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.

153 Gretna Green Way
Los Angeles, CA 90049
kd6ozh@amsat.org

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

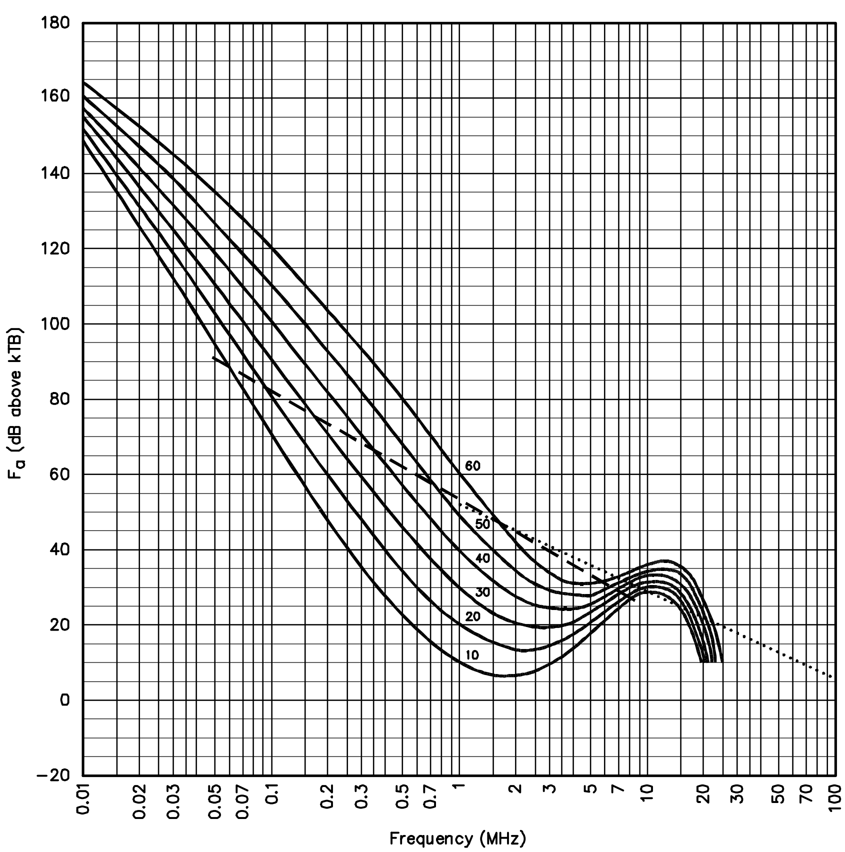

Fig 1—Atmospheric noise levels in a quiet location. Solid lines are the expected levels of atmospheric noise (in dBN_0) with curve 10 showing the quietest locations on Earth. The dashed line is the expected level of galactic noise and the dotted line is the expected level of manmade noise. The peak at 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

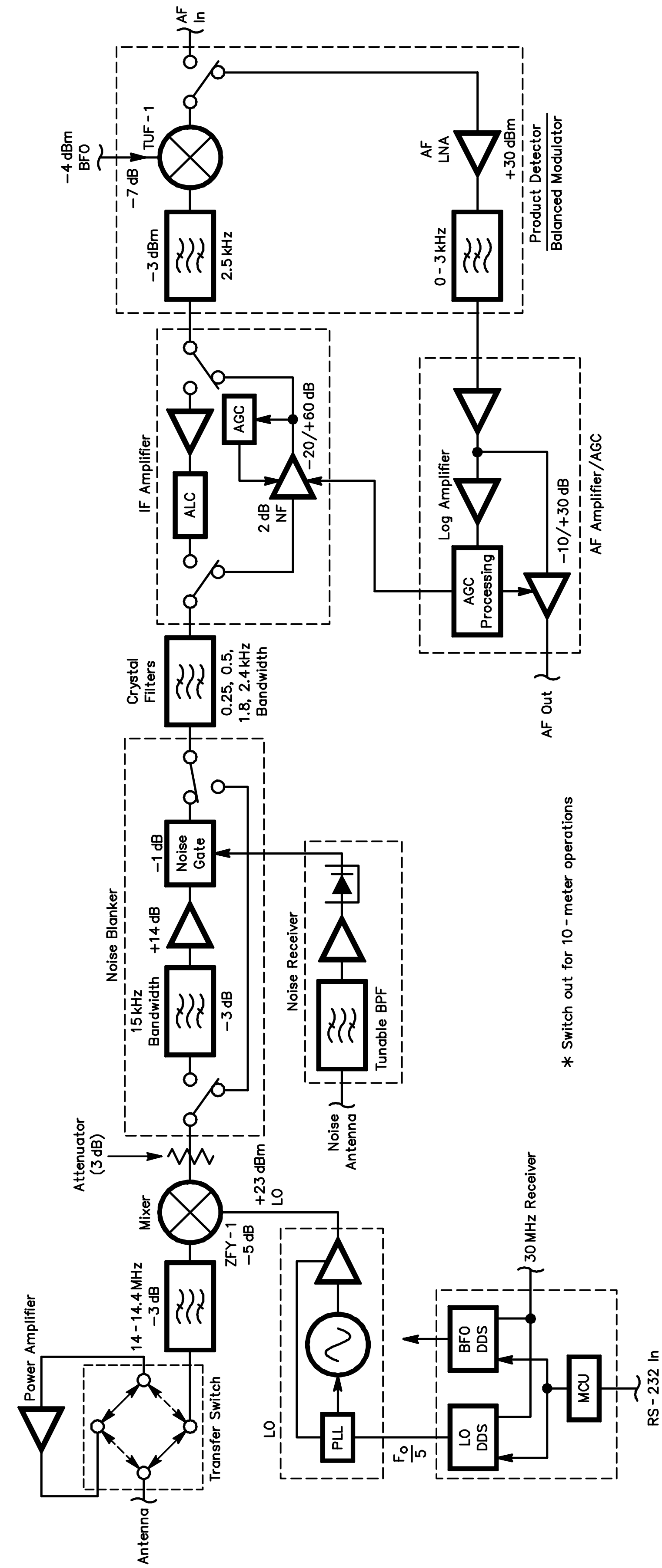

Fig 2—Transceiver block diagram.

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-

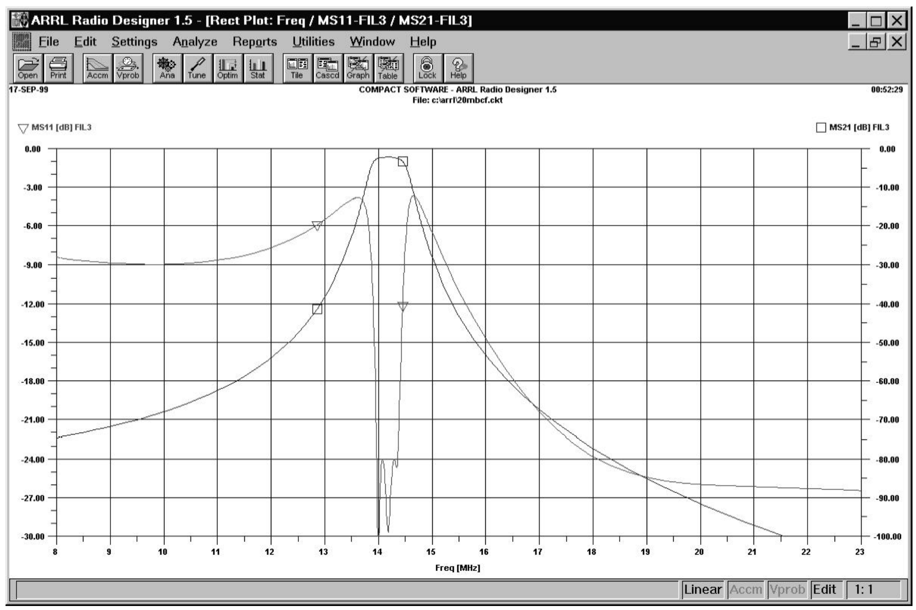

Fig 3—20-meter bandpass filter characteristics.

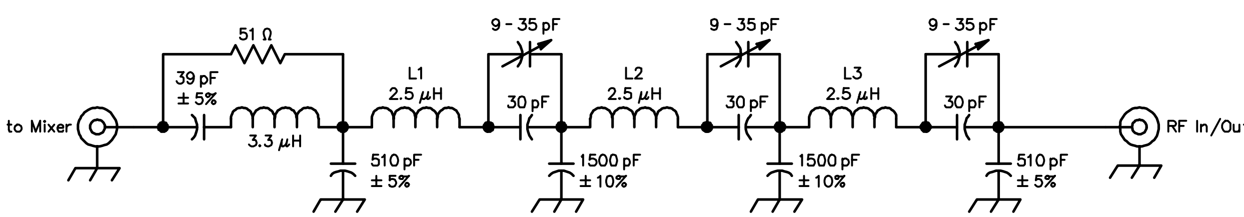

Fig 4—20-meter bandpass filter schematic diagram.
L1-L3—2.5 µH, 24 turns #22 AWG on T50-6 powdered iron core

