early production runs lack some improvements in construction that are found in later runs, so when they are found at flea markets it is best to consider them as “parts radios” (ie, for salvaging valuable components). Anyone who wishes to restore a CX7—an enjoyable and instructive “boat-anchor” project—is better off starting with a later version. The best are those with Florida labels and serial numbers in the 800s and 900s.

[15]Thanks to Paul Kollar, W8CXS, for 25 years of advice on repairing CX7s and especially for detailed instructions for overhaul of the PTO mechanism.

[16]Coil adjusting has been tried in a different context and found feasible. In one of the CX7s converted to cover 50-51 MHz, shifting only two turns gained 150 kHz at the upper end, enough to cover the ZL segment near 51.110 MHz during solar cycle 22.

[17]Dick Cunningham, K0HHP (SK), produced LED counters to replace the Nixie counters in the CX7. This was not only a fine improvement that eliminated the flickering Nixies, but helped keep many CX7s on the air, since nearly all the original Nixie counter boards were defective.

[18]Each regulator requires input and output bypassing. Minimum requirements vary with the type of regulator and even with the manufacturer. To avoid consulting the data books every time a regulator is needed, I keep a stock of inexpensive surplus 2.2-µF tantalum electrolytics (a size sufficient for any type regulator) on hand. Two of these are installed directly at each regulator. Also, a 1-Ω, 1/4-W (or a higher wattage where needed) series resistor is installed at the input and output of each regulator. A DMM set to the millivolt range and connected across the shunt will read milliamperes directly. This helps to monitor current drain and locate defects. The resistors also serve as fuses in case of shorts.

This note affords a chance to insert a bit of voltage-regulator trivia. This concerns the LM317, used to power the counter in this radio. There are several misconceptions involved in use of the LM317 as commonly seen in published schematics. Most often one sees a 240-Ω resistor used to establish the minimum load current and the bias current for the control terminal. This value is seen in the data books and is apparently simply copied without further thought (see Note 24). However, there is nothing special about this particular value, which is not a common 10% value normally stocked by hams. It is never good to specify unusual components when common types will suffice. The data sheet specifies a minimum current for proper operation; with the 1.25-V reference voltage maintained between the output and control terminals and a specified 5-mA minimum load current, this works out to 250 Ω. The next lower 5% value of 240 Ω (available for factory design) is shown in the data books. In a ham workshop, the next lower 10% value of 220 Ω may be used. Of course, the required resistance at the control terminal must be calculated using the bias current obtained and chosen to provide the desired output voltage. Nonetheless, 220 Ω would still be wrong! The second common misinterpretation of the data sheets stems from the fact that the application diagrams are shown using the military or industrial versions of the regulator, the LM117 or LM217. The LM117/LM217 does have a 5-mA minimum load-current specification. This version is seldom seen at ham suppliers, and it would cost too much even if located. A careful reading of the specification sheet shows that the commercial/consumer version, the LM317, has a minimum load current of 10 mA. Thus the resistor should be a maximum of 125 Ω; either 120 Ω or 100 Ω could be used. All this is really trivia, of course; the circuit usually draws more than the required minimum. But an LM317 with a 240-Ω resistor in a schematic for a weekend ham project needn't send readers rushing out to look for a precise component.

[19] Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76063; tel 800-346-6873, fax 817-483-0931; **sales@mouser.com**; **www.mouser.com**.

[20] Thanks to Harold Johnson, W4ZCB, for information on the construction of the Signal/One PTOs.

[21] Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952; tel 800-524-6464, fax 800-524-5414; **hosfelt@clover.net**; **http://www.hosfelt.com/**

[22] Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677; tel 800-344-4539 (800-DIGI-KEY), fax 218-681-3380; **http://www.digikey.com/**

[23] LMB Heeger, Inc, 6400 Fleet St, Commerce, CA 90040; tel 213-728-5108, fax 213-728-4740; **www.lmbheeger.com**. Ask for a catalog and information on direct ordering with free shipping.

[24] For example, pp 10-10 to 10-17 in *1982 Voltage Regulator Handbook*, 1981, National Semiconductor Corp, 2900 Semiconductor Dr, Santa Clara, PO Box 58090, CA 95051; tel 408-721-5000; **www.national.com**.

High-Quality SSB Audio

I know you can hear me, but can you understand me? Come read the case for and the methods to achieve full-range SSB audio.

By Bob Heil, K9EID

The audio used to modulate Amateur Radio carriers has traditionally been a constant roller-coaster ride and an ongoing subject for a majority of QSOs since the invention of the plate-modulation transformer. Some Amateur Radio stations produced audio equaling many small-market broadcast stations in their day.

During the late 1950s, SSB moved onto the scene and most of that great audio quality was traded for a much narrower signal that gave the Amateur Radio fraternity much less noise and interference. It just became accepted practice to roll off low-frequency response and concentrate on articulate midrange audio that cuts through pile-ups and provides clean, clear speech signals.

DX and contest stations experimented with anything that would increase their country counts or contest points, only to finally discover that the most important single item is the *frequency response* of their transmitted audio. The goal is to match it with the Fletcher-Munson response chart for the human ear. (See the sidebar, "Decibels, Hertz, Dr Fletcher and Dr Munson.") This, of course, placed high-fidelity transmit audio on hold as the results of high-powered midrange equalization of specialized microphone elements continued to raise those country counts and contest points.

Amateur Radio operators have always been experimenters. For years, they built most of their transmitters, antenna tuners and antenna systems. An interesting observation is that the audio quality has been degrading for years. It became a rather neglected element in the chain of equipment necessary to communicate. If the microphone plug fit and the meter moved, that was sufficient.

During the early 1980s, an interesting movement began with the introduction of a simple, two-band microphone equalizer. The August 1982 *QST* cover award was given for the article "Equalizing Your Microphone." This article enlightened Amateur Radio operators to the

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importance of microphone equalization and matching their transmitted audio response to that all-important Fletcher-Munson curve. A whole generation of radio amateurs began paying attention to the quality of their transmitted audio. Again, the microphone had been rediscovered.

Keep in mind that a single-sideband signal has traditionally been a fairly narrow-bandwidth mode. The low-frequency roll-off is set at 300 Hz, while the high end is usually switchable in the transceiver at 1800, 2100 or 2400 Hz. It's important to remember that these filters *roll off* the high and low ends of the response curves. The filters do not simply shut the door on any frequencies outside of their selected bandwidths, but reduce the audio at a rate of about 18 dB/octave. By cascading filters or using higher-technology, 24 dB/octave and greater roll-off rates are possible. For analog filters, anything above 18 dB presents some ringing that is very annoying when listening to speech.

Frequency Sensitivity of the Ear

Thousands of hours of on-the-air tests have proven that there are two important kinds of audio for SSB transmitters. When signals are noisy because of either storm-related noise or heavy interference from pileups, full-range "bass" signals are not easily copied. Many Bell Labs studies have shown that most of the important speech sounds fall in the 600-2200-Hz region with a *very* important range centered around 1800 to 2000 Hz. These all-important frequencies make up the puffy, sibilant parts of speech the mind *must* hear to tell a "d" from a "b," an "f" or an "s." Along with that, the correct balance of highs and lows creates a voice characteristic. Many tests proved that the intelligibility of an SSB signal totally depends on the frequency content of speech. With the proper microphone and equalization peaking the frequency response at 1800 Hz, articulation is perceived much better than when the audio peak was centered around the 600 to 800 Hz region. This latter situation *harms* the intelligibility of the signal. Of course, we are talking about punching through a noisy, crowded band.

When things are quiet and you are listening to 30-over-S9 signals, the transmitted frequency range can be greatly increased and, in most cases, this is quite desirable. It becomes very unpleasant to listen to bright, ear-piercing audio (the kind the contest and DX crowd love!) from big, strong signals.

TX Equalization

Thus, a SSB station really needs two different microphone elements or a system capable of achieving two different patterns of equalization. It becomes extremely important that the full-range audio not just become booming bass or "boxy" sounding. Since we are reproducing audio frequencies outside of the nominal filter bandwidth, one must listen very carefully in another receiver to hear whether this boxy sound is created; each voice is different and effects the equalization so much differently. We are not just looking to greatly increase low-end response. We are looking to create a very nice, warm *balance* of frequencies, and many different factors help to make the perfect signal.

Only by monitoring yourself on a second receiver can you absolutely know what you sound like to the outside world. Disconnect the receiving antenna and plug in a pair of high-quality headphones. This will ensure that you won't be causing any feedback from the speaker to the microphone. Listen—mentally dissect what you hear. Changes in the equalization, com-pression, noise gate, etc will become readily apparent. Recordings and reports from your friends cannot tell you how you sound. You *must* listen to a receiver in real time. The "monitor" feature of many transceivers *never* gives the exact picture. It is simply a power-amplifying system from the microphone to the speaker. You must have the filters in a receiver involved to know exactly how others hear you.

So many signals on the air today have that booming, boxy sound. Just because the equipment is producing a lot of bass response does not make for a well-balanced signal. The problem lies mainly in the microphone. Most microphones use elements intended for public-address or stage work. Not one single microphone shipped with equipment today has been carefully developed for SSB communications. When dealing with those terrific SSB-filter roll-off rates, we discover that the 100 to 200 Hz region response so prevalent in the "matching" microphones shipped with transceivers strengthens the low end at exactly the *wrong* spot. Repeated tests have proven that we need at least 8 to 10 dB of boost *below* 100 Hz, then *reduced* gain from 150 to 350 Hz, as we pass the 3-dB point of the SSB filters. Increases of these frequencies cause "boominess."

One must constantly remember that the focus of any Amateur Radio station is communication: exchanging information. For many decades, that was the sole point. In recent years, however, amateurs have finally come to realize that most transceivers sound about the same—given like-bandwidth filters. The big difference in signal *quality* begins at the microphone. Through listening, we can determine that one mike may yield clean, clear articulate audio while another may sound like it's wrapped in three layers of cotton—muffled and difficult to understand. Behind this subjectivity however, there resides a high degree of objective quantification. Specification sheets, response curves and other measurements sometimes lead only to very interesting arguments. The response of a particular microphone element is evident only when it's connected to a particular transmitter. That is the *real* answer.

Seldom do response curves, specification sheets and such give information about IMD, transient attack time, proximity effect, sensitivity to breath blasts and so forth. The truth in all this emerges from very precise listening: mental dissection of the output to determine good, better, best performance for a given situation, transmitter, voice and—last but not least—desired sound.

The Hi-Fi Movement

Audio quality was at its peak during the AM heyday of Amateur Radio. With the introduction of SSB and eventually an almost complete move away from AM in the 1960s, operators had to sacrifice "broadcast-quality" AM audio for a narrower bandwidth to help accommodate increased activity. Audio quality has been deteriorating over the past 20 years as many operators have paid more attention to output power, antenna systems and the bells and whistles of their new transceivers. Their circle of friends never told them they sounded like their matching microphone was still in its box. DX and contest operators focus their audio on intelligibility. Plain and simple: Get the message across. Waste no power on transmitted information that does not contain vital characteristics of communications.

When your signal is 30 over S9, rag chewing with your friends is not comfortable with ear-piercing contest audio. To answer this, several transceiver manufacturers and at least one microphone manufacturer have acknowledged a movement toward what could be called high-fidelity SSB: single-sideband audio that rivals the quality of an AM transmitter.

One of the leading factors in achieving this higher-quality SSB audio is the inclusion of DSP in just about every transmitter. Even tiny mobile rigs have powerful DSP units on board, giving operators the ability to adjust many parameters heretofore simply unavailable. Changing the carrier level, filter bandwidths, equalizing the microphone-audio preamplifier and a host of useful items that—when coupled with a high-quality studio microphone—can help achieve very high-quality SSB audio. This audio can be further enhanced by the addition of outboard audio equipment used primarily in broadcast and recording studios.

No matter which filter the manufacturer has chosen, the goal of a SSB voice signal is to communicate—to transmit and receive articulate speech. Following frequency parameters set by the telephone industry, the SSB AF spectrum has traditionally been 300 to 2400 Hz.

Yes, the movement from AM to SSB meant that full-range audio quality was given up, but the trade-off was much narrower signals that allow two or three stations to occupy the same bandwidth as one AM counterpart.

As we have previously discussed, Amateur Radio operators are the ultimate experimenters. They constantly push the envelope, never needing to say "No." Recently, there has been a movement by some of these experimenters to push the audio envelope, to see just how much audio they can produce through those narrow-bandwidth filters. Several developments in the Amateur Radio marketplace make it very easy to achieve beautiful, full-range, broadcast-quality SSB audio.

A definite chain of equipment is needed to achieve this full-range SSB audio. Of course, it all starts at the microphone—and we are not talking about a little transceiver-matching desk-stand mike that was packed with the transceiver. Just as in most broadcast stations, a studio microphone's audio must be properly processed. A processor in the broadcast world does much more than the processors built into Amateur Radio gear. Mostly, the latter is an amplitude-leveling device: It brings low-level audio up to a preset level and reduces peaks to that same level. In many cases, transceiver processors do more harm than good.

Broadcast processors have very high-gain microphone preamplifiers, multiband equalizers, split-band processors, noise gates and line drivers—all in one single-space rack chassis (see Fig 1, Table 1). They are indeed self-contained voice processors that—when properly adjusted—enhance intelligibility, reduce off-mike background noise and increase perceived presence, all with extremely low distortion.

On some Saturday evening or Sunday afternoon, tune in your local National Public Radio station and listen to Garrison Keillor. Pay attention to the segment when he presents his "News from Lake Wobegon." It's a perfect example of a broadcast voice processor at work. The engineers have always done a terrific job adjusting the voice processing equipment on that show. Every syllable is extremely articulate, presenting the illusion that you could just reach into the speaker and touch Garrison. *That's* the effect of properly designed and adjusted voice processors.

No, we are never going to get a ham SSB transmitter to sound like your NPR affiliate. After all, we only have about 2 kHz of useful bandwidth. Nonetheless, we *can* present that same illusion of closeness, voice presence and articulation. The frequencies from 500 to 1200 Hz need a 4- to 5-dB boost. These frequencies are where we find many of the puffy sounds. This is where careful listening to the individual voice is so important. These frequencies support just about every phrase or syllable in the English language, but too much gain in this range creates a "honky" midrange thrust to each word. With careful tuning of the equalization while listening to a second receiver, you can build a great foundation to add the upper "mids" and sibilant highs.

As we have discussed earlier, the 1800- to 2200-Hz spectrum is very important. It is where the defining parts of sibilance lie. Without the correct emphasis on these frequencies, the receiving station has trouble distinguishing an "f" from an "s" and a "b" from a "d." It is *extremely* important. The human ear is very sensitive in this region but must have the correct balance centered near these frequencies.

The extreme top end of our audio spectrum is around 3000 to 3500 Hz. Since it lies outside of the SSB filter

Fig 1—The panel of a Rane VP-12 voice processor.

Table 1—Broadcast Voice Processors

Mfgr	*Model*	*Description*	*Price*
Rane	VP-12	Mike processor, two-band Parametric EQ	$599
Rane	DC-24	Mike processor, two-band Parametric EQ, Limiter, Compressor, Gate	$699
Symetrix	528E	Mike preamp, three-band Parametric EQ, Limiter, Compressor, Expander	$549
DBX	286E	Mike preamp, EQ, Compressor, Expander, Gate	$349

bandwidths, the equalization will again be at least 6 to 8 dB of boost to properly balance the audio signal across the range needed for a full-range SSB signal.

Audio Compression

Once the equalization is adjusted, we need compression. The nature of an audio-compression circuit is to reduce the higher peaks and increase the lower levels to a common level that will drive the radio's microphone input properly. Now we have a *huge* problem. Having just spent considerable money and time to equalize each important frequency band for the ultimate signal, directing that complex audio signal through a single compressor will simply destroy the results of all of our prior effort. This audio signal has many peaks and valleys. We want it to stay that way.

Therefore, only a split-band compressor will do. It has two or three *separate* compressors that can be assigned to different frequency points, hence compressing the low end harder than the midrange and differently than the all-important top end. With the split-band unit, we can successfully compress audio without all of the distortion and "pumping" so prevalent with the single-band compressors incorporated in most transceivers.

Noise Gates

Because the microphones are very sensitive and the entire audio front end is designed for use in a recording studio, the output of a broadcast voice processor usually sounds like you are seated behind the engines of the space shuttle. There will be tremendous background noise. Blower noise from any high-powered amplifier will be quite evident. The solution is to patch all of your high-gain microphone preamplifiers, equalizer, split-band processors and line drivers through a very active noise gate. This will eliminate sounds around you, except for the speech audio of the microphone.

Inevitable RFI

Finally, thousands of dollars and many months of installation and adjustment can bring all of this wonderful, full-range audio to your SSB transmitter—maybe. Patching all of these pieces of marvelous broadcast equipment together seems fairly straightforward; however, one item will probably raise its ugly head to cause sleepless nights and head-banging: RFI.

You have placed all of this extremely high-gain audio circuitry in front of a high-power Amateur Radio transmitter. We have all heard for years to ground *everything everywhere*—long copper ground rods, electrical entrance panels—anything you can find that hasn't been grounded, *ground it!* Now, with all of this new audio gear on board, we just might be back at square one. High-gain audio circuits must see one—and only one—ground. Your transmitter likely does not ground the microphone shield. This may sound crazy, but most manufacturers do not ground the shield of the microphone cable! Now you are pumping all of your power through an unshielded microphone line that is driving the front end of a 1500-W, 20-meter transmitter—oh yes, there will be RFI!

There are many horror stories out there about station operators who have spent six months to work out the many ground loops, empty grounds, dual ground paths and what have you. Have patience: Just when you want to shovel all this high-dollar audio voice-processing equipment out the back door, things may begin to fall in place and your signal will become very well received. (Figs 2 and 3 show processing systems at my station.)

It is important to realize that no matter how much you invest in your audio chain and RF deck, your station will only sound as good as it its weakest link; it all starts at the microphone.

Microphones

A microphone—simple as it may seem—is a magnificent device. It is

Fig 2—An audio rack at my station. Equipment is as follows: (top unit, including row of slider pots) INTER M Model 1242 12-channel mixing console; (second row, three units) SLM Model SM4CL compressor/limiter, SM3PE three-band PARAMETRIC EQ and CBL cable doctor patch bay/ cable tester; (third row) SYMETRIX Model 528 microphone preamp/ processor; (fourth row) Heil Sound Model 435 four-band parametric equalizer. Also in photo: Heil Sound Model GM-5 GOLDLINE microphone; Luxo Model LX-1 microphone boom; AT-1 Model SM-1 microphone shock mount.

Fig 3—Equipment in my shack: (top row, three units) SM3PE three-band PARAMETRIC EQ, SLM Model SM4CL compressor/ limiter and a homebrew patch panel; (second row) SYMETRIX Model 528 microphone preamp/processor; (third row) Collins KWM-2A/312 B-5 (on desk) SPIRIT four-channel mixing console.

one of the few items in an audio chain that actually changes one form of energy into another. Specifically, it changes acoustic waves into electrical signals. Microphones can be divided into three groups, depending on the specific process for converting sound into electricity: dynamic (or moving coil), ribbon and condenser.

Back in the 1930s, crystal mike elements were popular, but they are fragile and very sensitive to heat and humidity. They have a much wider response than early dynamic mikes, but over the years, dynamic microphone technology has been able to produce frequency ranges equal to condenser elements. Very few crystal elements are in present-day use.

Unlike most FET/condenser and ribbon microphones, dynamic mikes are very rugged and reliable. They tolerate heat, cold and high humidity well and can handle high input levels without distortion. In a dynamic mike, a coil of wire is attached to a diaphragm suspended in a magnetic field. As a sound wave vibrates the diaphragm, the tiny hair-like wire—or voice coil—passes in and out of the magnetic field and generates an electric signal similar to the incoming acoustic wave.

In a ribbon microphone, a thin metal foil or ribbon is suspended in a magnetic field. Acoustic waves vibrate the ribbon and generate an electrical signal. Ribbon microphones are generally prized for their warm, smooth tone quality but they are extremely delicate as the thin metal ribbon can be easily damaged. They also have very low output levels, which necessitate a very high-gain, low-distortion mike preamplifier. RCA was the leader in ribbon-microphone technology. Their famous Model 77s can be seen each night in front of David Letterman and Larry King. It is interesting that these wonderful old icons are strictly props: The audio you hear comes from wireless Lavalier or overhead-boom microphones.

A FET/condenser microphone has a conductive diaphragm and metal back plate in very close proximity to form a capacitor. The capacitor is charged from an external voltage source. When acoustic sound waves strike this diaphragm, it vibrates, varying the spacing between the plates. This varies the voltage across the capacitor and generates a signal similar to the incoming sound wave. Condenser mikes include the true condenser and the newer-style FET/electret.

The true condenser mike is externally biased. In the later-style FET/electret, the diaphragm and the back plate are charged by an electret material on the diaphragm and the back plate. Both electret and true condenser mikes can sound equally good, although some engineers prefer the more-costly true condensers. They usually feature a wide, smooth frequency response with detailed sound and extended highs as well as excellent low-fre-

Decibels, Hertz, Dr Fletcher and Dr Munson

Probably the most used (and misused) term in today's electronic society is the decibel (dB). The term represents one tenth of a bel. The bel, named for Alexander Graham Bell, is a logarithmic measurement of power ratios. The decibel is *never* a unit of quantity like a gallon, watt and so on.

Every time power is increased by 3 dB, that power is doubled. An increase of 10 dB represents a power ratio of 10. The human ear can barely detect a 3-dB increase, so power-level increases may be undetectable by ear. They are sometimes only numbers to support the ego! It takes some incredible increase of power to double or triple perceived signal strength.

Another important term is "cycles per second," changed several decades ago to hertz. This term is easily understood, as it is simply the number of times per second something vibrates or oscillates. The human ear can usually begin hearing an audio signal around 20 Hz; its range extends to around 15,000 Hz.

Frequency Response of Human Hearing

The human ear has one great problem: None of us hears all frequencies from 20 to 15,000 Hz at the same loudness. Dr Fletcher and Dr Munson were two audio pioneers working for Bell Laboratories. They made a detailed study of this effect after early telephone engineers couldn't understand why some inventions did not work as planned. Back in the early 1930s, Dr Fletcher and Dr Munson made the startling discovery that no human hears frequencies less than 100 Hz very well. Oh, we hear them, but we hear them at least 3 to 6 dB weaker than we do other components in the spectrum. Human hearing response also starts rolling off about 14,000 Hz. The frequency response of the human ear is much flatter at louder volume levels. At low levels, a response curve for the human ear looks like a ride at Disneyland! (See Fig A.) This becomes extremely important when balancing or equalizing the tonal response of an audio system.

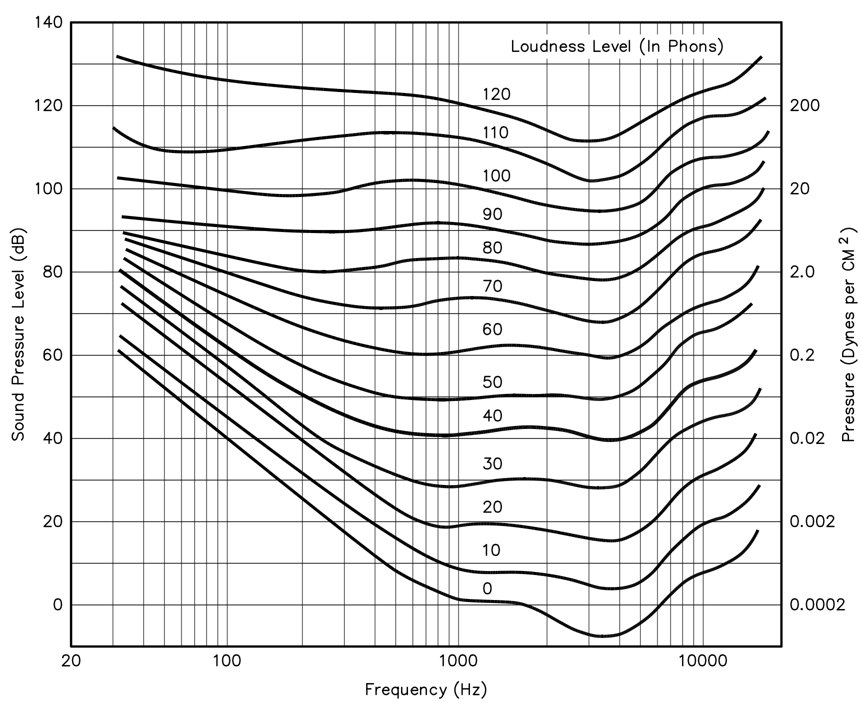

Fig A—The typical sound response of a human ear.

quency response. Because of its lower diaphragm mass and higher damping, the condenser responds much faster than all other mikes to transient sounds, such as those coming from percussion instruments, including pianos. Of course, this response is not important for speech.

These microphones can be powered by an external battery pack or a "phantom" power supply located in a mixing console or at the microphone preamplifier input. Phantom power places +12 to 48 V dc on the mike connector through a resistor. The microphone receives dc power and sends its ac signal down the same wire. When using phantom power with a microphone feeding an RF transmitter, take care to suppress RFI through extensive bypassing and decoupling.

Many transmitters feeding dc power through the microphone audio line are susceptible to a small amount of distortion. It usually is 40 dB below the audio level, but careful listening or viewing of *audio* signals on a spectrum analyzer will show this phenomenon.

Dynamic microphones, however, need no power supply. You can connect these microphones to phantom-supplied inputs, but the dynamic element must be decoupled with an in-line 1 µF, non-polarized capacitor. Dynamic microphones vary greatly in their general responses. Some have smooth, wide frequency ranges. Some have very narrow, "bassy" responses with little voice articulation. Some dynamics do not reproduce any frequencies below 300 to 400 Hz. In the past few years, great strides in dynamic-microphone technology have given us mikes that rival the responses and sounds of some studio condenser units.

Directional Patterns

Microphones differ greatly in the way they respond to sounds coming from different directions. *Omnidirectional* mikes "hear" equally well from all directions. *Unidirectional* microphones are most sensitive to sound waves arriving from the front of the microphone, least sensitive to those from the sides or rear. *Bidirectional* mikes are most sensitive to sounds entering from two directions: the front and back. They reject sounds entering from the sides.

When using dynamic microphones for speech communications, the unidirectional pattern is preferred over the omnidirectional pattern, as the later does not offer much suppression of background noise. When using a unidirectional microphone, its main lobe can be directed to the person speaking, thereby reducing unwanted background sounds.

Proximity Effect in Dynamic Microphones

One of the most damaging effects in a high-quality dynamic microphone is a phenomenon called "proximity effect." This effect is very desirable to a rock singer, but very degrading to high-quality communications. Proximity effect increases the bass response as the microphone is "swallowed" (brought close to the mouth). As the microphone is placed farther away from the mouth, the response is equalized.

Observe a well-trained vocalist and you can hear the response change as he or she "eats" the microphone. Understanding how to use a dynamic microphone's proximity effect can be a very worthwhile part of getting the most out of a microphone's response and dynamic range.

For speech communications, it is *vital* to speak very close to the microphone. Distances greater than four or five inches will create several problems. Many operators set a microphone on a desk stand 12 to 15 inches from their mouth. Room echo, sounds from adjacent rooms, blowers and other ambient noise become louder than the voice. To properly use a dynamic microphone for speech communications, place it on a professional studio boom and no more than four or five inches from the mouth—*very* important. This much improves the dynamic range, transient and high-frequency responses.

Frequency Response

When selecting a microphone, think about exactly what the microphone will reproduce. In a recording studio used for instruments such as violins, harps and pianos, for example, one must use a microphone with a flat (equilevel) response. Usually, ±3 dB is adequate from 60 Hz to 16 kHz—20 to 9000 Hz for bass instruments, 80 Hz to 12 kHz for brass instruments and voice. For speech communication such as Amateur Radio, a high-quality unidirectional microphone with a frequency response of 80 Hz to 8 kHz is needed; however, selection doesn't stop there. It is vital that the microphone have a 6 to 8 dB rise centered at 1800 to 2000 Hz. This allows the microphone to reproduce those important puffy sounds. Learning more about the Fletcher-Munson curve (in the "Decibels, Hertz, Dr Fletcher and Dr Munson" sidebar) will substantiate this and help one to listen carefully for these important frequencies.

A rising high-end or presence peak around 5 to 10 kHz gives the speech pattern the necessary balance of fricative sounds—the "s" and "z" sounds. The warmth added by proximity effect, the dynamic range of a close-talked voice, the midrange rise for the articulate "p" and "f" sounds all balance with the sibilance of the "s" sounds and together create a very special audio signal.

All the above-described microphones have definite individual sounds, and the audio engineer may chose the correct one for any given occasion: condensers for violins and choirs; dynamics for solo vocalists and so on. When choosing a correct microphone for Amateur Radio, things get a bit tougher. One would think that since we only have to fill a 300 to 2400 Hz bandwidth, why worry? With such a small bandwidth, what you put into it and at what frequencies become extremely important.

Many tests have proven that in weak-signal conditions—especially VHF SSB—excessive low-end response hampers the ability of the receiving station to copy. In a recent test of 50-MHz aurora signals, one of the HC4 midrange Heil elements produced amazing speech quality through the aurora when nothing else was intelligible.

When a signal is 40 over S9, however, we have an entirely different situation. You would never want to have that piercing, high-pitched audio for your rag chew friends. For strong signals, full-range sound is preferred. Nonetheless, remember that the microphone must be equalized with a rise in the midrange to give that articulate speech audio and balance the extreme highs and lows in reference to those all-important middle frequencies.

There is a true art to adjusting equalization; to complicate matters, we only can work across about *four* octaves—yes, four. I know, you want a 31-band hot-dog-brand, high-dollar equalizer to impress your radio friends, but you can only work from about 100 to 4000 Hz. All of those extra controls are for naught. They're great for live sound or in a recording studio, but wide frequency range is not needed for Amateur Radio.

Reality Check

If you want anything beyond off-the-shelf ham audio, you need to thoroughly evaluate the performance you desire and complete a system—from microphone to antenna—that achieves that goal. If you want a wide-band hi-fi SSB signal, you may invest tens of thousands of dollars in quality audio gear only to have the weakest link in your chain of equipment determine the overall quality.

All of this is science. The results are determined not by one piece of better gear or by my paying more than you, only by the science of audio. Here's a thought to ponder: Most of the popular transceivers tested here in my lab—costing upwards of $4000—do *not* pass frequencies under 250 Hz and barely pass 3000 Hz on the top end. This is a standard situation with present-day transceivers. The SSB audio spectrum is 300-2700 Hz. Current designs use input and output coupling capacitors suited to achieve that; this is the case to match the filter networks and DSPs installed. Why pay thousands of dollars for a studio microphone, a broadcast equalizer and a complete audio chain for a digital synthesizer, when the performance of those high-dollar components will be limited by the simple audio restrictions designed and built into your $4000 transceiver?