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

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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
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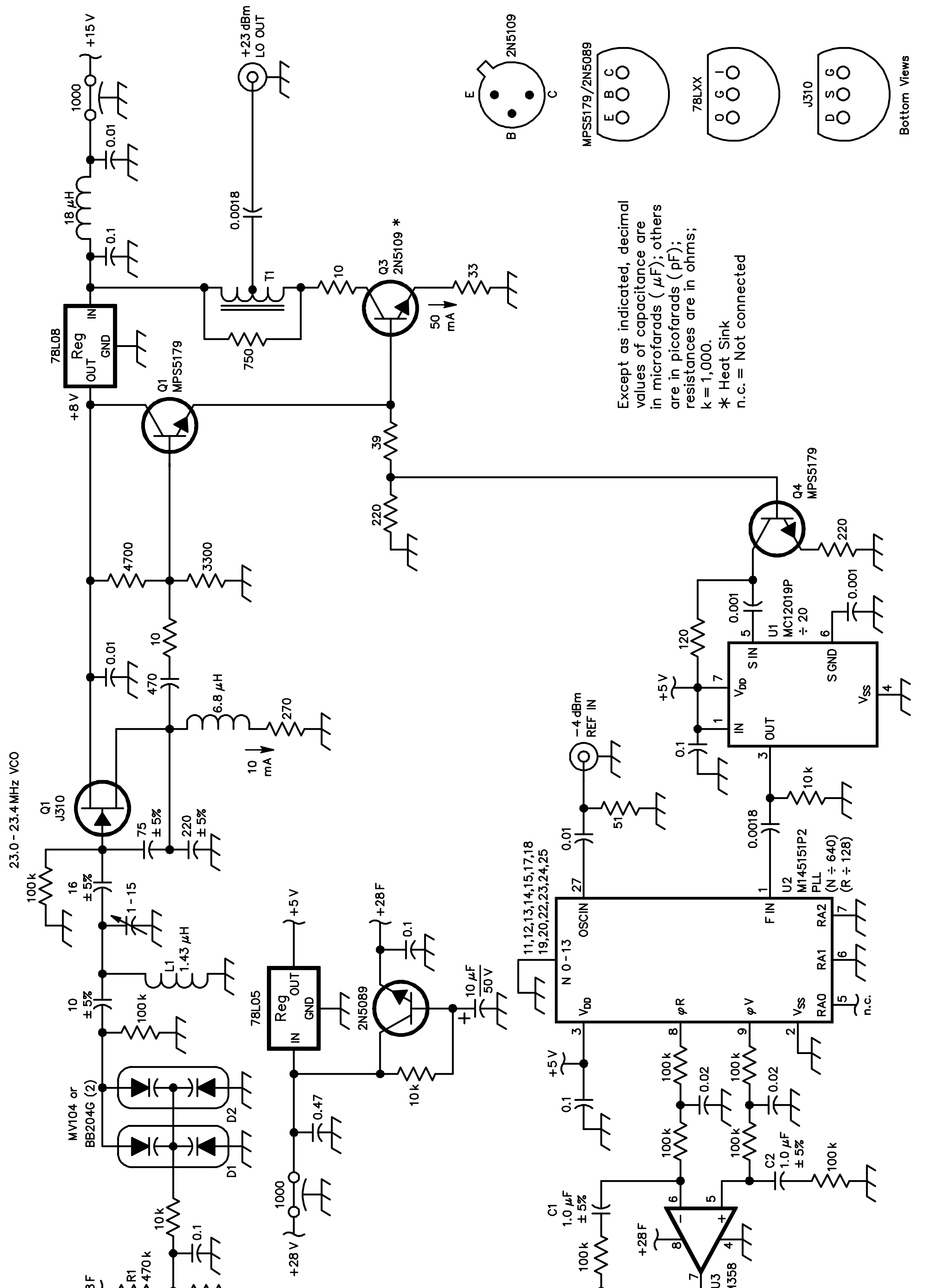
Fig 5—VCO simulation.



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
+15 V
+23 dBm LO OUT
2N5109
MPS5179/2N5089
78LXX
J310
Bottom Views
78L08
Q1 MPS5179
Q3 2N5109 ∗
Q4 MPS5179
U1 MC12019P ÷ 20
U2 M145151P2 PLL (N ÷ 640) (R ÷ 128)
U3 LM358
23.0 - 23.4 MHz VCO
Q1 J310
MV104 or BB204G (2)
78L05
2N5089
−4 dBm REF IN
+28 V
+28 F
+5 V
+8 V

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

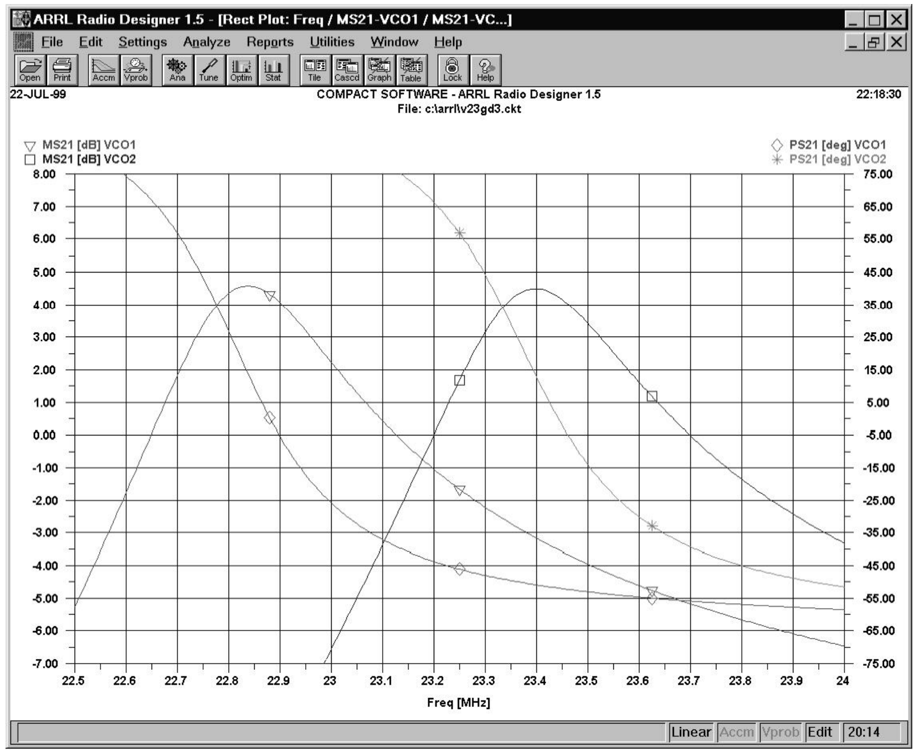


Fig 7—LO gain and phase plot.

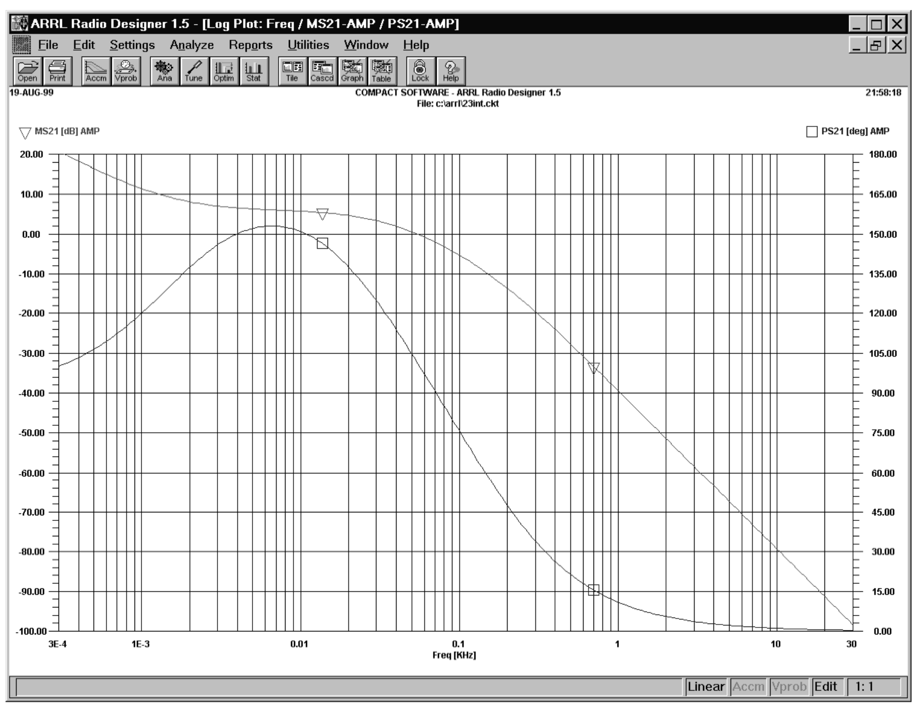


Fig 8—PLL filter amplitude and phase response.