As Arsenio Hall used to say: "Things to make you go hmmmm...."

Licensed in 1956, Bob Heil's FCC license became his education: experimenting, building and learning basic electronics. It was his vast knowledge learned from huge VHF moonbounce arrays built in 1959 that led him to pioneer the rock and roll sound business. Heil was the sound engineer for such touring groups as the WHO, the Grateful Dead, Peter Frampton and his Amateur Radio friend and client, WB6ACU, Joe Walsh, for whom Heil invented the "Talk Box." Since 1977, Heil has presented his "High Tech Heil" consumer electronic shows on KMOX (CBS radio) and KSDK (NBC TV) in St Louis.

An author of five text books, teacher, lecturer and consultant, Heil is recognized as the "guru" of the home theater movement, designing over 1000 custom AV systems for homes as well as church and school sound systems. To the Amateur Radio world, Bob is best known as the audio genius behind his famous "DX Dream Machine" microphone, headset and hardware. Since 1982, Heil Sound has provided the leading Amateur Radio operators with highly articulate audio for their transmitters.

Heil, an accomplished theater pipe organist, plays the four-manual Wurlitzer at the Fox theater in St Louis, and Lincoln theater in Belleville, Illinois. He just released a new CD: Bob Heil—Live! *His years as a child learning to build and tune theater pipe organs taught him how to listen, and his interest in Amateur Radio was the path that led Heil to his creative and colorful life.*

A Class-B Audio Amplifier

Do you need a low-power, high fidelity amplifier? Would you like to wet your feet in a small amplifier construction project? Here's a simple amplifier suitable as a mike preamp or speaker driver.

By Parker R. Cope, W2GOM/7

The class-B amplifier is the workhorse of high-power audio and linear RF amplification because of its high efficiency and low tube or transistor dissipation. The theoretical maximum possible efficiency is 78.5% compared with 25% for a class-A amplifier [untuned—*Ed.*] and about 80-85% for a class-C amplifier. Class-A and -B amplifiers are linear and can serve in audio applications. Class-C amplifiers are non-linear and must have a tuned circuit for a load.

While the theoretical maximum efficiency of a class-B stage is 78.5%, the realizable efficiency is a function of drive. If the efficiency with maximum-permissible output signal is 60%, then the efficiency with half this applied signal will be 30%. Only switching-class amplifiers offer greater efficiency, but that's a different story.

The class-B amplifier is typically characterized by two devices in push-pull with each device biased to operate for 180° or more of the input cycle. One device operates during the positive half-cycle of the input, while the other operates during the negative half-cycle. Class-B amplifiers are usually biased to conduct over a little more than 180° in an attempt to improve linearity, but efficiency is sacrificed and device dissipation increases.

Complementary emitter followers make a simple class-B amplifier; biasing is simple and there is no need for push-pull transformers. When properly biased, the distortion of complementary emitter followers is certainly low enough for use in communications applications, but probably too high to be classified as high fidelity. Nevertheless, not all is lost. An inexpensive op amp like the LM741 can reduce distortion to levels that will satisfy discriminating listeners.

The major distortion in complementary emitter followers arises from errors in setting the operating point at the threshold of conduction and from the drivers' output resistance. Driving the emitter follower with high impedance introduces distortion because a transistor's h_{fe} changes with collector current. Consequently, the output

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current does not follow the input current. When driven from a voltage source (low impedance), changes in h_{fe} merely cause the input resistance of the emitter follower to change, not the output voltage.

Therefore, a low-distortion complementary emitter follower requires low-impedance drive and biasing at the threshold of conduction. Such biasing means that current starts to flow as soon as a change in the base signal places one of the transistors in a forward-biased condition. Proper bias can be provided by silicon diodes, as shown in Fig 1. The base of the 2N3904 is one diode-voltage drop above the signal voltage and the base of the 2N3906 is one drop below. The 2N3904 conducts as soon as the signal becomes positive, while the 2N3906 conducts as soon as the signal becomes negative. The static current in the diodes provided by R1 and R2 generates a diode voltage V_f that is approximately equal to the V_{BE} of the transistor. With V_f equal to V_{BE}, the bases are at the threshold of conduction under no-signal conditions.

Insuring proper base voltage even when the diode voltage doesn't exactly match the transistors' V_{BE} is neatly solved by including the emitter followers within a negative-feedback loop of an op amp, as shown in Fig 2. An op amp such as the LM741—when connected as a unity-gain buffer—has an output equal to the input. When the emitter followers are within the feedback loop, the output of the emitter followers is forced to be equal to the input of the op amp, thus no distortion is introduced by the emitter followers.

The distortion introduced by an amplifier with feedback can be expressed as:

$$\frac{D_f}{D} = \frac{A_f}{A} \qquad \text{(Eq 1)}$$

where D is the distortion without feedback, D_f is the distortion with feedback, A_f is the gain with feedback, and A is the gain without feedback. The gain, A, is the open-loop gain at the frequency of interest. For voice communications, the frequency of interest is about 2 or 3 kHz; for high-fidelity applications, it would be much higher, perhaps 20 kHz. The gain of an LM741 at 2 kHz is about 54 dB, and the gain at 20 kHz is 34 dB. Including the complementary emitter follower within the feedback loop of an LM741 reduces the uncompensated distortion of a 20-kHz signal by 34 dB.

Including the emitter follower within the feedback loop of an internally compensated amplifier is the ultimate in simplicity. The emitter followers' outputs are connected to the inverting input and the signal is applied to the non-inverting input. Since the emitter followers have 0° of phase shift to several megahertz, they can be included as part of the op-amp circuit without further consideration of their phase shift/frequency response.

The peak current of the 2N3904/6 is about 100 mA. That's only 40 mW into an 8-Ω load. 40 mW is sufficient for headphones, but a bit light for high-fidelity speakers. More output can be obtained with complementary Darlingtons like the TIP120 and TIP125. Of course, another choice is to use Darlington-connected 2N3904/TIP120s and 2N3906/TIP125s, as shown in Fig 3.

With either Darlington implementation, bias is provided by two silicon diodes instead of one. If the drive requirements for the Darlingtons is more than can be provided by a single LM741, a class-A emitter follower can be used following the op amp. A class-A emitter follower has a broadband

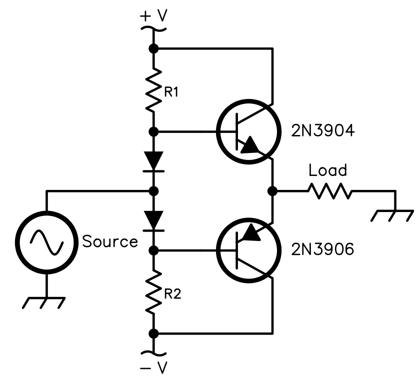

Fig 1—Complementary emitter followers can be biased with silicon diodes.

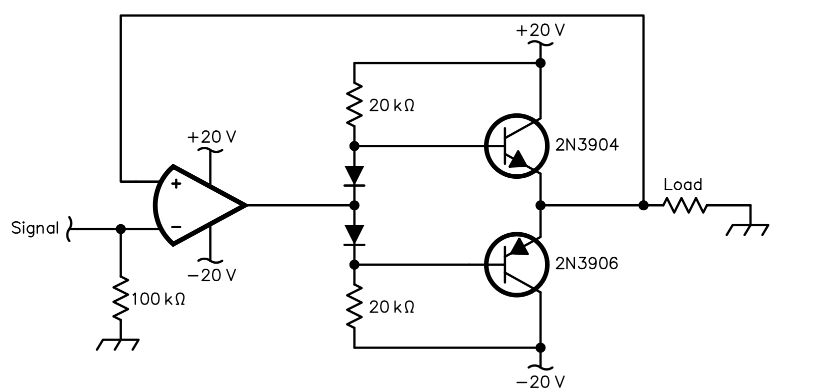

Fig 2—An op amp controls the operating point.

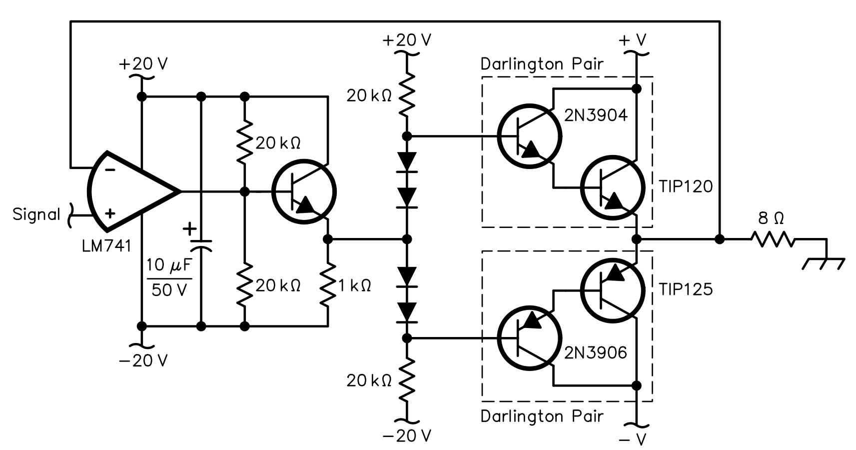

Fig 3—Output current is increased with power Darlingtons.

0°-phase-shift response and introduces no difficulties when it follows the op amp.

The power supply for this unit need not be exotic, since the op amp rejects power supply ripple by about 90 dB, and the emitter-follower response is also essentially independent of power supply voltage. Balanced 20-V supplies can allow ±18 V to be applied to the load. The 20 V can be provided by two full-wave bridge rectifier supplies using two 24-V, 2-A transformers. 24-V transformers will produce about ±32 V with a capacitive input filter or ±20 V with a choke-input filter. A supply of ±32 V is too high for the LM741, but the output voltage can be dropped with a bypassed resistor. A capacitor-input filter does not make the most efficient use of transformers and rectifiers (because the current's peak-to-average ratio is high), but capacitor-input filters are certainly the least-expensive option. A couple of resistors can drop the 741 supply voltage to about ±20 V. The absolute value of the voltage is not critical, but it does limit the maximum power output. With ±20 V to an LM741, the maximum output would be ±18 V into 2 kΩ.

The drive current required by the TIP120/TIP125 is on the order of 20 mA—a little high for an LM741, but not for a 2N3904 operated as a class-A emitter follower.

A garden variety IC and a pair of complementary Darlingtons can make a moderate-output class-B audio power amplifier that will satisfy critical high-fidelity listeners. The parts are available from Radio Shack. Since non-critical parts are used, the junk box can be called on to hold down the costs. □□

A Synthesized Down-Converter for 1691-MHz WEFAX

Here's a down-converter to receive WEFAX images. It uses a PLL/VCO LO instead of the traditional crystal oscillator and multiplier chain. A wide-range LO makes the converter useful for many amateur bands. Tune-up is very simple, and the price—with new parts—is around $100.

By Jim Kocsis, WA9PYH

This microwave down-converter can be used to obtain weather satellite imagery from several geosynchronous satellites. The satellite names are GOES 8, GOES 10 (USA), GOMS (the former USSR), GMS (Japan) and Meteosat (European Space Agency). They all transmit pictures on 1691 MHz (HRPT transponder). Several photos show sample imagery.

Instead of using multipliers to obtain the relatively high-frequency LO signal, this down-converter generates the LO directly at 1553.5 MHz. This converts the 1691-MHz signal from a five-foot dish/feed-horn assembly and LNA to an IF of 137.5 MHz. A simple VHF scanning receiver and input card (or sound card) in your PC complete the setup. A block diagram of a complete system is shown in Fig 1.

By changing the crystal, RC loop-filter values and VCO, the circuit described here can be used to convert many other microwave frequencies to IFs that can then be received on simple receivers. Ham bands that can be converted to lower frequencies using this circuit include 902, 1296 and 2300 MHz. The oscillator I used covers 512-2800 MHz, so by using low- and high-side injection, this board can down-convert signals ranging from 400-3000 MHz to the VHF range.

For many years, I have sporadically tried to make a down-converter for WEFAX service but was never able to get the required output level or purity in a 1553.5-MHz LO. Brian, NQ9Q, who also dabbles in weather satellite imagery, told me several years ago: "The hardest part of the down-converter will be the local oscillator." He was right—until now. A new $12 IC from Motorola changed all that.

Technical

Fig 2 is a schematic diagram of the converter. The 1691-MHz signal enters the board at J1 and is amplified by U1. From there, it goes to mixer U2. The LO consists of the PLL/crystal-reference oscillator IC (U3), the VCO (U4), the loop filter R1/C1 and U5, a low-noise dc amplifier with a gain of two. The LO output mixes with the desired signal and is converted to

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137.5 MHz. This is the preferred IF because receivers used to obtain satellite pictures from polar-orbiting (APT-transponder) satellites usually tune to this frequency.

The PLL is locked when 6.068 MHz × 256 = 1553.5 MHz. The PLL chip is the workhorse of this circuit. It divides the VCO output (somewhere between 1300 and 2000 MHz) by 256, compares it to the on-board crystal oscillator (6.068 MHz) and outputs a series of pulses. Those pulses are integrated to produce a VCO-control voltage that keeps the VCO at exactly 1553.5 MHz. The PLL can output a maximum of 5 V, so U5 (a low-noise op amp) raises the level—approximately 8.3 V maximum—to tune the VCO.

Separate voltage regulators are used for the PLL and VCO to keep any noise in one stage from getting into the other. If any noise from the PLL power line were to appear on the VCO supply line, it would modulate the LO signal, adding noise to the incoming satellite signal. U7, an LM317, provides a clean, regulated 8-V dc source for the VCO. The R1/C1 loop filter integrates the phase-detector pulses and removes reference modulation, resulting in a quiet LO signal.

VCOs are known for phase noise that tends to mask weaker signals. The R1/C1 filter values must be chosen to minimize this phase noise. Inside the loop bandwidth, the loop actually cancels VCO phase noise in proportion to the loop gain. The values shown are *not* optimum. An Excel spreadsheet is available from Motorola to help determine optimum R1/C1 component values. I spent many hours trying to

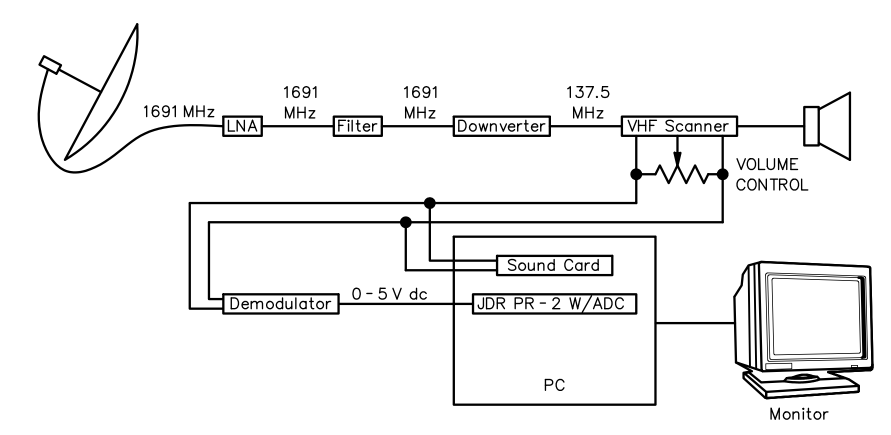


Fig 1—Block diagram of down-converter.