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,

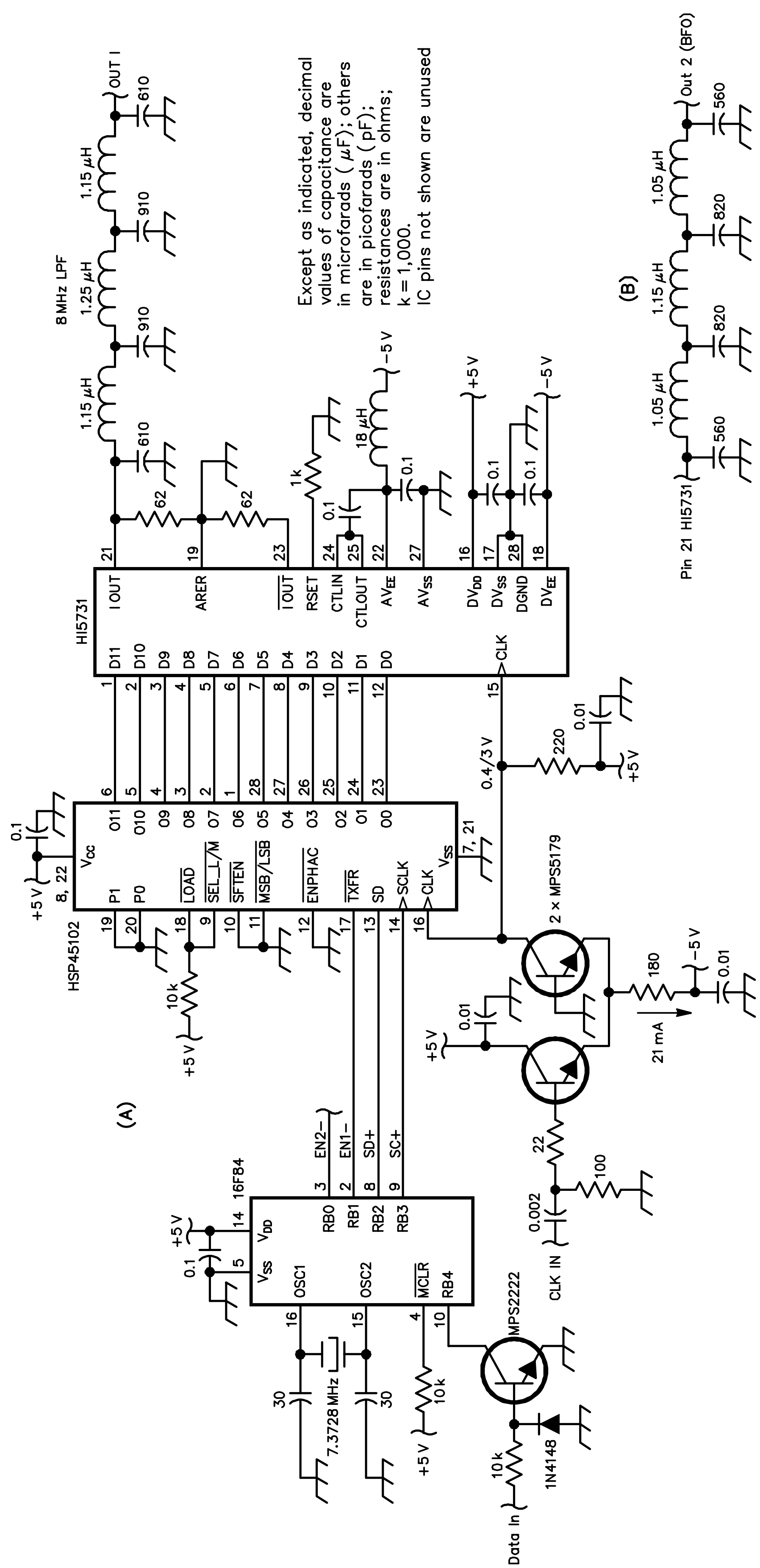


Fig 9 (right)—(A) PIC and LO DDS schematic diagram. The low/high clock levels to pin 15 of the HI5731 should be 0.4/3-V. (B) BFO low-pass filter schematic diagram.

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:

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

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.

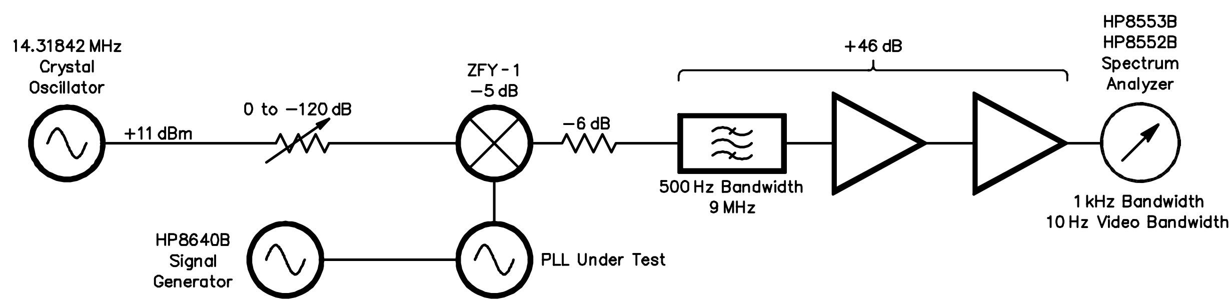


Fig 11—LO phase-noise test setup.

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

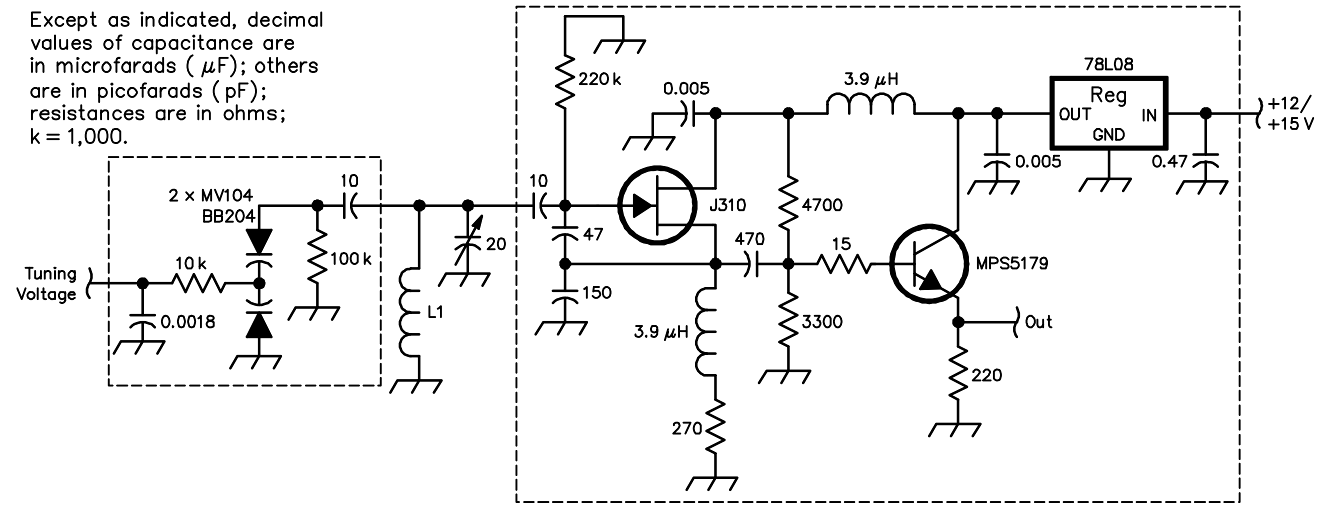


Fig 12—10-meter VCO schematic diagram.

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.

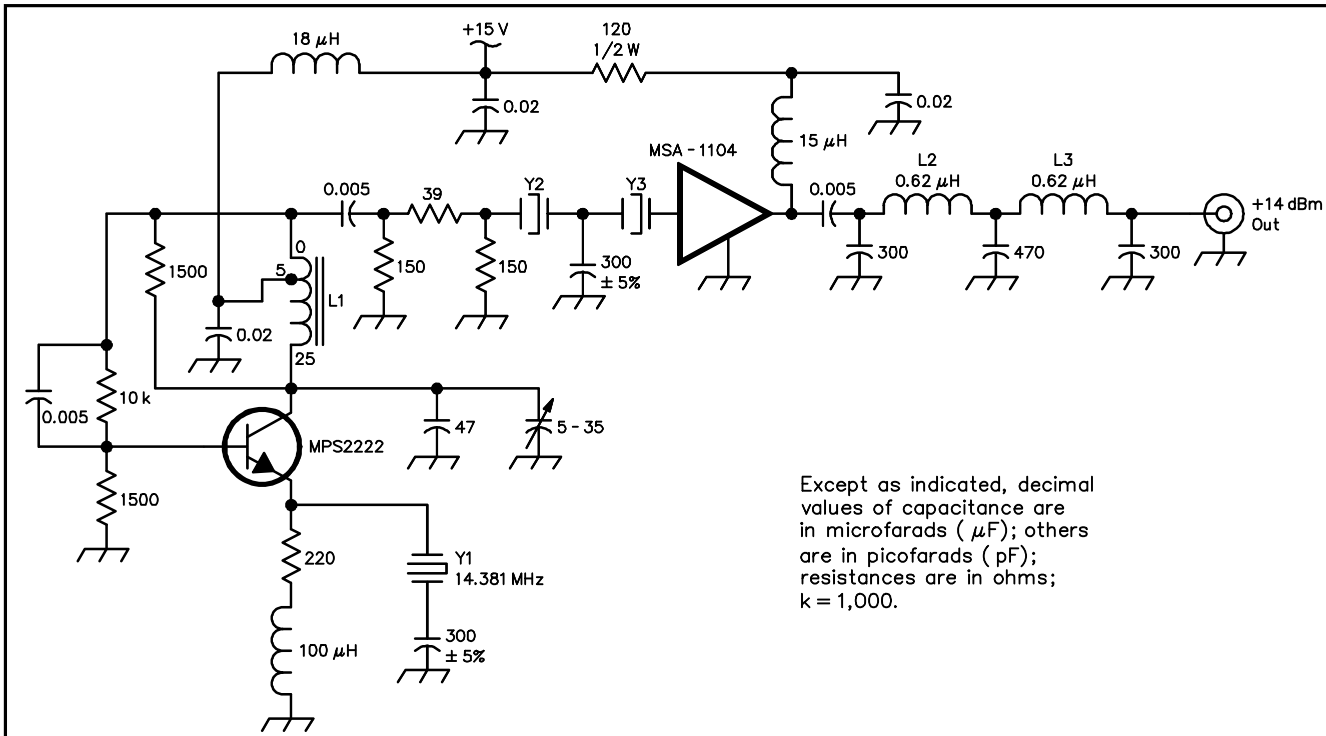


Fig A—Low-noise crystal oscillator schematic diagram.