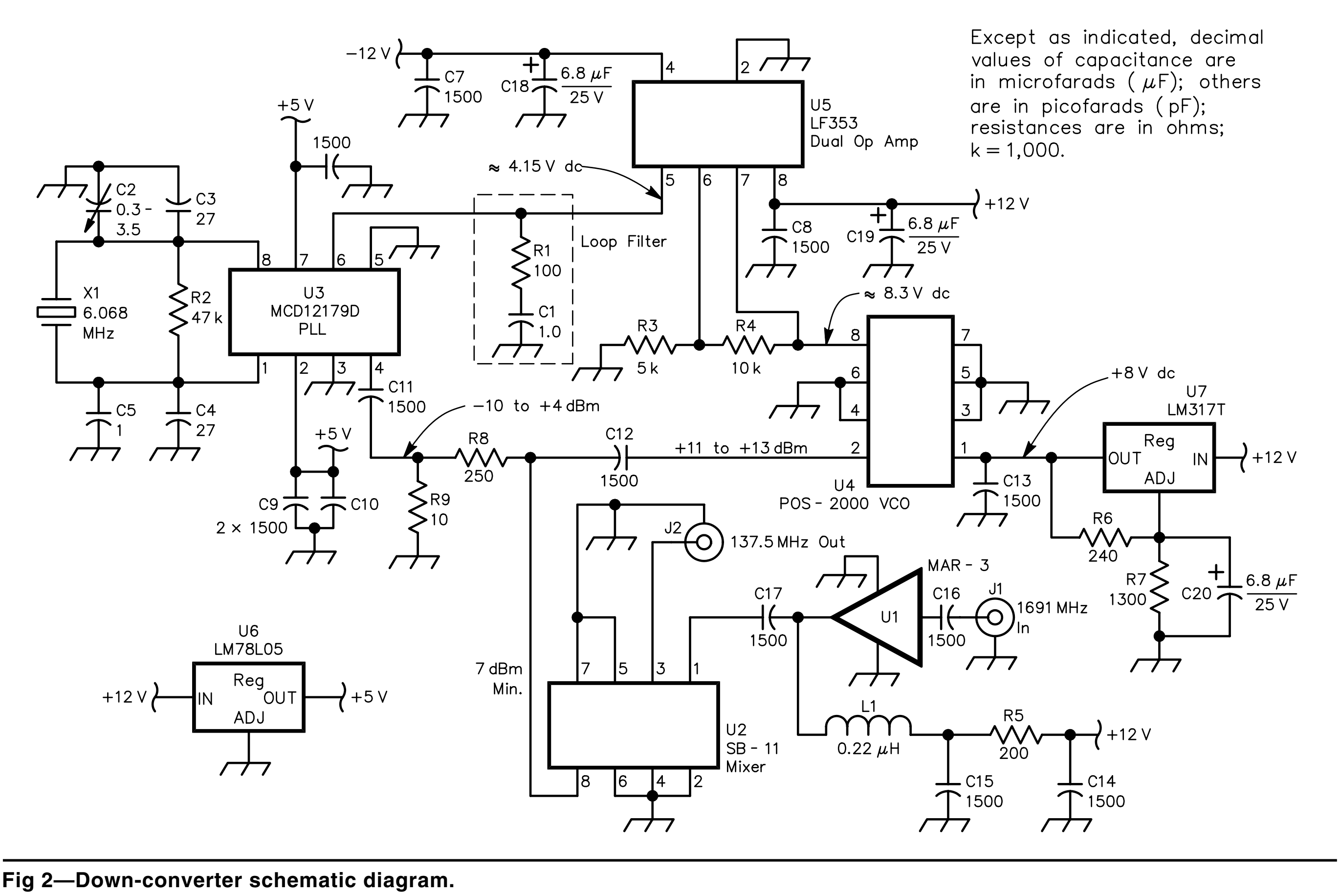


Fig 2—Down-converter schematic diagram.

obtain the proper values with this software, and as yet have not been able to figure it out completely. The values in the schematic were determined by the "cut-and-try" technique. I welcome anyone who can help me to obtain the optimum values with this software! A spectrum analyzer plot of phase noise is shown in Fig 3. Although it is not as quiet as a crystal-oscillator/multiplier circuit, it is quiet enough to yield very clear WEFAX pictures.

An RF filter is needed at the input to this down-converter. Without it, the signal is buried in noise because both the desired signal and the image noise appear in the IF. The filter I used appears in the reference of Note 1. (I'm working on a new down-converter board that incorporates this filter and other features on one board—more details at the end of this article).

Construction

The file containing the artwork is available from me (if you include an SASE diskette mailer and diskette) or from the ARRL *QEX* Web site.[2] I used "Press-N-Peel" to make the actual board. See the reference of Note 1 for a full description of how to make the board using Press-N-Peel.

The MMIC (MAR-3) mounts more efficiently with a 3/32-inch clearance hole drilled in the board, so that the input and output leads lie flat on the board surface. The input lead (dot is near the input) and output lead come straight out and are soldered directly to the traces. The two ground leads are bent straight down at the body and then along the ground-plane side of the board. See Fig 4.

A microwave circuit like this requires short, low-inductance feed-throughs to the ground plane near all ground pins of the PLL IC, the regulator ICs, and at the grounded end of all bypass capacitors. These ground connections should be soldered onto the board before other components because they require more heat than most other components. Locations of these feed-throughs and all components are shown in Fig 5. Keep in mind that you can't have too many ground points, but too few of them can cause problems. A ground feedthrough that is located too

[1]Notes appear on page 54.

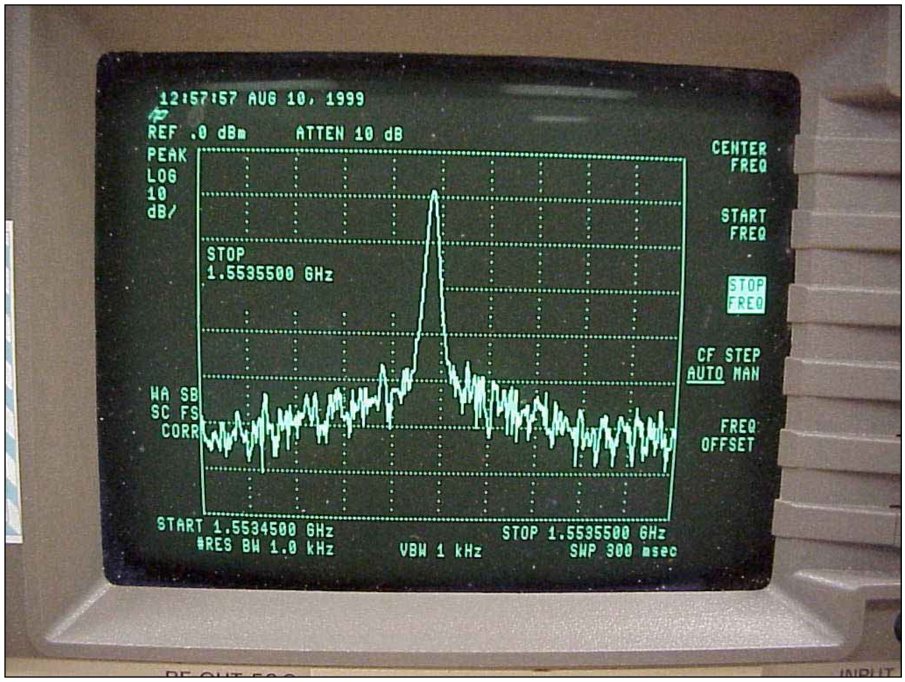


Fig 3—Spectrum analyzer plot showing close-in phase noise of LO.

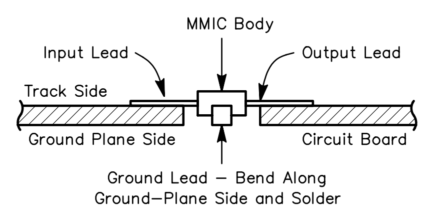


Fig 4—MMIC mounting details.

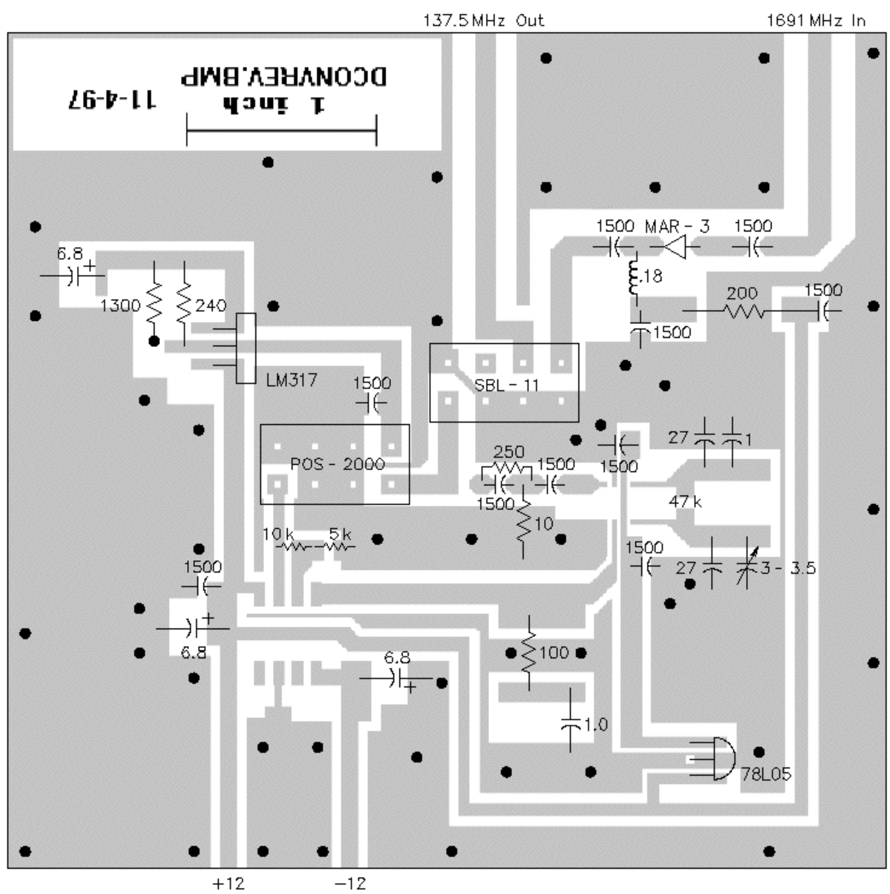


Fig 5—Parts placement diagram.

far from an IC can make a circuit oscillate or can reduce the effectiveness of a bypass capacitor. I used a 0.043-inch drill and #20 to #26 AWG solid wires to make my feedthroughs.

The POS-2000 VCO and SBL-11 mixer need several of their pins bent to make contact with various circuit traces. The remaining ground leads pass through holes in the board and are soldered to the ground plane.

I elected to use a few leaded components because I didn't have the required values in surface-mount parts. These leaded components are at noncritical locations: the op amp U5, loop-filter components R1, C1, R5, R7, R8 and L1. They should all have their leads formed as shown in Fig 6.

The PC-board artwork is shown in Photo F. Note that I have trimmed off some of the board to fit the chassis box that houses all other components of my system.

Caveats

Remember several things while constructing this unit:

1. A complete unit is shown in Photo E. Note the change in the artwork at U2 pin 3. The original artwork was wrong so I cut a track and corrected it. The circuit-board artwork provided by me and at the ARRL Web site includes those corrections.

2. The circuit side of the board should be completely tinned with a very thin layer of solder before mounting any components. Make sure the layer is smooth; a bump of solder where an SMT part lies will lift some of the leads off the board and make the part difficult to attach. Solder all components with a small soldering iron. I used a 25-W iron element and a 0.070-inch round chisel tip. Practice soldering inexpensive surface-mount parts (such as the resistors or capacitors) on a dummy board until you feel comfortable doing it and can do it quickly so the components don't overheat. They are tricky to solder and can stick to the iron tip if not held in place by a toothpick.

3. The crystal, VCO and mixer should be installed last, since they are the most expensive components.

4. Static discharge can damage the active components, such as the PLL, regulators, VCO, op amp, MMIC and so on. Use a grounded workstation, wrist strap and soldering-iron tip when installing these parts. Also, ground the PC board itself.

5. The dot on the bottom of the PLL IC does *not necessarily* denote pin 1. The beveled top edge of the IC is the side with pins 1, 2, 3 and 4. The data sheet doesn't make a big point of pin 1's location; it just shows a double line on one side and a single line on the other. The double line is the beveled edge. I learned all this the hard way: I installed the IC backwards and the chip was destroyed. I had to *very* carefully remove the damaged part without removing any trace with it. (That was $12 gone and a few more gray hairs!)

6. The VCO and mixer are marked much more clearly, but you should be careful when you bend the leads to fit the PC tracks. Pin 1 of each is identified by blue around it where it exits the body. The letters "MCL" (MiniCircuits Labs) are marked on top directly over pin 2. These two components are a lot more rugged than the PLL IC, but their leads can only be bent a few times before they break off. To paraphrase a saying: Check it twice and bend/solder it once (carpenters measure twice and cut once).

7. Don't substitute another part for the LF353 op amp. Since it is in the loop, I selected it for its low-noise char-

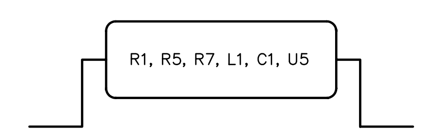


Fig 6—Lead-forming guidelines.

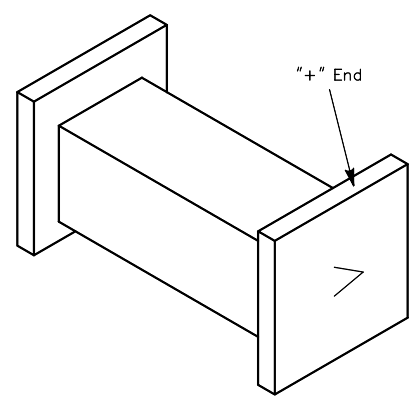


Fig 7—Tantalum-chip capacitor marking.

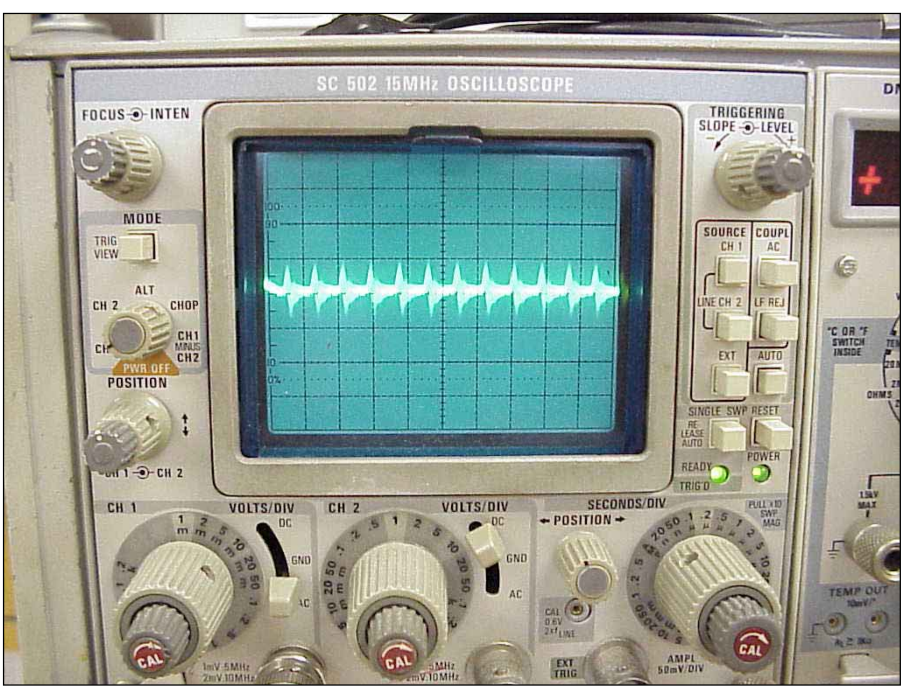


Fig 8—The PLL IC pin-6 waveform.

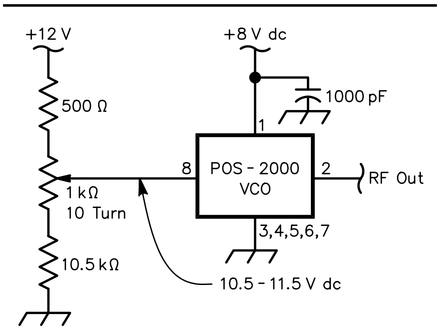


Fig 9—Test oscillator circuit.

acteristics. Any noise it generates will show up as noise on received signals.

8. The three, 6.8-µF power-supply-filter chip capacitors that I used have their positive ends marked with a small point; see Fig 7. Reversing them will apply the voltage backwards, destroying them. Yes, I did this, too. They short, and then smoke. You'll know if you install them backwards the moment you apply power! Electrolytic chip capacitors made by other vendors may be marked differently. Verify their polarity before soldering them onto the board. To keep the cost down, you can also use radial-lead electrolytic capacitors at these three locations. I had a large stock of 6.8-µF chip capacitors, so I used them.

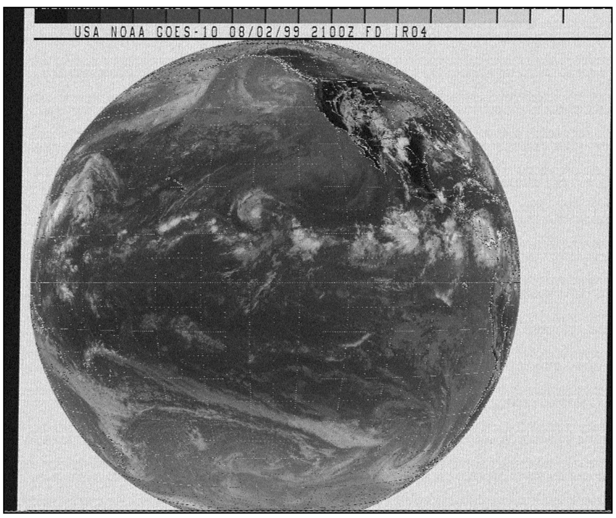
Check-out and Alignment

Tune up consists of adjusting a trimmer capacitor on a 6 MHz crystal oscillator. There are no other adjustments.

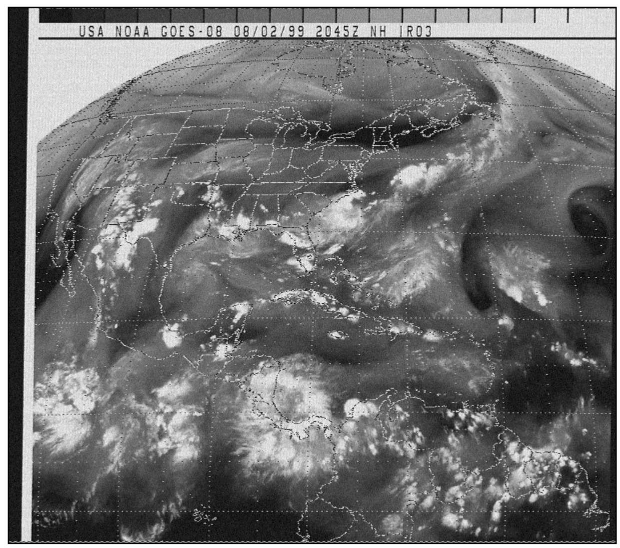
Correct operation is initially indicated by the board drawing approximately 72 mA from the +12-V supply and 3.5 mA from the +12-V supply. Correct operation of the LO is indicated by approximately 4.15 V dc at pin 6 of the PLL IC (as read with a DVM). A sample of the waveform at this point is shown in Fig 8. Twice this voltage should appear on pin 7 of the op amp since it has a gain of two. Adjustment of the crystal frequency is accomplished by varying trimmer capacitor C2. Monitor the frequency using a sensitive frequency counter attached to pin 4 or 5 of U3, using a very low-capacitance probe.

Approximately 400 mV of signal is available at either pin. The frequency should be close to the frequency marked on the crystal. Excessive probe capacitance will change the frequency more than the trimmer. Just a few picofarads are allowed. Remember that a one-hertz change in the crystal frequency results in a 256-Hz change in the LO.

The bandwidth of the satellite signal is only ±9 kHz. If the crystal is off more than 80 Hz and your receiver bandwidth matches the signal deviation, you won't even hear the satellite! A good way to make sure that the crystal is very close to the desired frequency is to use an HF receiver set to



(A)



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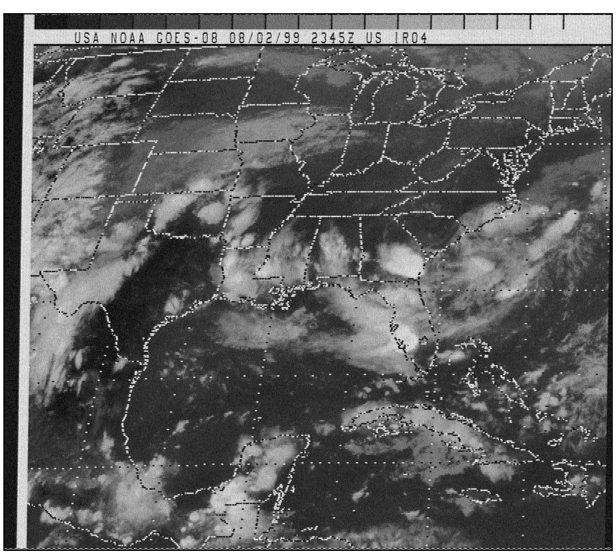
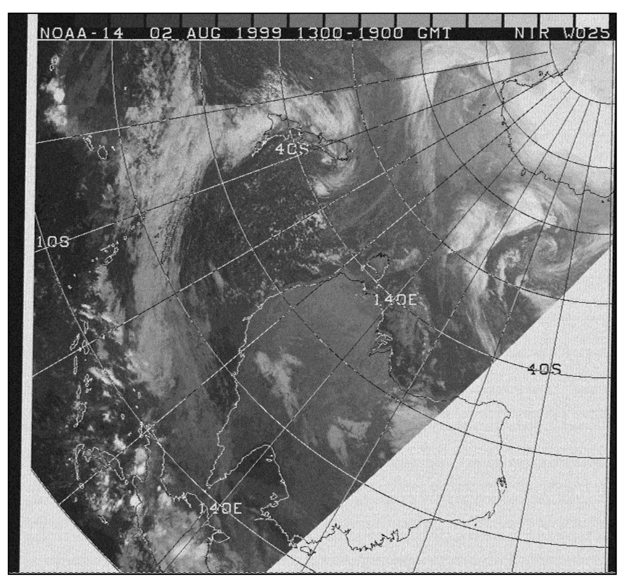


(C)



(D)

Photos A-D—Sample HRPT WEFAX images

6.06836 MHz. Change from USB to LSB and back while listening to the tone of the carrier. Adjust trimmer C2 until both tones are the same. The crystal oscillator's signal isn't very strong even with a short lead draped across the circuit board. In fact, it is so weak that at night, a shortwave station covered it up with only two feet of wire on my receiver's antenna jack! Propagation is worse during the day at this frequency, so do this check in the daytime.

Make the final trimmer-capacitor setting when you can hear a WEFAX signal on your VHF receiver. Tune for zero discriminator output if your receiver uses a discriminator. If possible, borrow a spectrum analyzer to view the phase noise of the LO. It should look very similar to Fig 3. If the analyzer has AM and FM detectors and audio output available, listen to the LO to make sure it has no hum, hiss or noise on it. I was fortunate enough to have access to a very powerful HP analyzer with both AM and FM detectors and a minimum resolution bandwidth of 1 kHz. To determine whether the converter's mixer is working, you can use a signal generator at 1691 MHz—or its image at 1416 MHz, *without* the RF filter installed. If you don't have access to a signal generator, you can use a second POS-2000 as a signal source. Approximately 11 V on the VCO control line is needed to obtain 1691-MHz output. A circuit for this test oscillator is shown in Fig 9.

Parts

The input connector is an SMA type that has very low loss at these frequencies. I obtained several end-mount units at a hamfest for a few dollars each, but I haven't seen them since—even at Dayton! New units are available from Newark, but they cost over $10: Too expensive for me! A type that is available at hamfests mounts like SO-239 flange-mount connectors. They mount on the ground-plane side of the board rather than the edge.

The PLL IC is available from Newark Electronics.[3] Be sure to get the data sheet for it from Motorola at **http://mot-sps.com/cgi-bin/get?/books/dl110/pdf/mc12179rev*.pdf**. The data sheet is in PDF format. The Excel software for selecting the loop filter components is called *DK-307/D* and is available from Motorola at their Literature Center.[4] You will need *Excel* version 5.0 or higher to run it.

The mixer, VCO and MMIC are available from MCL.[5] Purchase a few

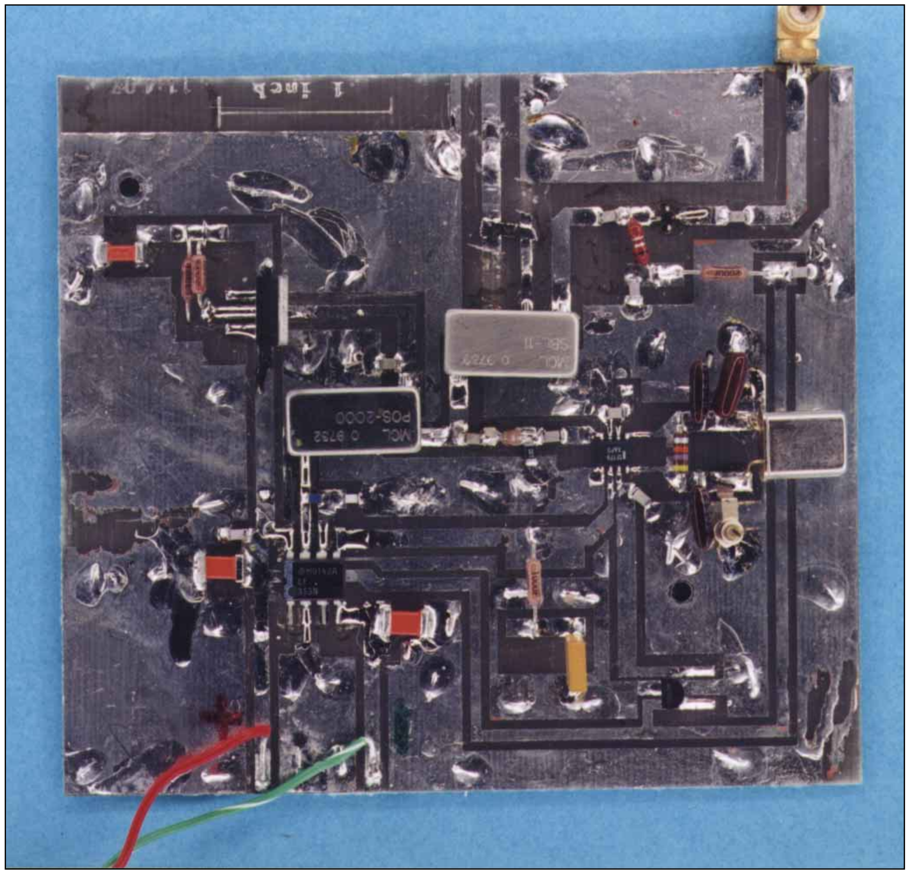


PHOTO CREDIT: CHRIS HALL.

Photo E—Complete WEFAX down-converter

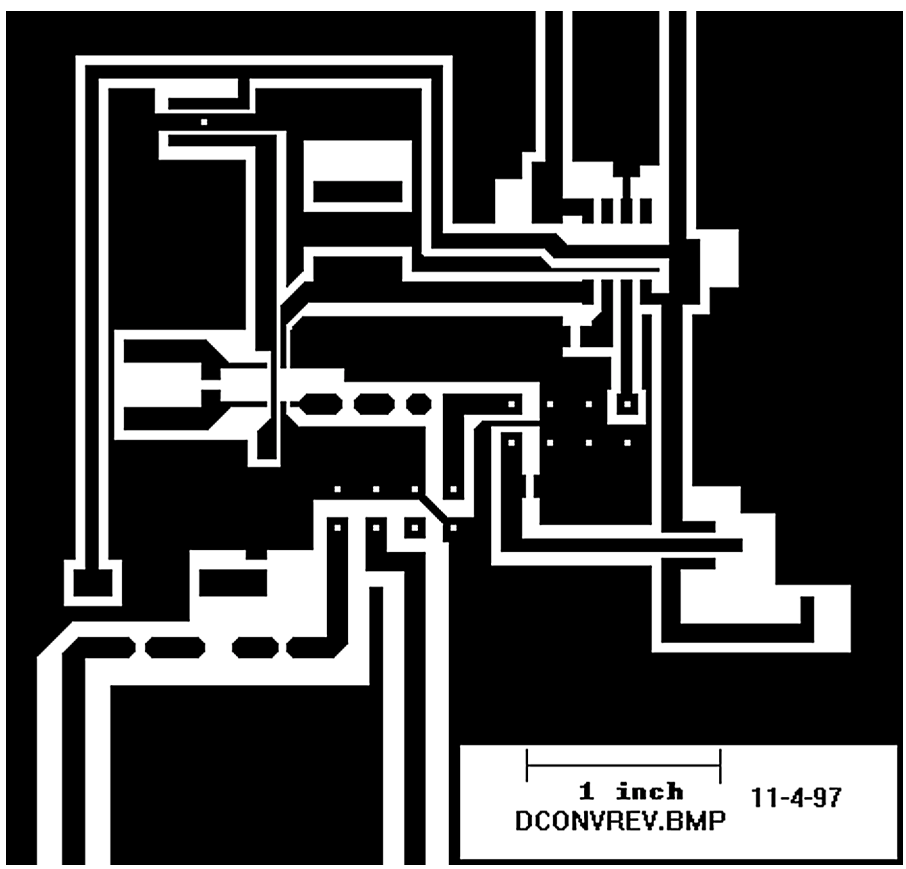


Photo F—PC-board artwork.

spares of the MMIC—you can use them in many other projects. I recommend buying at least one extra VCO for testing or as a spare. Be sure to ask for their catalog and *VCO Designers Handbook*. Both have very good practical-guideline sections. The VCO theory section is especially good! Both documents are free.

The voltage regulators, 6.8-µF power-supply chip capacitors and the op amp are available from Digi-Key[6] or at some hamfests. The 1000-pF chip capacitors, MMIC, chip resistors, inductor and GaAsFET for the HP LNA (mentioned later) are available from Downeast Microwave.[7] Order extras of each value: They're inexpensive, and you can use them for other projects. You can also use the spares when you practice soldering surface-mount parts.

I have not been able to find a company that sells 0.062-inch double-sided glass/epoxy board. I found several large pieces at a hamfest for $10 each (enough for many projects). A search on the Web might lead you to a commercial source.[8] Make sure it is glass/epoxy and not phenolic. The glass/epoxy type has a greenish-white appearance while the phenolic type is light brown.

Improvements

There are many changes I want to make to improve this unit's performance. However, so that I could get the information on this new IC from Motorola to the amateur community quickly, this article only describes a basic downconverter. I have designed artwork for a unit with the RF filter and a crystal heater on the board. After I make these changes and verify that they work, I plan to add an LNA to this board. I will make this circuit available when I'm done.