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

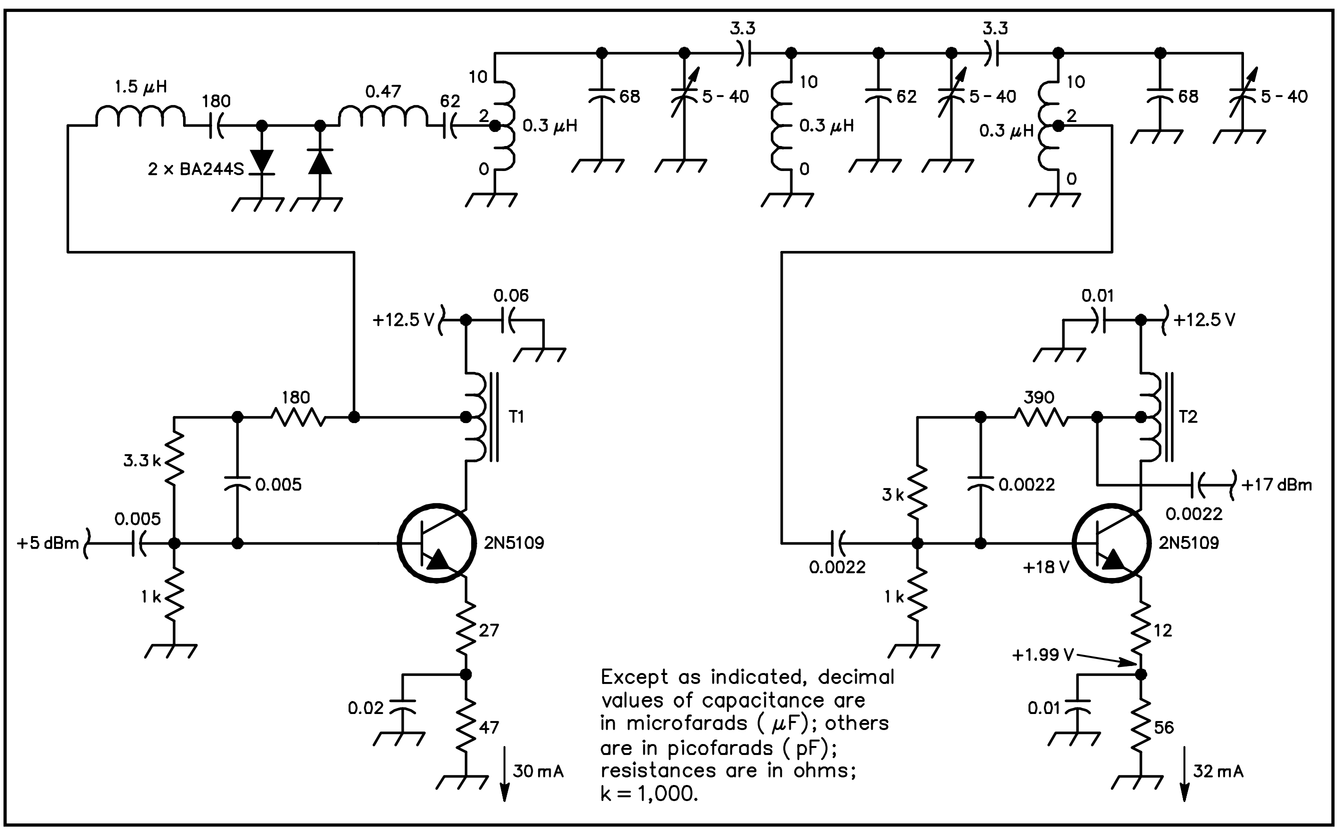


Fig B—10 to 30 MHZ tripler schematic.

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:

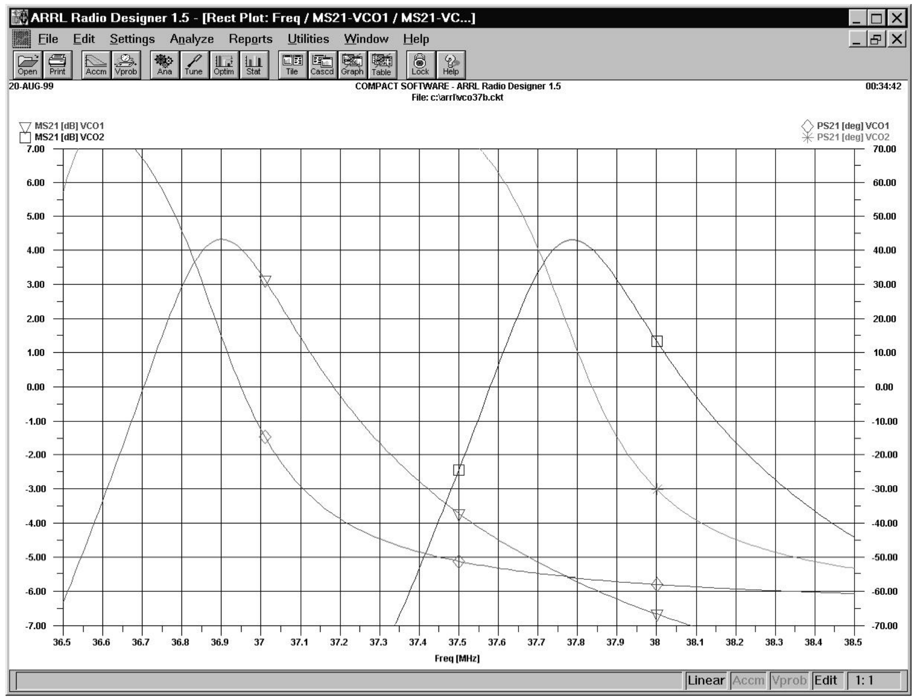


Fig 14—Predicted 10-meter VCO gain and phase.

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

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:

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

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 1 1/4 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 2 1/4×2 1/4×5-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 1 1/2 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.

[9]D. DeMaw and Wes Hayward, "Modern Receivers and Transceivers: What Ails Them?," *QST*, Jan 1983.

[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.

[13]M. Mandelkern, "A High-Performance Homebrew Transceiver: Part 1," *QEX*, Mar/Apr 1999.

[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 mes aging 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.

The ATR-2000: A Homemade, High-Performance HF Transceiver, Pt 2

Part 2 describes the IF and audio sections, including IF amplifier, product detector/balanced modulator, RF compressor, AGC and PC-interface circuits.

By John B. Stephensen, KD6OZH

This article series describes my homebrew HF transceiver. Part 1 described the general architecture and the front end, including the synthesized local oscillator and BFO, mixer and RF band-pass filter.[1] This part describes the IF and audio sections of the transceiver, including IF amplifier, product detector/balanced modulator, RF compressor, AGC and PC-interface circuits.

Dynamic-Range Considerations

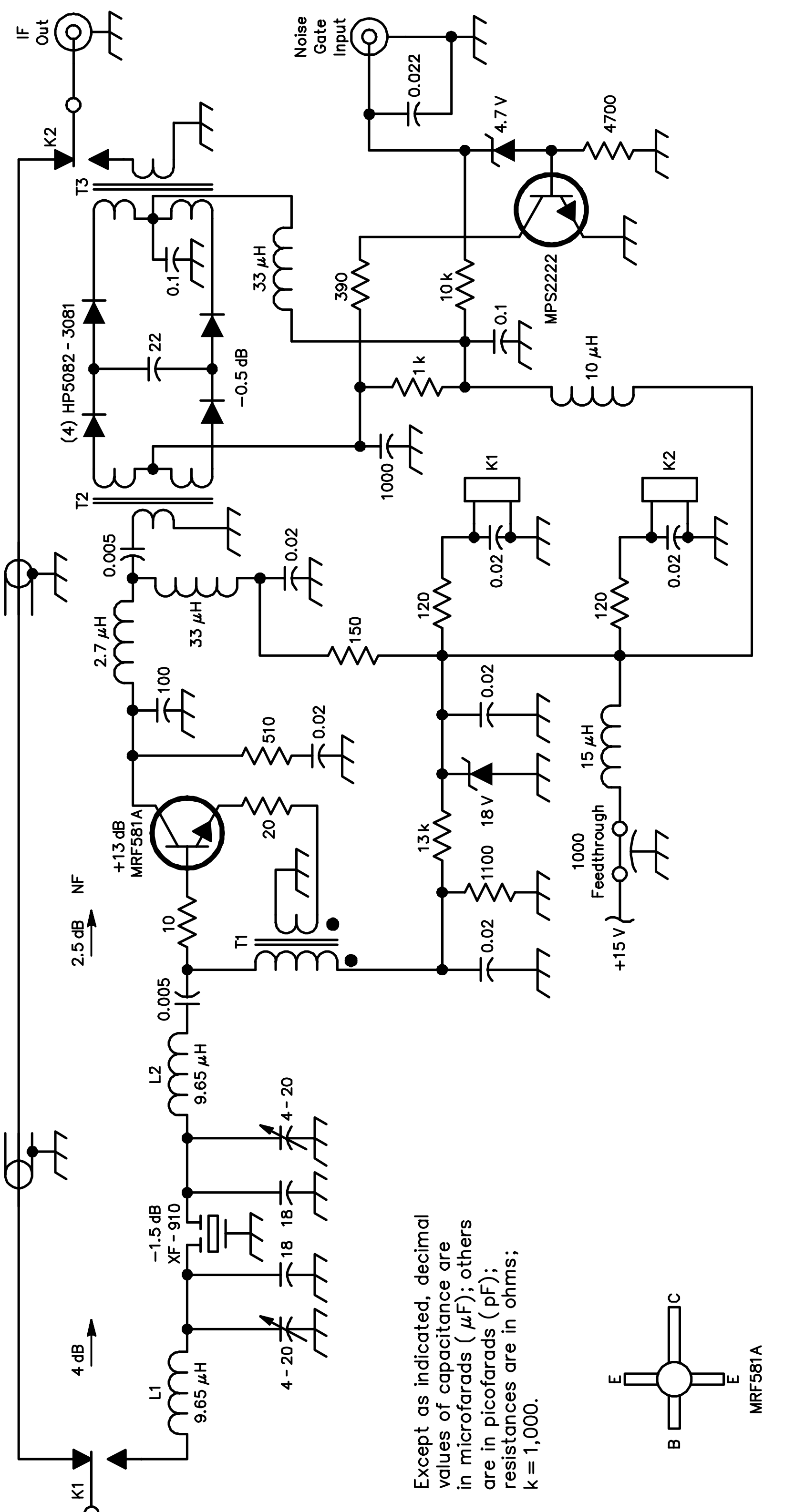
Gain distribution is an important facet of transceiver design.[2] Gain ahead of the main band-pass filter should be minimized, so signals on adjacent frequencies do not cause serious intermodulation distortion (IMD). With the IF Preamplifier/Noise Gate module switched out, the ATR-2000 has no gain ahead of the main crystal filter, but rather a loss of 12 dB. As described in Part 1 of this article, noise figure is 18-22 dB depending on which filter is selected, and the 1-dB compression point is +23 dBm. The mixer is not terminated in an ideal load, so we expect at least 3 dB of degradation for a +35 dBm third-order intercept point (IP3).