There are five other manufacturers' VCOs that cover this range. I have the artwork for one of them if you're interested in trying it (the artwork and circuit are very different). These vendors are listed in Note 9. They are not nearly as rugged as the POS-2000, require a buffer MMIC and attenuator network and require much more careful soldering. The package/pin-out is very different and the cost is the same, but they have a much narrower frequency range. I built a downconverter using one of these VCOs and it worked very well. The LO's phase noise was 10 dB lower than when using the POS-2000(!), but something damaged every active component on the board after a few weeks operation. I suspect a spike from the power supply was the villain, since even the regulators were zapped!

To date I have only used JVFAX to display the pictures on my PC. I used a JDR Microdevices PR-2 PC prototyping board with an ADC0804 eight-bit A/D converter. I can provide information on this—send an SASE if you're interested. This interface requires that you use a demodulator. The demodulator I've used for the past 14 years is from an excellent article by Grant Zehr.[10] I plan to try a sound card with WXSAT software. It will eliminate the need for the interface card and demodulator.

Dish and Feed Horn

The dish I use is a homemade, five-foot mesh unit, and the feed horn is a two-pound coffee can.[11] *The Weather Satellite Handbook* by Taggart provides a lot of information on dish construction as well as many other aspects of weather satellite imagery.[12]

LNA

The LNA I use comes from an application note by Hewlett Packard (AN1076). It uses an inexpensive GaAsFET and MMIC (MAR-3) to attain a stated noise figure (NF) of around 0.7 dB—quite good at 1691 MHz. The gain of the circuit is stated at approximately 25 dB. I was not able to measure either parameter. The numbers are good but the real proof of performance is a clear picture. The LNA is so good that I am able to hear the signal without the dish. I just point the feed horn toward the satellite and hear the characteristic 2400 Hz audio signal! It's not clear enough to obtain a picture, but I can hear it. If you decide to build this LNA, be sure to read and understand all of the construction details. The cost of this LNA—using all new parts—is in the $25 range. As pointed out in later versions of the same PC-board layout for similar GaAsFETS, beware that the positions of the RF choke and resistor used to supply gate bias are incorrect on the layout. The RF choke should be closest to the gate lead. In addition, the trace at the 78L05 regulator is wrong. I can supply instructions to correct this small error. The circuit can be simplified and the cost reduced by not using the voltage-inverter (Maxim MAX1044ESA) circuit and instead using a 79L05 and the +12-V supply used in my downconverter. The GaAsFET is static sensitive. Use all appropriate anti-static precautions. Also, keep the soldering heat to a bare minimum—this part is *very* small.

I have described an LNA that is easy to build.[13] The parts are inexpensive and physically larger so they are easier to handle. The unit needs only +5 V dc. The NF isn't as good as the HP LNA (it's only 1.7 dB), but the gain is higher: 39 dB. Although gain isn't as important as NF. LNAs are also available from Downeast Microwave (Model 1691LNA priced at $65, see Note 7). If you don't feel comfortable working with very small parts (the GaAsFET), buy an LNA from Downeast!

Conclusion

In addition to the WEFAX weather imagery on 1691 MHz, other weather information (called EMWIN) is sent by GOES 8 on 1690.725 MHz. The data sent includes weather radar. Using this downconverter, you can receive these data by tuning your VHF receiver to 137.225 MHz. I have not decoded the data yet, but my scanner does stop at 137.225 MHz, indicating that a signal is present. The signal sounds like noise, but it is, in fact, 9600-baud data. A schematic of a decoder is shown at **iwin.nms.noaa.gov/emwin/wintip.htm**. (*Do not* precede this address with the typical "**www**"). Click "Tech Info," then "Demodulator Schematics," and "ZE DM-96 Demodulator Schematic." Free software is available at **www.weathernode.com** for Win 95, 98 and NT.

I had a lot of fun building this downconverter and learned a lot during its design and construction. If you build it, please write or e-mail me; I'd like to see your questions or comments. I would especially like to know if you were able to make the *Excel* software work and what values you found make the LO quieter. Happy soldering! Many thanks to my wife Yvonne for proofreading this article.

Notes

1"Defogging Microstrips—An Introduction to Microstripline Filters," *73 Magazine*, Oct 1999.

2You can download this package from the ARRL Web **http://www.arrl.org/files/*qex*/**. Look for 0300KOC.ZIP. It includes a parts list.

3Newark Electronics, 4801 N Ravenswood Ave, Chicago, IL 60640-4496; tel 800-463-9275, 773-784-5100; **www.newark.com/**.

4There's an on-line literature order sheet at **http://www.mot-sps.com/home/lit_ord.html** or you can telephone 800-441-2447.

5Mini Circuits Labs, PO Box 350166, Brooklyn, NY 11235-0003; tel 800-654-7949, 718-934-4500, fax 718-332-4661; **www.minicircuits.com/**.

6Digi-Key Corp, 701 Brooks Ave S, PO Box

677, Thief River Falls, MN 56701-0677; tel 800-344-4539 (800-Digi-Key), fax 218-681-3380; **www.digikey.com/**.

[7]Down East Microwave, 954 Rte 519, Frenchtown, NJ 08825; tel 908-996-3584, fax 908-996-3072; **downeastmicrowave.com/**.

[8]Circuit Specialists Inc, 220 S Country Club Dr #2, Mesa, AZ 85210; tel 800-528-1417, 480-464-2485, fax 480-464-5824 (8AM-6PM MST) **info@cir.com**. Their Web catalog shows a 6×9-inch two-sided epoxy or fiberglass board at **http://store.yahoo.com/webtronics/printed-circuit-board-supplies-printed-circuit-boards.html** as part #22-267—*Ed.*

[9]VCO alternatives: Z-Comm **www.zcomm.com**; Tetra **www.tetra.co.il**; Varil **www.varil.com**; PES **www.pesinc.com**; Champion Tech **www.champtech.com**.

[10]"The VIP: A VIC Image Processor," *QST*, Aug 1985.

[11]"Dish Antenna for Weather Satellite Images," *73 Magazine*, July 1995.

[12]R. Taggart, *The Weather Satellite Handbook* (Newington: ARRL, 1994, Order #4483).

[13]"A Low-Noise Amplifier for 1691 MHz," *73 Magazine*, July 1995.

Jim Kocsis is a test engineer at AlliedSignal Aerospace where he has access to a limited amount of test equipment that can be used at low-microwave frequencies. His interests include casual HF DXing, DX contests, Field Day, ragchewing on HF, antennas and homebrewing anything. He has been interested in weather-satellite imagery since 1985 and has decoded images from NOAA satellites, HF fax and now (finally!) GOES satellites. He has a degree in Physics from Indiana University (1976). He enjoys reading, music, some cooking/baking and bicycling. Jim was first licensed in 1964 as WN9LDB and in 1965 upgraded to WA9PYH (General). He now has an Extra class license.

RF

By Zack Lau, W1VT

A 2-Meter Input Circuit for an 8877

A classic construction project for serious VHF operators is a legal limit 2-meter amplifier using the Eimac 8877 triode. This tube is often available inexpensively on the surplus market. Just as importantly, designs exist in the both Bill Orr's *Radio Handbook* and *The ARRL Handbook*.[1, 2, 3, 4, 5] As many have found, however, the tunable coil used on the input circuit is tough to find. Thus, it is necessary to devise a suitable substitute.

The first step is determining the exact function of the original circuit. According to Bill Orr, the T network is designed to match 54 Ω in parallel with 36 pF to 50 Ω resistive. Edward Meade found a slightly different capacitance, 26 pF. Such a variation isn't surprising—this is why tunable networks are used. A high-Q network also isolates the exciter from variations in input impedance presented by the tube at different parts of the RF cycle. This is why final testing of the network is done under full power conditions—the impedance varies with power level.

At HF, the input matching network also forms a low-impedance shunt from the cathode to ground, but the parasitic capacitance of the tube is enough to perform this function at VHF. (This is why you often see π networks at HF.) Thus, you get a rough schematic like Fig 1.

The next step is to determine the known component values. If the original design is known to work well, it makes sense to copy of as much of it as possible, to minimize the chance for error. Inductance can be easily calculated with formulas or simulation programs. Thus, L2 is 145 µH.

With a sophisticated program like *ARRL Radio Designer*, one can use an optimizer to solve for the unknown values. Using ARD, a good match is obtained if L1 is 204 nH and C1 is 15 pF. This is shown in Table 2. Be aware that there may be many practical solutions. Other factors, such as bandwidth or insertion loss, may help determine which solution is best. The last step is devising a suitable substitute for the 204 nH inductor.

[1]Notes appear on page 58.

225 Main St
Newington, CT 06111-1494
zlau@arrl.org

Table 1

Air core inductance values for #14 wire and 1.38 inches of total lead length

Inside Inductance (nH)	*Turns*	*Diameter (Inches)*	*Length (Inches)*
192	4	0.562	1.00
203	4	0.562	0.85
220	4	0.562	0.70
232	4	0.562	0.60
196	3	0.75	1.0
205	3	0.75	0.85
216	3	0.75	0.70
236	3	0.75	0.55

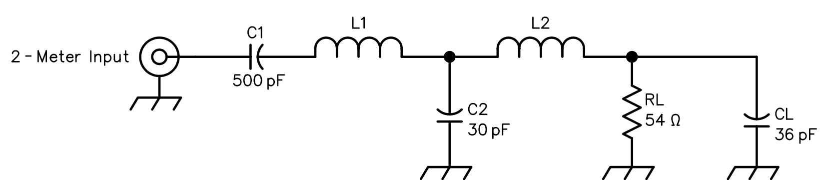


Fig 1—Model of 8877 input circuit based on project description.

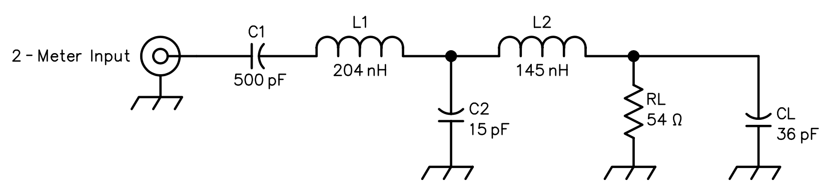


Fig 2—Model of 8877 input circuit "tuned up" with the ARD optimizer.

Table 2

Freq MHz Input	*MS11 dB Input*	*SWR*
140.000	–9.98	1.93:1
140.500	–10.94	1.79:1
141.000	–12.09	1.66:1
141.500	–13.47	1.54:1
142.000	–15.20	1.42:1
142.500	–17.46	1.31:1
143.000	–20.63	1.21:1
143.500	–25.87	1.11:1
144.000	–41.86	1.02:1
144.200	–39.98	1.02:1
144.500	–28.83	1.08:1
145.000	–21.89	1.17:1
145.500	–18.07	1.29:1
146.000	–15.42	1.41:1
146.500	–13.42	1.54:1
147.000	–11.81	1.69:1
147.500	–10.48	1.85:1
148.000	–9.35	2.03:1
148.500	–8.39	2.23:1
149.000	–7.55	2.44:1
149.500	–6.82	2.68:1
150.000	–6.17	2.93:1

Both designs specify adjustable inductors with white tuning slugs. According to J. W. Miller documentation, white is a color code for slugs made out of IRN 9, recommended for use between 50 and 200 MHz. In case you are lucky enough to come across a source of these obsolete coil forms, white coil forms are indicated by a "–4" suffix.

```
*Input T network for an 8877 at 144 MHz
*Page 31-52 1991 ARRL Handbook
BLK
SLC 1 2 L=20NH C=500PF Q=500
f=144MHZ
IND 2 3 L=?200NH 204.389NH 2UH? Q=200
F=144MHZ
CAP 3 0 C=?5PF 15.2547PF 25PF? Q=500
F=144MHZ
ind 3 4 l=145nh Q=815 F=144MHZ
*simulate 8877 input
res 4 0 r=54
cap 4 0 c=36pf
input:1por 1
end
freq
step 140mhz 150mhz .5mhz
144.2MHz
end
opt
input ms11 .01 lt
end
```

Fig 3—Amateur Radio Designer Simulation of the 8877 amplifier input circuit.

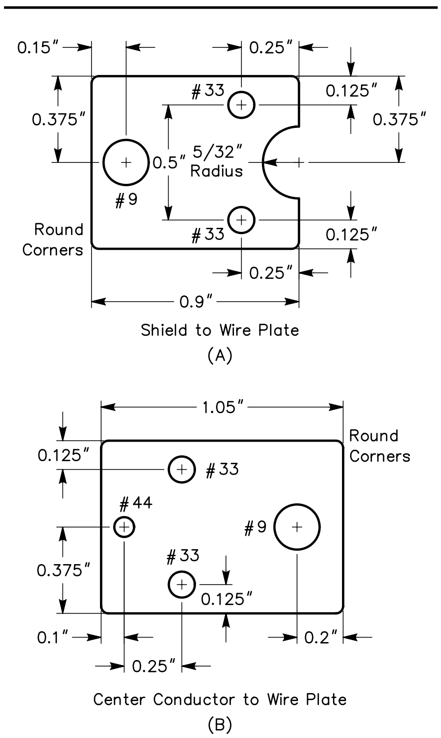


Fig 4—(A) Coax shield to wire plate made out of 0.020-inch brass or copper sheet. (B) Coax center conductor to wire plate made out of 0.020-inch brass or copper sheet.

Surplus Sales of Nebraska is a good source of slug tuned coils.[6] An inexpensive solution may be to substitute an air-core inductor, the value of which is tweaked by spreading or compressing turns. This can be rather time consuming to adjust, since it is necessary to turn off the amplifier and remove the voltages before safely tweaking the amplifier. Table 1 gives some examples of how the inductance can vary as an inductor is expanded or compressed.

A BNC Feed Insulator for Wire Antennas

Do you need a center insulator to go between a piece of coax with a BNC male connector and some antenna wires? I devised this simple design for my portable QRP HF station. It has a female BNC connector that connected to a pair of #10-32 screw posts. A pair of wing nuts makes it easy to change wires in the field without tools. The use of machined flat metal plates instead of round wires results in a rugged, compact assembly. This might be a good project for someone who wants to learn machine shop skills—the difference between a good job and a great one is quite noticeable when parts are attached to a clear piece of plastic. Small Parts sells everything for the feed insulator except the BNC connector.[7]

I found a bargain-priced bag of #10 wing nuts at a hamfest, so choosing the size was easy. I've also had good luck with #8 wing nuts on antennas, but my fingers are smaller and stronger than those of most hams. (I find #6 wing nuts are too small, so I'd avoid them unless you plan to use a tool to tighten them. That defeats the purpose of using wing nuts, though.)

If you want to use larger wing nuts, increase the spacing between the screws and the BNC connector. The

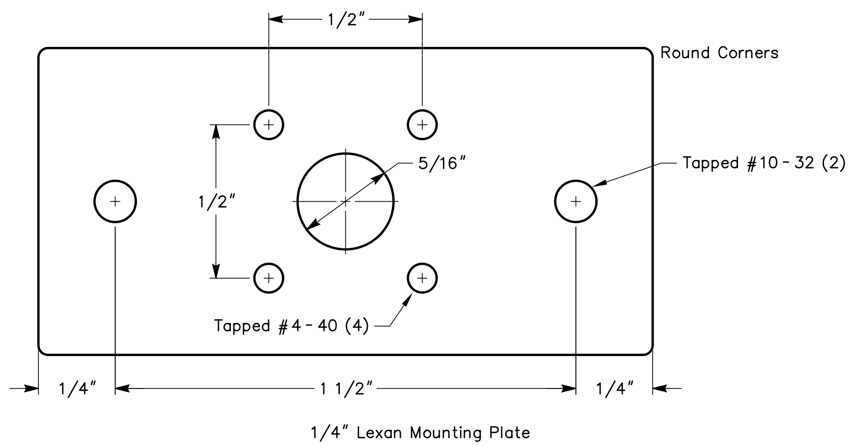


Fig 5—Polycarbonate mounting plate (1/4-inch thick).