When the IF Preamplifier/Noise Gate module is switched in, it adds 11 dB of gain, bringing total gain ahead of the main crystal filter up to –1 dB. The noise figure of this module is 4 dB, bringing the system noise figure down to 16 dB. The 1-dB compression point is reduced because of the additional gain. It is determined primarily by the compression point of the IF preamplifier itself, which is +13 dBm at the output to the crystal filter. This is intentionally limited to prevent destruction of the crystals by strong signals. The receiver's input compression point is therefore +14 dBm.

When the incoming signal is within the main filter bandwidth, the compression point is determined by the first few stages of the IF amplifier when it is running at minimum gain. This is the condition when a strong signal is present and AGC is on. This transceiver uses the Analog Devices AD603 low-noise, 90-MHz variable-gain amplifier. Its advantages include

[1]Notes appear on page 51.

153 Gretna Green Wy
Los Angeles, CA 90049
kd6ozh@amsat.org



a +13 dBm 1-dB input compression point and a gain reduction mechanism that does not reduce input signal-handling capability. A post-filter amplifier is included to maintain noise figure; it reduces the 1-dB input compression point to –1 dBm. A 4-dB loss in the SSB crystal filter and 12-dB loss in the front end result in a +15 dBm 1-dB compression point at the antenna terminals without the IF preamplifier, and +4 dBm with it. The resulting dynamic ranges for various input frequencies are shown in Table 1.

IF Preamplifier/ Noise Gate Module

Fig 1 shows the IF preamplifier and noise-gate module. It consists of a diplexer, band-pass filter, low-noise amplifier and noise gate. The diplexer provides a termination for the mixer at the LO and image frequencies. The Q is low and insertion loss is 0.4 dB.

When not in use, power is removed and K1 and K2 bypass all stages after the diplexer. Mechanical relays are used to ensure low IMD as this circuit precedes the main filter. The Omron relays are designed for RF use; they have more than 60 dB of isolation between contacts. At least 80 dB of isolation between input and output is required when the noise blanker is operating.

When the preamplifier is in operation, a two-pole monolithic crystal filter with a 15 kHz bandwidth filters incoming signals and noise. L1, L2 and associated capacitors provide transformation to and from 50 Ω. Because impedance of the KVG XF-910 monolithic crystal filter is very high (6000 Ω), two-thirds of the 1.5 dB loss in this circuit actually comes from the matching networks. High-Q toroidial coils are used to minimize the loss. Note that the filter bandwidth must be several times the information bandwidth (to ensure minimum spreading of pulses), but narrow enough to delay the arrival of noise pulses until after blanking is in effect.

The amplifier consists of a Motorola MRF581 configured as a low-noise

Fig 1—IF preamplifier and noise gate. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors.
K1, K2—Omron G6Y-1 relay.
L1, L2—43 turns #26 AWG on a T50-2 powdered-iron toroid core.
T1—3 t #26 AWG primary, 1 t #24 AWG secondary on a BLN-43-2402 binocular ferrite core.
T2, T3—6 t #24 AWG (trifilar) on an FT37-43 ferrite toroid core.

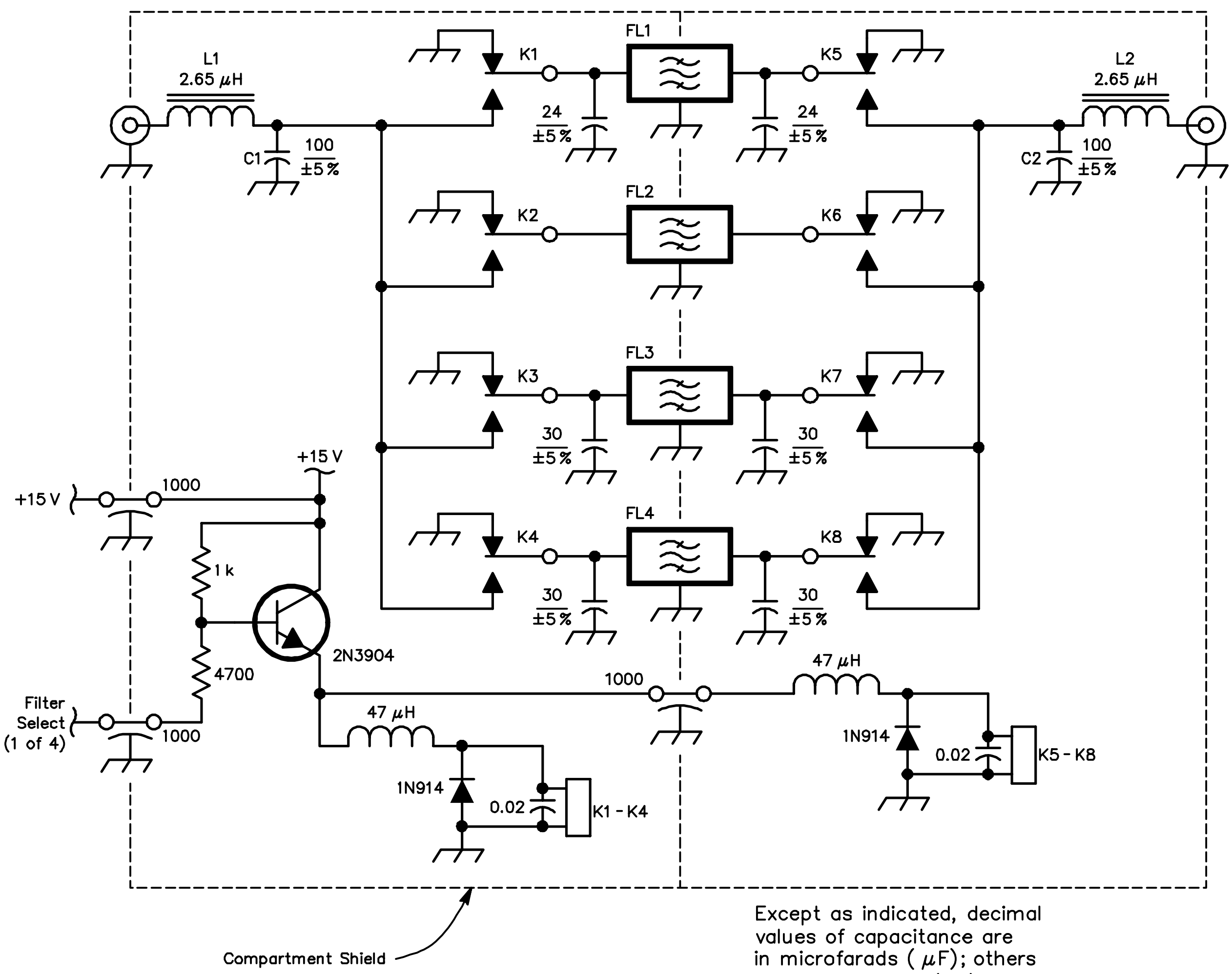
amplifier with a 50-Ω input impedance and 13 dB of gain. Unlike many other "noiseless feedback" circuits, this particular circuit has the advantage of providing high isolation between input and output. The MRF581 is biased at 20 mA to provide a low noise figure; the resulting +13-dBm compression level provides some protection for the following crystal filters. The 510-Ω resistor in the collector circuit sets the output impedance of the amplifier at 50 Ω to provide a proper termination for the main crystal filters.

Following the amplifier is the noise gate. A balanced circuit is used to minimize switching noise at the output. HP 5082-3081 PIN diodes are used to minimize IMD and transient generation during turn-on and turn-off. Four PIN diodes are used in the circuit to achieve a high attenuation when the gate is off. The 22-pF capacitor resonates broadly with the inductance of the two transformers at the IF. When the gate is off, it provides shunt reactance to form an "H" attenuator with the junction capacitance of the PIN diodes. This achieves 78 dB of attenuation.

The circuit at Q2 is a time-delay circuit. Normally, the 0.022-µF capacitor is charged through the 10-kΩ resistor, causing the 4.7-V Zener diode and the MPS2222 transistor to con-

Table 1—Predicted Receiver Dynamic Range with 2.4 kHz bandwidth

Frequency Offset	*Total Noise Figure*	*1-dB Compression*	*Receiver IP3*	*Dynamic Range*
IF Preamplifier/Noise Gate out				
> 1.5 kHz	18 dB	+23 dBm	+35 dBm	104.7 dB
0-1.5 kHz	18 dB	+15 dBm	—	—
IF Preamplifier/Noise Gate in				
> 10.0 kHz	16 dB	+23 dBm	+35 dBm	106.0 dB
1.5-10.0 kHz	16 dB	+14 dBm	+29 dBm	102.0 dB
0-1.5 kHz	16 dB	+4 dBm	—	—



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.

Fig 2—Crystal-filter selection and matching circuits. Only one relay control circuit is shown; the others are identical. Unless otherwise specified, use ¹/₄ W, 5%-tolerance carbon composition or film resistors.

FL1—KVG XF-9810, 2400 Hz bandwidth.
FL2—International Radio 2308, 1800 Hz bandwidth.
FL3—KVG XF-9NB, 500 Hz bandwidth.
FL4—KVG XF-9P, 250 Hz bandwidth.
K1-K8—RS-241 SPDT relay, 12-V coil.
L1, L2—19 t #24 AWG on a T50-6 powdered-iron toroid core.

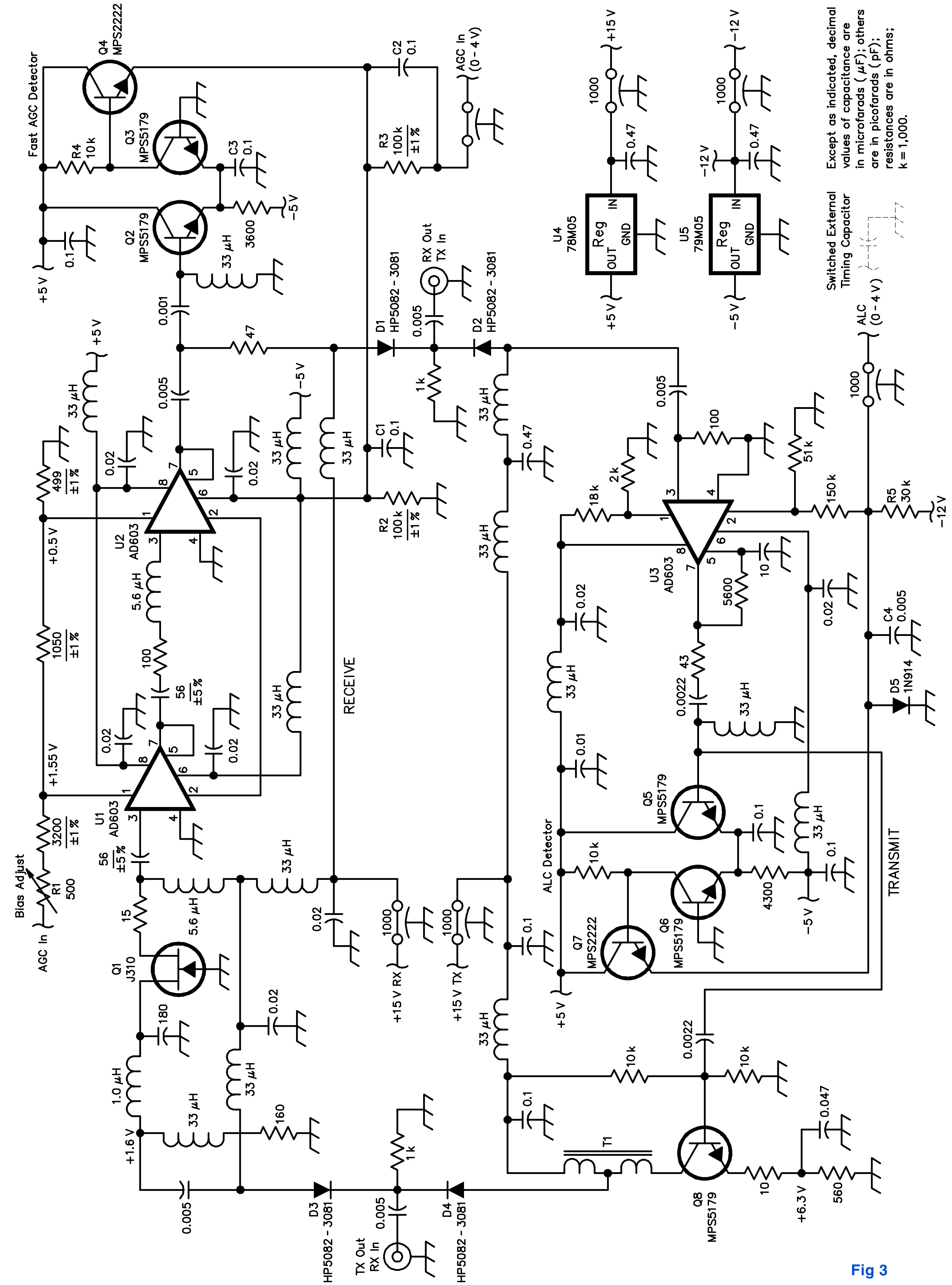

Fast AGC Detector
Q4 MPS2222
Q3 MPS5179
Q2 MPS5179
R4 10 k
C3 0.1
C2 0.1
3600
−5 V
+5 V
0.1
33 μH
0.001
AGC In (0 - 4 V)
R3 100 k ±1%
D1 HP5082 - 3081
D2 HP5082 - 3081
RX Out TX In
0.005
47
1k
C1 0.1
U4 78M05
Reg
OUT
GND
IN
+15 V
1000
0.47
U5 79M05
−12 V
−5 V
Switched External Timing Capacitor
ALC (0 - 4 V)
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
499 ±1%
0.02
U2 AD603
+0.5 V
5.6 μH
100
R2 100 k ±1%
RECEIVE
1050 ±1%
56 ±5%
+1.55 V
U1 AD603
3200 ±1%
R1 500
Bias Adjust
AGC In
Q1 J310
15
180
1.0 μH
+1.6 V
160
0.005
D3 HP5082 - 3081
TX Out RX In
D4 HP5082 - 3081
+15 V RX
+15 V TX
18 k
2 k
U3 AD603
5600
10
51 k
150 k
R5 30 k
−12 V
C4 0.005
D5 1N914
43
0.0022
0.01
ALC Detector
Q5 MPS5179
Q6 MPS5179
Q7 MPS2222
4300
TRANSMIT
10 k
0.0022
T1
Q8 MPS5179
0.047
560
+6.3 V

Fig 3

duct, turning on the noise gate. When a noise pulse is detected, an external transistor discharges the capacitor. This turns off Q2 and the noise gate. The time required to recharge the capacitor ensures that the noise pulse (stretched by the 15 kHz filter) has ended before the noise gate is re-enabled.

Crystal Filters

The main crystal filter immediately follows the IF preamplifier. This receiver has four selectable filters with 250, 500, 1800 and 2400 Hz bandwidths for PSK, FSK and SSB. (See Fig 2.) The filters' 500-Ω input and output impedances are matched by two L-networks: C1/L1 and C2/L2. Since the Q is only three, these are fixed-tuned using 5%-tolerance components. Additional capacitors are placed near each filter to provide the proper termination. Some filters require a pure 500-Ω source and load impedance while others require 15 or 30 pF of parallel capacitance.[3]

The filters are selected by mechanical relays with unused filter inputs and outputs grounded to minimize feed-through. Power for the relays is filtered to ensure that no coupling occurs via power leads or capacitance to the solenoid coil. The filter-selection relays are controlled from the PC-Interface module via four PNP transistors Q1-Q4. This allows one end of the relay coils to be grounded in order to minimize cross coupling via the power supply.

To preserve the stopband attenuation characteristics of these filters, good mechanical design is required. Unwanted coupling between the filter input and output must be minimized. The filters are mounted directly to the chassis with the input and output pins separated by a shield that partitions the chassis into two compartments. I found that small aluminum boxes didn't work here—they only gave 70 dB of isolation. An old-fashioned 4×5×2-inch welded chassis with a bottom plate gave 90 dB of isolation, which doesn't compromise the 80-dB stopband attenuation of the narrow SSB filter.

Fig 3—There are separate IF amplifiers for receive (upper circuitry) and transmit (lower circuitry) paths. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. All capacitors are ±20% and RF chokes are ±10% tolerance unless labeled otherwise.
T1—8 t #28 AWG bifilar wound on an FT23-43 ferrite toroid core.

IF Amplifier

The IF amplifier is shown in Fig 3. Independent circuits are used for transmit and receive, with the signal flow controlled by PIN-diode switches (D1-D4) for low distortion. The first amplifier in the receiver IF strip and the last amplifier in the transmitter IF strip are switched off when not in use by removal of the power-supply voltage.

The receive path consists of a low-noise amplifier followed by two variable-gain amplifiers and an AGC detector. Q1 is a J310 FET in a common-gate circuit with low-Q L networks for input and output matching. The circuit has a gain of 12 dB and brings the IF amplifier noise figure to 3 dB. A 3-MHz-wide series-tuned band-pass filter between the two variable-gain amplifiers limits the broadband noise at the AGC detector.

U1 and U2 are Analog Devices AD603 amplifiers having logarithmic gain control. They are capable of handling 3 V (pk-pk) input signal levels. Gain is variable over a 40-dB range with 1 V of control signal variation. The AGC voltage is applied differentially between pins 1 and 2 of each device; over 80-dB of total gain variation is achieved. The network of 1% resistors connected to pin 1 of each device biases them so that gain is controlled sequentially. R1 is adjusted to provide the correct bias voltages as shown in the schematic and compensates for any variation in the voltage from U4. As the AGC voltage increases, the gain of U2 is reduced from 30 dB to –10 dB before any reduction in the gain of U1. This maximizes the signal-to-noise ratio of the IF strip.

Receive IF gain is controlled from two sources. The control voltage for the slow AGC loop is applied to the 2:1 voltage divider formed by R2, R3, C1 and C2, resulting in a 0-to-4-V AGC range. Note that there is no long time-constant associated with driving the divider since the capacitors balance each other.