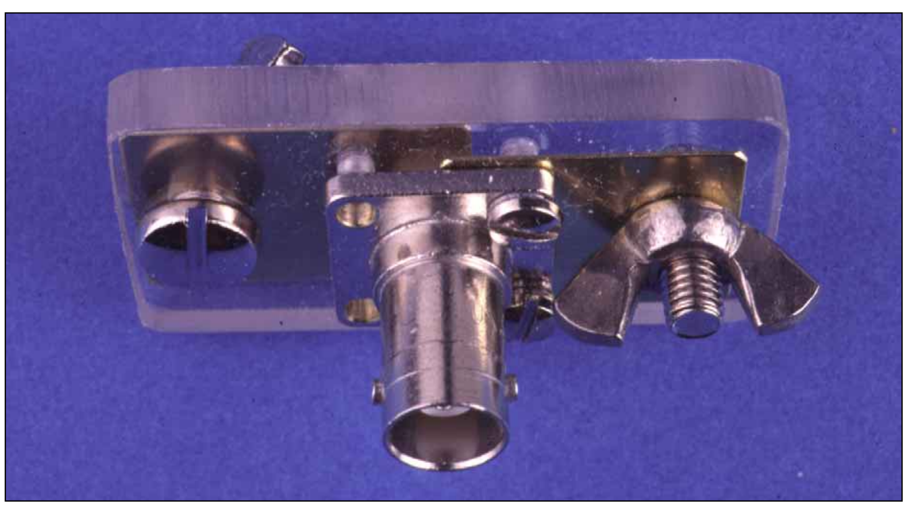


Fig 6—Side view of the BNC to wire transition.

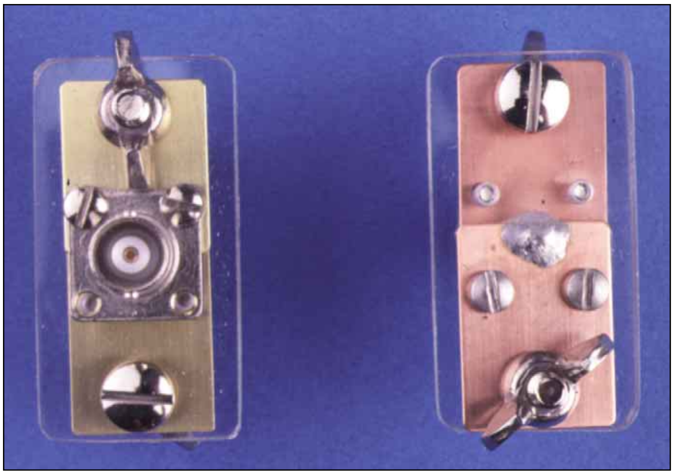


Fig 7—Top and bottom views of the BNC-to-wire transition.

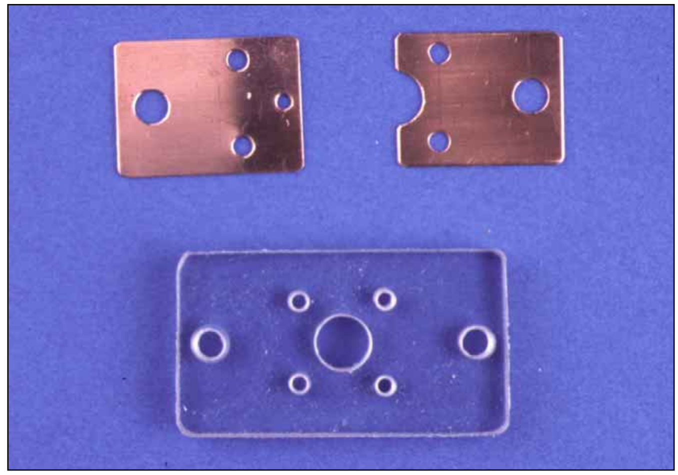


Fig 8—The machined parts before assembly.

1.5-inch spacing between the wing nuts is designed to accommodate a wingspan of 0.82 inches. Otherwise, the connector will prevent the nut from turning.

There is a trick to making the metal plates. In order to put a $^{5}/_{32}$-inch semicircle at the edge of a plate, I drill a $^{5}/_{16}$-inch hole in a sheet and then cut the sheet in half, forming two plates. I used a metal shear to cut the plates. It should also be possible to cut the sheet metal with tin snips or a nibbling tool—both have their disadvantages. A tin snip tends to bend sheet metal, while a nibbling tool is slow and tedious. As shown in Fig 4, I've made plates out of both brass and copper.

I strongly recommend that you use Lexan or polycarbonate plate. Not only is it rugged (it's used to make bullet-proof windows), but it is UV stabilized. It doesn't degrade when exposed directly to sunlight like cheaper plastics. I used a drill press to accurately drill the holes. Interestingly, the #3-56 tapped holes on an UG-290 connector are small enough to be used as drilling guides for a #43 drill bit. I've not used this technique myself, however, I've had good results using scribed lines as a guide. The holes in UG-290 connectors must be enlarged with a #33 drill bit to pass #4-40 screws.

After drilling and cutting the metal and plastic plates, I rounded all the sharp edges. This is important—you don't want to cut yourself when using it out in the field. I also polished up the metal plates with steel wool. Clean surfaces are important for good electrical contact. It aids in soldering to the center pin of the BNC connector. I used a 100-W soldering iron—a big iron with a large chisel tip makes it possible to quickly solder the connection without melting the plastic.

Chuck Hutchinson, K8CH, suggests placing a flat washer and an internal-tooth lock washer between the metal plate and the wing nut. This allows you to tighten the wing nut without a metal surface scraping against the wire. I've not used this extra hardware—this way I have less stuff to lose in the field. A little bit of metal scraping isn't necessarily bad—it can help make a good electrical connection. Strain relief is another area that's ripe for improvement—wires last longer if you distribute the area of stress. As anyone who has played with paper clips has learned, wire can easily be broken if flexed back and forth at the same location. This becomes important if you are designing a center insulator for permanent use, as opposed to temporary portable installations.

Parts List

1	1×2×$^{1}/_{4}$-inch Lexan or polycarbonate sheet
2	#10-32 stainless steel wing nuts (Small Parts R-PWNX-032)
2	#10-32×$^{5}/_{8}$-inch machine screws (they serve as binding posts for the wing nuts)
2	#4-40×$^{3}/_{8}$-inch machine screws (to attach the connector and shield plate to the plastic)
2	#4-40× $^{3}/_{16}$-inch machine screws (to attach the center conductor plate to the plastic)
4	#4-40 lock washers
1	0.75×3×0.020-inch brass or copper sheet
1	UG-290 BNC connector
2	#10 flat washers
2	#10 internal-tooth lock washers

Notes

1"2-kW PEP Amplifier for 144 MHz," *ARRL Handbook* (Newington, CT: ARRL, 1988-1994), pp 31-51 through 31-53.

2"2-kW PEP Amplifier for 144 MHz," *ARRL Handbook* (Newington, CT: ARRL, 1987) pp 31-59 through 31-61.

3"2-kW PEP Amplifier for 144 MHz," *ARRL Handbook* (Newington, CT: ARRL, 1986), pp 31-67 through 31-69.

4Bill Orr, W6SAI, Ed., *Radio Handbook* (Indianapolis, IN: Sams, 23rd edition, 1986) "A High Performance 2-Meter Power Amplifier," pp 18-2 through 18-7. The 21st edition has the article as well, on pages 22.79 through 22.85. This earlier edition gives the impedance as 54 Ω with 26 pF of capacitance, while the later edition uses 36 pF.

5Edward J. Meade, Jr, K1AGB, "A 2-KW PEP Amplifier for 144 MHz," *QST*, Dec 1973 pp 34-38; Jan 1974, pp 26-33.

6Surplus Sales of Nebraska, 1502 Jones St, Omaha, NE 68102; tel 402-346-4750, 800-244-4567; grinnel@surplussales.com; www.surplussales.com.

7Small Parts Inc, 13980 NW 58th Ct, PO Box 4650, Miami Lakes, FL 33014-0650; tel (orders) 800-220-4242 (M-F, 8:00 AM to 6:30 PM ET); fax 800-423-9009; http://www.smallparts.com.

Upcoming Technical Conferences

Southeastern VHF Society Conference Committee Members

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"Four Days in May" 2000 QRP-ARCI Conference

QRP Amateur Radio Club, International (QRP-ARCI) proudly announces the fifth annual "Four Days In May" QRP Conference commencing Thursday, May 18, 2000—the first of four festive days of 2000 Dayton Hamvention activities. Mark your calendar and register early for this not-to-be-missed QRP event of the new millennium. Amateur Radio QRP presentations, workshops and demonstrations will be the focus of the full-day (8 AM to 4:30 PM) Thursday QRP Symposium to be held at QRP-ARCI headquarters—the Days Inn Dayton South. Topics include:

- "The 2N22/6"—Jim Kortge, K8IQY—Last year's first-place award winner!
- "The Super Gainer Regen Receiver"—George Dobbs, G3RJV
- "Do the VOMBA: The Dance of Vertically Oriented Multi-Band Arrays"—L. B. Cebik, W4RNL

Many more speakers and great surprises are yet to be announced! Symposium Attendance requires a contribution of $15. There will be a Thursday evening author social from 7 to 11 PM (no charge). It's a great chance to meet and talk with the QRP Symposium speakers.

QRP ARCI Awards Banquet

Friday evening starts with the annual QRP-ARCI Hall of Fame Awards Banquet ($25 per ticket) honoring QRPers who have made major contributions to QRP and Amateur Radio. There will be fantastic door prizes, a great speaker and tons of fun—be there! Following the banquet, there will be a "no charge" QRP vendor social. Come review the latest equipment and talk with the vendors.

Building and Design Contests, Radio Show

Saturday evening provides time for QRPers from around the world to socialize. Show off your projects and collections at the Radio Show! The evening culminates with three building contests:

- Open category—bring your latest homebrew or kit project.
- 1-volt challenge—build a rig that operates on 1.5 V!
- 48-volt challenge—build a tube transceiver!

Judges will select winners from all categories. Winners will receive prizes, and their projects will be subjects of future feature articles in the *QRP Quarterly*.

Hotel Reservations

Hank Kohl, K8DD, has arranged for a Dayton-QRP rate at the "re-*newed*" Days Inn Dayton South. Rooms are $72 per night plus tax (pending confirmation from Days Inn)—with as many occupants as will fit in a room! For reservations, don't call the hotel yet, but contact Hank at 1640 Henry, Port Huron, MI 48060-2523. Hank can be reached via e-mail at **k8dd@arrl.net**.

Registration

For more information and a registration form, visit the QRP-ARCI Web site at **http://www.qsl.net/k4zol/fdim/index.html**. Send payment for seminar ($15 each) and banquet ($25 each) registration to Ken Evans, 848 Valbrook Ct, Lilburn, GA 30047. Make checks payable to QRP ARCI.

2000 Southeastern VHF Society Technical Conference

The fourth annual conference of the Southeastern VHF Society will be held on April 14-15, 2000. The conference location will again be the Marriott Hotel at Windy Hill in Marietta, Georgia. The hotel and conference center is conveniently located northwest of Atlanta, Georgia, at exit 110 of Interstate 75. The room rates for attendees will again be $69 per night, but that does not include breakfast this year. For reservations, call Marriott at 1-800-228-9290. For commercial air transport, Atlanta-Hartsfield Airport is about 45 minutes away by Atlanta Airport Northside Shuttle. The hotel is also equidistant from three local airports for private aircraft. They are Peachtree-DeKalb, Brown-Fulton County and McCollum-Cobb County airports.

Complete conference details, including a tentative schedule and nearby attractions are available at the SVHFS Web site **http://www.svhfs.org/svhfs/svhfs2000.htm**. The Conference Committee Members table lists key individuals and their e-mail addresses. □□

Letters to the Editor

Technology and the Future of Amateur Radio (Nov/Dec 1999)

Doug,

Thanks for your thoughtful article "Technology and the Future of Amateur Radio" in the Nov/Dec 1999 *QEX*. After surveying the articles in that issue, I find them to be of great value. One thing that bothers me is that many of the articles in *QEX* are limited to those with considerable electronics education. I don't mean that they should be written otherwise; I am commenting that it seems a shame that many thousands of hams seem to have a feeling of being "shut out" as a result of a lack of background. It seems to me that it would be very worthwhile to develop some means of upgrading people from older technology to new skills. Education is really the key to the future. Perhaps that is too big a goal. Anyway, you are doing a good job and I enjoy both *QST* and *QEX*.—*Frank Merritt, VE7FPM, 1851 Meredith Rd, Nanaimo, BC, V9S 2M6, Canada;* **fmerritt@nanaimo.ark.com**

Where Are the Schematics?

◊ I was disappointed in two of the articles in the Nov/Dec 1999 *QEX*. The first was PA0HRK's article on the noise-figure meter. I have been in correspondence with Luis Cupido and had already begun to build the meter of his design. Being blessed with two spectrum analyzers, this was the logical choice.

Although I had started on Luis's design, I was still very interested in PA0HRK's design, as was another ham with whom I correspond. Both of us were disappointed that only part of the schematic was published. Only a block diagram is shown of the vital AD8307 circuitry and the following dc amplifier. Now I am enough of a technician that I could find the AD8307 application note and eventually figure out the circuit. But why go though all that trouble when the design is completed and tested? Surely, there was enough room in *QEX* to publish the complete schematics. N0ADL's article "Signal Sources" suffered from the same ailment. After reading the preview in the previous issue, I was waiting—and it seemed like a long time, too—for that article. Like many, I badly need a microwave signal generator. Instead of schematics, all we got was, for the most part, a bunch of block diagrams. Yes, I am very familiar with the Mini-Circuits' and other VCOs, and yes, with a lot of work, I could figure out the design. Nonetheless, wouldn't it be better to publish complete schematics?

Perhaps in the future it would be possible to fill in the holes and publish complete circuits? I would certainly like to see them.—*David Metz, WA0AUQ, 725 Climer, Muscatine, IA 52761;* **davemetz@mut1.muscanet.com**

Hi Dave,

Thanks for your note. Very often, we assume readers can follow through unclear areas and fill gaps in articles. We therefore insist that authors be easily reachable for questions and comments and that the sources given for additional information be accurate. Sometimes when circuits are copied from manufacturers' application notes, duplicating them in their entirety isn't appropriate. In general, we run the schematics that authors provide. Let me know if you have trouble getting what you need.—Doug Smith, KF6DX

RF (Sep/Oct 1999)

Dear Zack, W1VT,

I read your article in *QEX* on MF/HF 50:450 Ω toroidal transformers. I have built several for commercial applications and have found that using different wire sizes for each of the three windings improves the bandwidth: #24 AWG for the 50-Ω winding, next #26 AWG and #28 AWG for the 450-Ω winding. I tightly twisted all three wires before winding them. I used ferrite cores from Palomar Engineers (**Palomar@compuserve.com**), mix 61. I have also built a loop antenna using current-mode operation and two MRF901 transistors feeding a push-pull version of this type of transformer. This results in a no-tune antenna with sharp nulls that receives signals from 100 kHz to over 30 MHz. The loop is two turns of #14 AWG wire in a 15-inch diameter, spaced about 1 inch apart. The prototype runs on a 9-V battery.—*Art McBride, KC6UQH, 705 Bozanich Cir, Vista, CA 92084;* **kc6uqh@k-online.com**

Temperature Limiting the 100-W MOSFET Amplifier (Nov/Dec 1999)

◊ The amplifier I described in the Nov/Dec 1999 *QEX* uses some fairly expensive MRF150 transistors. As a follow-up to the project, I decided that it would be prudent to add a circuit that protects them from excessive junction temperature, in case the fan fails or the dissipation increases too much for some reason.

A "precision thermistor," RadioShack 271-110, is rated at 10 kΩ ±1% at 25°C. I measured its resistance in boiling water to verify (at 100°C) the calibration chart that comes with the thermistor. At 90 to 95°C, its resistance is about 1230 Ω. I attached the thermistor to the ceramic button of one of the FETs, using a small drop of epoxy to hold it in place as shown in Fig A. The ceramic button is the hottest and the fastest-responding location, so I decided to control it rather than the flange or heat sink. This helps assure early detection of an FET temperature problem. The MRF150 has a junction-to-case thermal resistance, P_{JC}, of 0.6°C/W. If the "case" is the ceramic button, at a temperature of 95°C and a worst-case dissipation of 110 W per FET, the junction temperature reaches 160°C, which is 40°C below the 200°C maximum allowed. The maximum-allowed dissipation per FET is 300 W, derated to 180 W at a 95°C case temperature, which provides a 70-W safety margin at that temperature.

Fig B shows the circuit that controls temperature. The 4.7-V Zener, the 8.45 kΩ, 1% metal-film resistor and the thermistor form a voltage divider. When the thermistor resistance falls to 1230 Ω, the V_{BE} of Q1 falls slightly below 0.6 V, Q1 comes out of saturation and Q2 quickly goes into saturation. The condition for this is the voltage divider equation:

$$0.6 = 4.7\left(\frac{R_{TH}}{R_{TH} + R_B}\right) \qquad \text{(Eq 1)}$$

From this, a value of R_B, in my case 8450 Ω (a standard value), is correct for an R_{TH} of 1230 Ω. This causes the FET gate bias to fall quickly to a low value, which shuts off the FETs. They remain off until the case temperature falls below 95°C. While the FETs are off, Q3 lights the red LED, which can be mounted on the front panel of the equipment. The thermistor is across V_{BE} of Q1, which is 0.6 V or less, and this minimizes thermistor self-heating. I modified the regulator board to add the circuitry; the modified board set is available from FAR Circuits (see Note 11 of the article). The new information has been added to the *QEX* download file, **http://www.arrl.org/files/*qex*/1199SABN.ZIP**.

In normal SSB/CW/Data operation at 100-W PEP output on all bands, the protection circuit is idle, as intended. It is also idle at 100 W when continuously keyed, except on 160 meters, where it toggles for short periods. On that band, the efficiency is a little less because of transformer T4, which is marginal for that band because of the type-43 ferrite; a 1-dB reduction of output level eliminated the toggling. If the fan quits or any other mishap occurs, the circuit takes over and keeps the case temperature at a safe 95°C; but if the red LED lights frequently, take prompt action. As mentioned in the article, the fan as shown in Fig 1 or on the front cover is more than adequate at the 100-W output power level, except as noted here. I do urge that the fan be mounted in a way that maximizes the air flow through the fins and that a high quality heat sink be used.—*William E. Sabin, W0IYH, 1400 Harold Dr. SE, Cedar Rapids, IA, 52403*; **sabinw@mwci.net**

Notes on "Ideal" Commutating Mixers (Nov/Dec 1999)

Doug,

I read your recent article on commutating mixers in the Nov/Dec 1999 *QEX*. However, your analyses are somewhat flawed in that they fail to account for the effect of terminating impedances seen at all the mixer ports. The effects of these impedances are to reflect harmonics back into the mixer in either aiding or opposing phase, so as to modify the mixer waveform. At microwave frequencies, these effects are far more significant than the LO waveform, which is likely to be a distorted sine at best. Pure square waves do not exist at a sufficiently high frequency, although the diodes might be current-driven to approximate such a characteristic, if desired.

Dr. A. A. Saleh wrote his thesis at MIT about the conditions required to minimize conversion losses in mixers, as well as optimal design of image-recovery or -rejection types. He came up with a figure of 4.78 dB for an ideal DBM that was terminated correctly—that is with the odd harmonics open-circuited and the even harmonics shorted, as I recall. He also came up with several designations for the class of terminations—some of these are still in use today. His book was published by the MIT press. I urge you to look at it. The text is quite readable even with the math. Failing to locate a copy, I believe application notes by Watkins-Johnson or material in *The Microwave Engineer's Handbook* will provide insight. 73 and good work on *QEX*.—*David M. Upton, WB1CMG, 25 Harwood Rd, RR1, Mount Vernon, NH, 03057;* **david@wb1cmg.mv.com**

Hi Doug,

Way back in 1973, an IEEE conference (or maybe IERE) had a paper by a guy from Plessey about diode commutating mixers. He showed that it was possible to reduce the theoretical loss to 3.14 dB with perfect diodes by reflecting the image back into the mixer, remixing it. Some work I did at the time showed this to have some merit—I could get down to about 4.5 dB conversion loss in a conventional mixer, but the IMD performance went to you-know-where in a you-know-what—and then some!

It remains to my mind an interesting academic thing, possibly useful at microwaves. Of course, I didn't look into analog multiplier operation. That is more usual with transistor trees or Gilbert-cell mixers. There are two basic modes for Gilbert-cell mixers. One is the analog multiplier mode, where the top transistors are operated in a linear manner. This gives, generally, more gain and worse IMD, while the conversion gain is dependent on the LO drive level. In the more usual mode, the top transistors commutate the signal. This gives relative insensitivity to LO level, and the IMD is mostly determined by the voltage-to-current transfer characteristic of the bottom transistors. The current in the top transistors can be very low without affecting the IMD—See Ref 1.

Improved IMD and signal handling can be achieved by increasing the

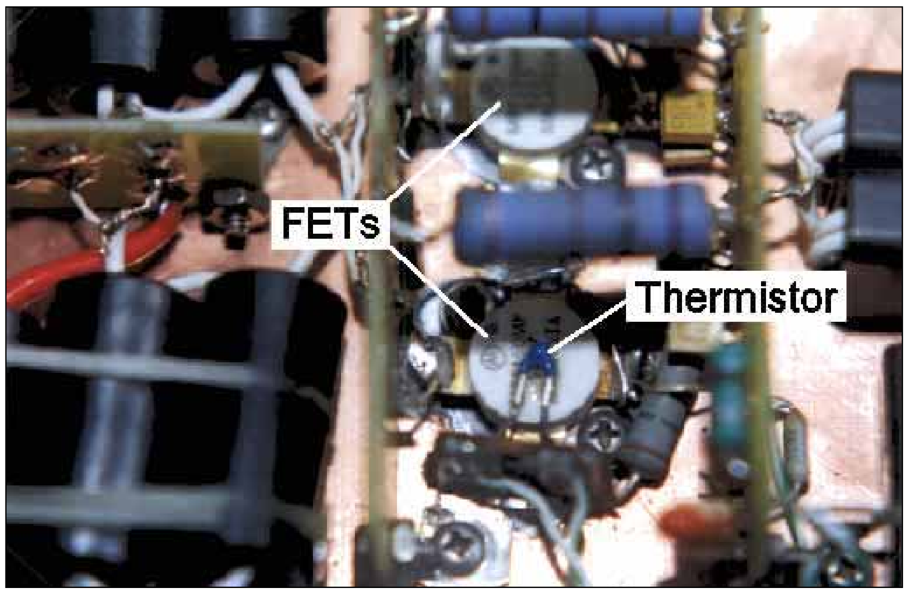

Fig A—The thermistor is bonded to one of two PA FETs.

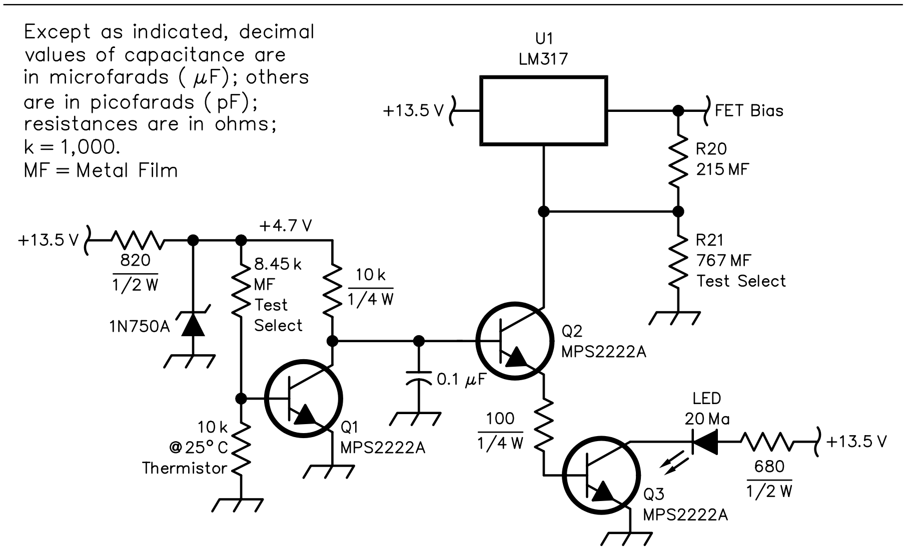

Fig B—A schematic of the temperature-protection circuit.

current in the lower transistors—see Ref 2. That reference mentions a method of applying negative feedback using PNP transistors, which wasn't practicable with the IC processes available then. I have had some interesting results with transformer noiseless feedback in a semi-discrete mixer.

I believe that the relative insensitivity to termination impedance of the tree or Gilbert cell happens because the transistors are effectively current sources, and the performance is thus relatively unaffected by the load. Indeed, for maximum signal handling, you need to short the unwanted sideband, as this reduces the voltage swing by 6 dB. In theory, you can then have 6 dB more signal before you hit the same amount of gain compression.

This suggests that an interesting possibility might exist by using discrete transistors at the bottom, and a matched integrated pair for the switching, with transformer feedback to linearize the bottom pair.—*Peter Chadwick, G3RZP, Sen. MIEEE, Three Oaks, Braydon Swindon, Wilts SN5 0AD, UK;* **Peter_Chadwick@mitel.com**

References

1. H. C. Nauta and E. N. Nordholt, *Proceedings of the 4th International Conference on Radio Receivers and Associated Systems*, (Bangor, Wales: IERE, 1986; ISBN 0 903748 66 5) "The Intermodulation-Free Dynamic Range of Bipolar Switching Mixers," p 141 et seq.
2. P. E. Chadwick, *WESCON Proceedings 1981*; Session 24, Mixers for High Performance Radio, "The SL6440 High-Performance Mixer."

On the Melding of the Magazines

Hi Doug,

I just received the letter about the merger with *Communications Quarterly*. Seems like a good idea if it keeps a good magazine in circulation. I find lots of stuff in both that I do not find in my professional literature. Please do what ever is needed and keep up the good work! Best regards.—*Richard Kiefer, K0DK, 4700 47th St, Boulder, CO, 80301;* **kiefer@csd.net**

Doug,

Congratulations on your new expanded *QEX*. As a subscriber to *Communications Quarterly*, I look forward to your new publication. As I recall, I became a subscriber to *Communications Quarterly* because of *ham radio* magazine's demise. From the beginning, I was not entirely satisfied because it didn't have the hands-on projects I enjoyed in *ham radio*. In any case, I look forward to supporting this new endeavor. Thanks.—*Michael J. Black, Sr., N6EGN, PO Box 211, McCloud, CA 96057;* **n6egn@snowcrest.net**

Doug,

I received your message about purchasing *Communications Quarterly*. I have subscribed for a long time and like it; I am glad to see you purchase it. I was sorry to see *ham radio* quit. I also like *Nuts and Volts*. Keep the same high-quality items that we hams can relate to: those that are not too easy, but need some effort. Thank you for including *QEX*. I will be glad to see what is there.—*Herman W Bansemer*, W6NDT, 14201 Rattlesnake, Grass Valley CA 95945; **w6ndt.jps.net**

Doug,

I was a charter *Communications Quarterly* subscriber. You can be sure, if the quality of *QEX* is maintained, I won't drop the subscription. You have come a long way and done some very hard work. I appreciate it a lot.—*Douglas Datwyler, WR7O, 1506 Plata Wy, Sandy, UT 84093-2348;* **douglas.datwyler@ieee.org**

Doug,

I have just finished the Nov/Dec 1999 issue of *QEX*, and I must say that it is a remarkable document. Clearly your efforts come through and I feel compelled to make this notation to you. I have been receiving *QEX* since its very first issue; the strides that you have made have been very large. Thanks.—*Dick Jansson, WD4FAB, ARRL TA, 1130 Willowbrook Tr, Maitland, FL, 32751;* **wd4fab@cfl.rr.com** □□

Next Issue in QEX/Communications Quarterly

In our 200th issue, learn about a mode that may catch on: digital voice over HF. Charles Brain, G4GUO, and Andy Talbot, G4JNT, explain how to do it and what to expect. Beware, though: US rules do not currently permit their 1800-baud, multi-carrier modulation format below 30 MHz. Requests for Special Temporary Authorities (STAs) should be coordinated through the ARRL to avoid duplication of effort. Contact ARRL Counsel Chris Imlay, W3KD (**w3kd@arrl.org**) if you are interested in obtaining an STA for this mode.

Johan Forrer, KC7WW, presents a tool that should be quite handy during planning and initial tests of digital modes on HF: a channel simulator. Johan explains standard models for ionospheric propagation, then goes into details of their implementation in the simulator. He also provides test results for three communication modes that show various levels of error-rate degradation under worst-case conditions.

ARRL TA L. B. Cebik begins a series on log-periodic dipole arrays (LPDAs). He gives the subject thorough treatment. These antennas may be quite desirable where high gain is needed over more than an octave and the acreage is available.

Editor Doug Smith, KF6DX, has researched some of the techniques used to compress digitized audio signals. He is applying variants of them to bandwidth compression of a mode that's been around for a while: analog voice over radio. He explains the basis of his DSP methods in the first part of a two-part article about "perceptual-transform coding." □□

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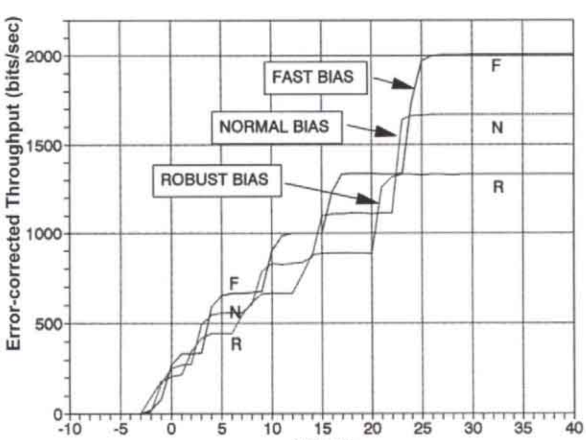

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MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25M	20	25	3½ x 19 x 9⅜	6.5
SRM-30M	25	30	3½ x 19 x 9⅝	7.0


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MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25M-2	20	25	3½ x 19 x 9⅜	10.5
SRM-30M-2	25	30	3½ x 19 x 9⅝	11.0


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*ICS – Intermittent Communication Service