Q2 and Q3 form an envelope detector for the fast AGC loop. Q2 acts as a rectifier. It is cut off when the base voltage falls below the sum of the emitter voltage and base-emitter barrier voltage. Q3 is a common-base amplifier that provides dc gain and compensates for temperature variations in Q2's emitter-base voltage in a manner similar to a differential amplifier. The compensation is not perfect because the currents through the two transistors are not identical. The actual variation in detector gain, however, was less than 1-dB from 16°C to 38°C (room temperature) and was all concen-trated at the low end, where there is little operational effect. The detector's rise and fall times are determined by the emitter resistance and C3 to be about 6 µs. The voltage developed across R4 is buffered by Q4, an emitter follower, then applied to the AGC pins of the IF amplifiers, U1 and U2. The advantages of this detector are that it responds to low-level signals and that the output is logarithmic—within 1 dB—over a range of input voltages from 70 mV (pk-pk) to 250 mV (pk-pk). This results in a reasonably constant AGC loop gain and good transient response.

Additional AGC filtering is provided by R2, R3, C1 and C2, amounting to 50 kΩ of resistance and 0.2 µF of capacitance as seen from the emitter follower. This provides a 6-µs attack and 10-ms exponential decay time for the fast AGC. Any voltage from the slow AGC loop that is more than twice the fast AGC detector's output will override the output of Q4 and take control of the IF amplifier gain.

The transmit path contains a variable-gain amplifier (U3) that drives the ALC detector (Q5, Q6 and Q7) and a fixed-gain amplifier (Q8). These circuits amplify the SSB signal from the product detector/balanced modulator module and compress it. The input to this module can range from –10 dBm to –58 dBm depending on the amount of audio applied to the balanced modulator. The output is leveled to a ±4 dBm range.

The ALC detector is identical to that used in the receiver's AGC detector. It generates 0 to 4 V of ALC with a logarithmic response. The ALC is applied to U3; it varies U3's gain from +43 to +3 dB as signal levels rise. This compresses the transmitted signal by a factor of six or more, translating the original 48-dB range to 8-dB. Most voices only vary by 20 dB during normal speech. This range is compressed to 3 dB and the rest of the control range compensates for variations in the dis-tance from the operator to the micro-phone.

The ALC attack time is approximately 6 µs, but the release time is variable. The minimum of 1 dB/µs is determined by C4 and R5. R5 provides a constant discharge current of approximately 0.5 mA. D5 prevents the ALC voltage from going below –0.6V. Since the ALC line is brought out of the module, additional filter capacitors can be placed across the line to increase the time constant. The short time constant provides an action like an RF clipper, but with less in-band distortion. Intermediate time constants on the order of 2-5 dB/ms provide RF syllabic compression. Long time constants of 25-100 dB/s set the RF output to the correct level but do not modify the modulation.

To limit the maximum gain and place background noises below the compression range, the PC Interface module can set a minimum ALC voltage. It can also be used to eliminate the ALC action on normal audio levels, but ALC is left on all the time so that the power amplifier cannot be accidentally over-driven and generate splatter.

The leveled signal at the output of U3 is amplified by Q8 and routed to the main crystal filter. This additional filtering is necessary to remove any out-of-band distortion products caused by rapid gain variations before transmission. It also provides additional carrier suppression when audio levels are low.

Product Detector/ Balanced Modulator

This module was the most straightforward to design. (See Fig 4.) The crystal filter and mixer are both bilateral and so are used for both SSB generation and detection. Q1 is an amplifier that increases the +4 dBm from the DDS BFO up to +10 dBm. This is a common-emitter amplifier with emitter degeneration to set the gain and a low-Q fixed-tuned L-network in the collector circuit for impedance matching. The BFO signal is attenuated 3 dB before application to the level-7 mixer (Z1).

The received signal first passes through a crystal filter (FL1) to strip away excess noise from the wide-band IF strip and eliminate the audio image. This filter need not have tremendous selectivity. I used a five-pole, 2.5-kHz-bandwidth filter from KVG, which has a minimum stopband attenuation of 50 dB. (This is about the minimum required.) Combined with the 80-dB minimum stopband attenuation of the filters ahead of the IF amplifier, at least 130 dB of attenuation is presented to out-of-band signals. This is the minimum necessary for a receiver with 120 dB of AGC range. Two L networks, C1/L1 and C2/L2, provide impedance matching. If an International-Radio

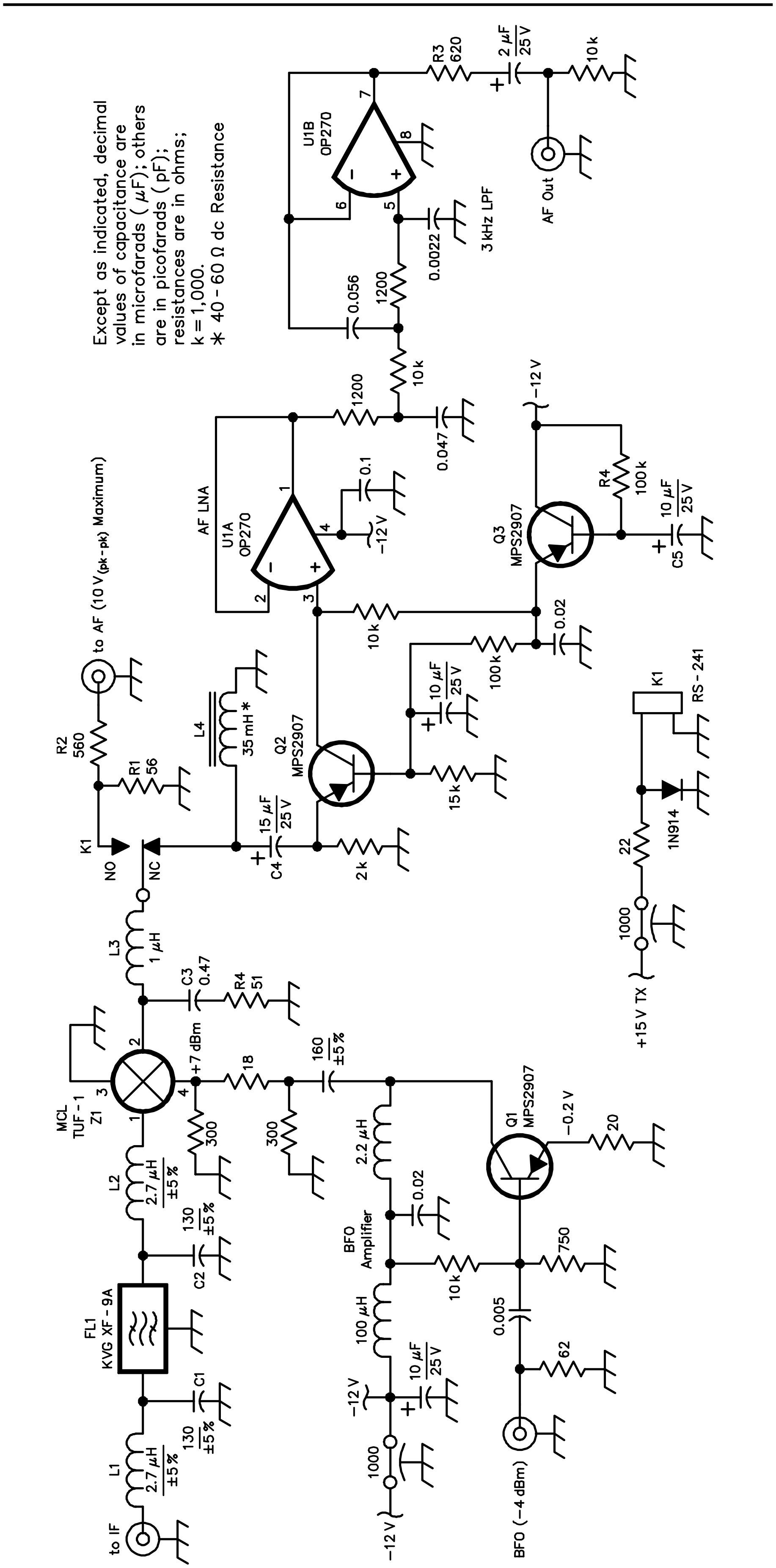


Fig 4—Product detector/balanced modulator circuit. Unless otherwise specified, use ¼ W, 5%-tolerance carbon composition or film resistors.

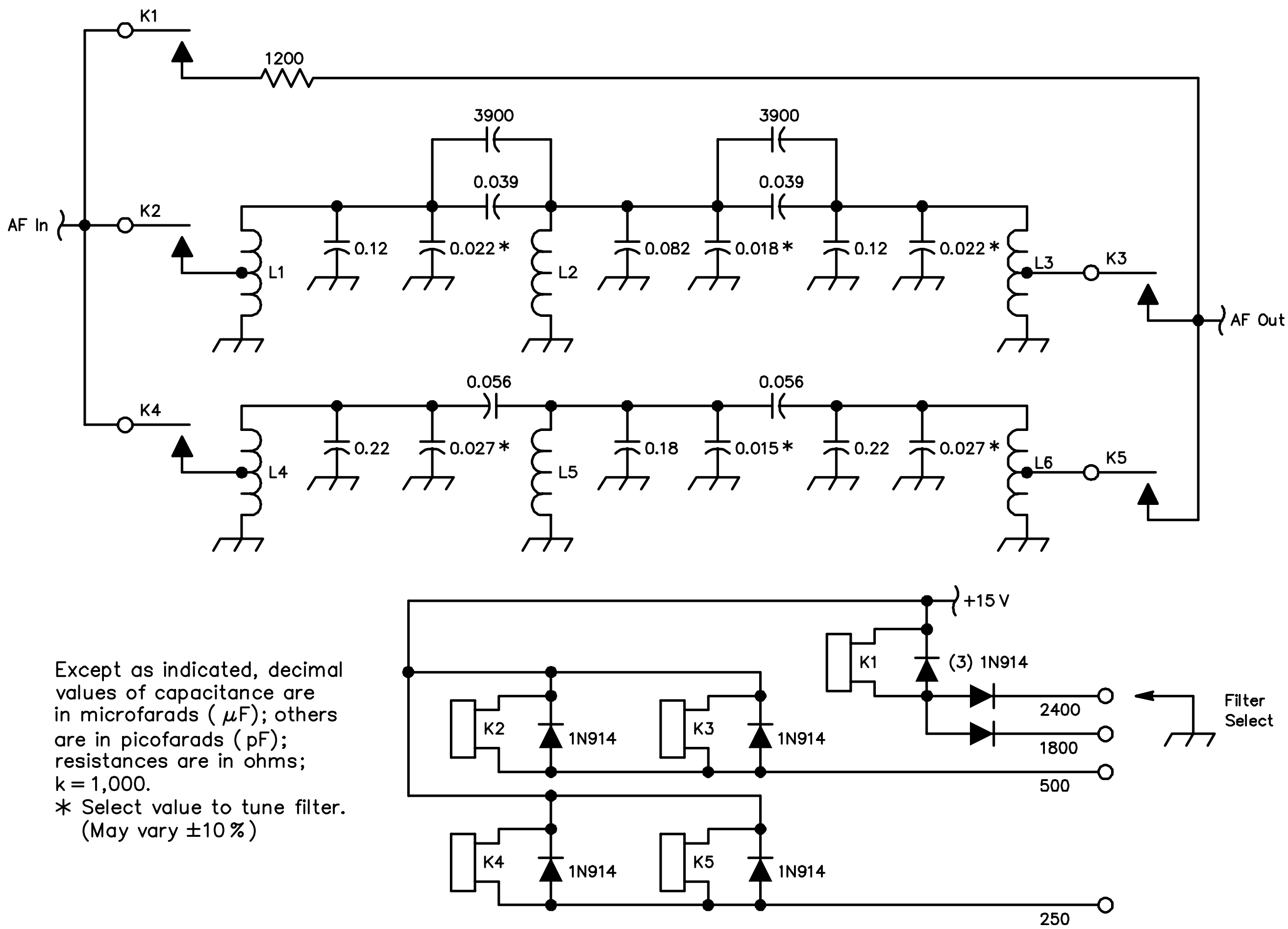


Fig 5—AF filter schematic diagram. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. All capacitors are ±5% tolerance mylar components.

K1-K5—Reed relay, SPST, 12-V dc coil.

L1-L6—88 mH toroid inductor, center tapped.

filter is substituted, C1 and C2 must be reduced to 100 pF each.

The double-balanced mixer (DBM) that follows the filter is used as the product detector. Since the IF level at this point is –77 to –37 dBm, low IMD is assured. The DBM is followed by L3, C3 and R4, which form a diplexer with a 6-kHz transition frequency. This is followed by a relay to switch between incoming transmitter audio and the receive audio-amplifier chain.

While receiving, a low-noise audio amplifier follows the relay. L4 provides a dc return for the DBM and forms a 220-Hz high-pass diplexer with C4. Q2 is used as a common-base amplifier and is biased to have a 50-Ω input impedance. An operational amplifier, U1A, configured as a voltage follower, buffers the output. This is followed by U1B, configured as a low-pass filter to attenuate high-frequency audio hiss. The filter is a three-pole Chebyshev with 0.5 dB of ripple and a cutoff frequency of 3 kHz. R3 establishes the output impedance. The voltage gain through the amplifier and filter is 40 dB.

Transmit audio is applied to the DBM through R1 and R2, which form a matching network and attenuator. The maximum audio input of 10 V (pk-pk) results in +3 dBm at the IF port of

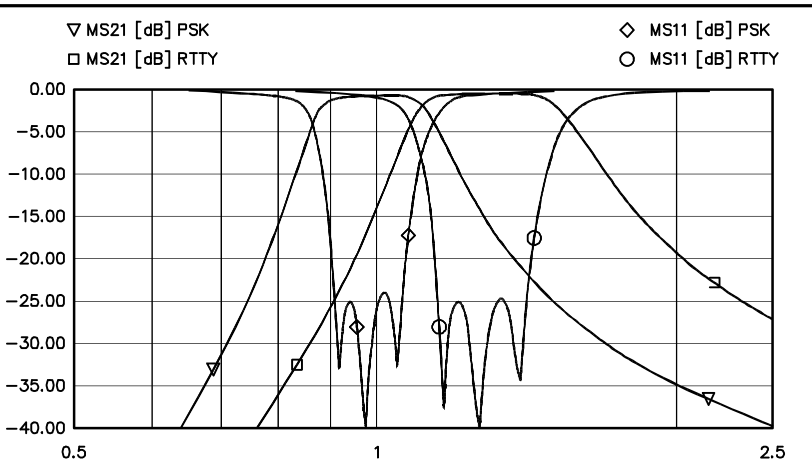


Fig 6—AF filter attenuation (S21) and return loss (S11).

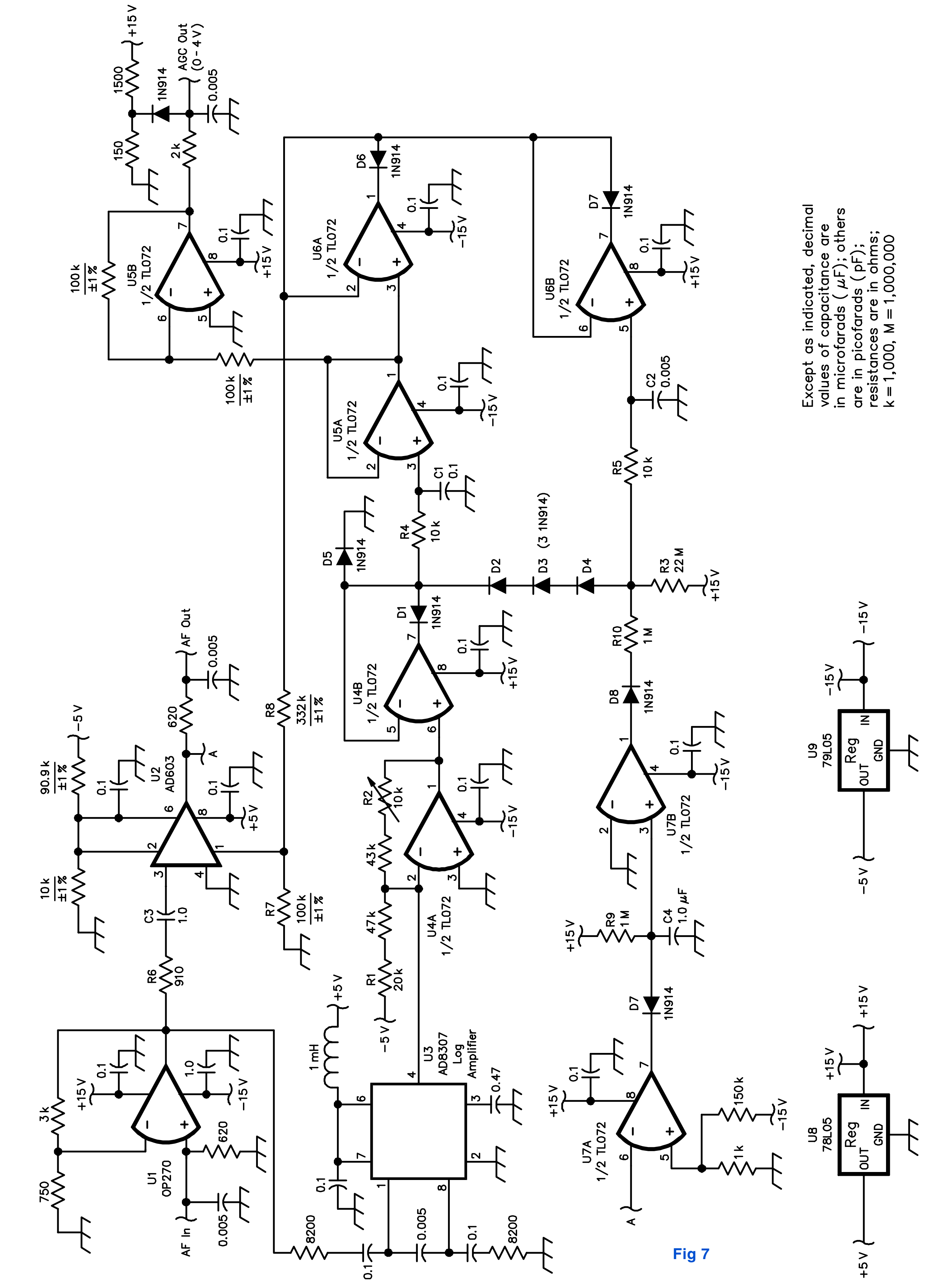

U1
OP270
AF In
750
3k
620
0.005
+15 V
−15 V
0.1
1.0
R6
910
C3
1.0
10 k
±1%
90.9 k
±1%
−5 V
U2
AD603
+5 V
620
A
AF Out
0.005
R7
100 k
±1%
R8
332 k
±1%
8200
0.1
U3
AD8307
Log
Amplifier
1 mH
0.47
−5 V
R1
20k
47k
43k
R2
10k
U4A
1/2 TL072
U4B
1/2 TL072
D1
1N914
D5
1N914
R4
10 k
C1
0.1
D2
D3
D4
(3 1N914)
U5A
1/2 TL072
100 k
±1%
U5B
1/2 TL072
150
2k
1500
1N914
AGC Out
(0 - 4 V)
U6A
1/2 TL072
D6
1N914
U6B
1/2 TL072
D7
1N914
C2
0.005
R5
10 k
R3
22 M
R10
1 M
D8
1N914
U7B
1/2 TL072
R9
1 M
C4
1.0 μF
U7A
1/2 TL072
150k
1k
U8
78L05
U9
79L05
Reg
IN
OUT
GND
Except as indicated, decimal
values of capacitance are
in microfarads (μF); others
are in picofarads (pF);
resistances are in ohms;
k = 1,000, M = 1,000,000

Fig 7

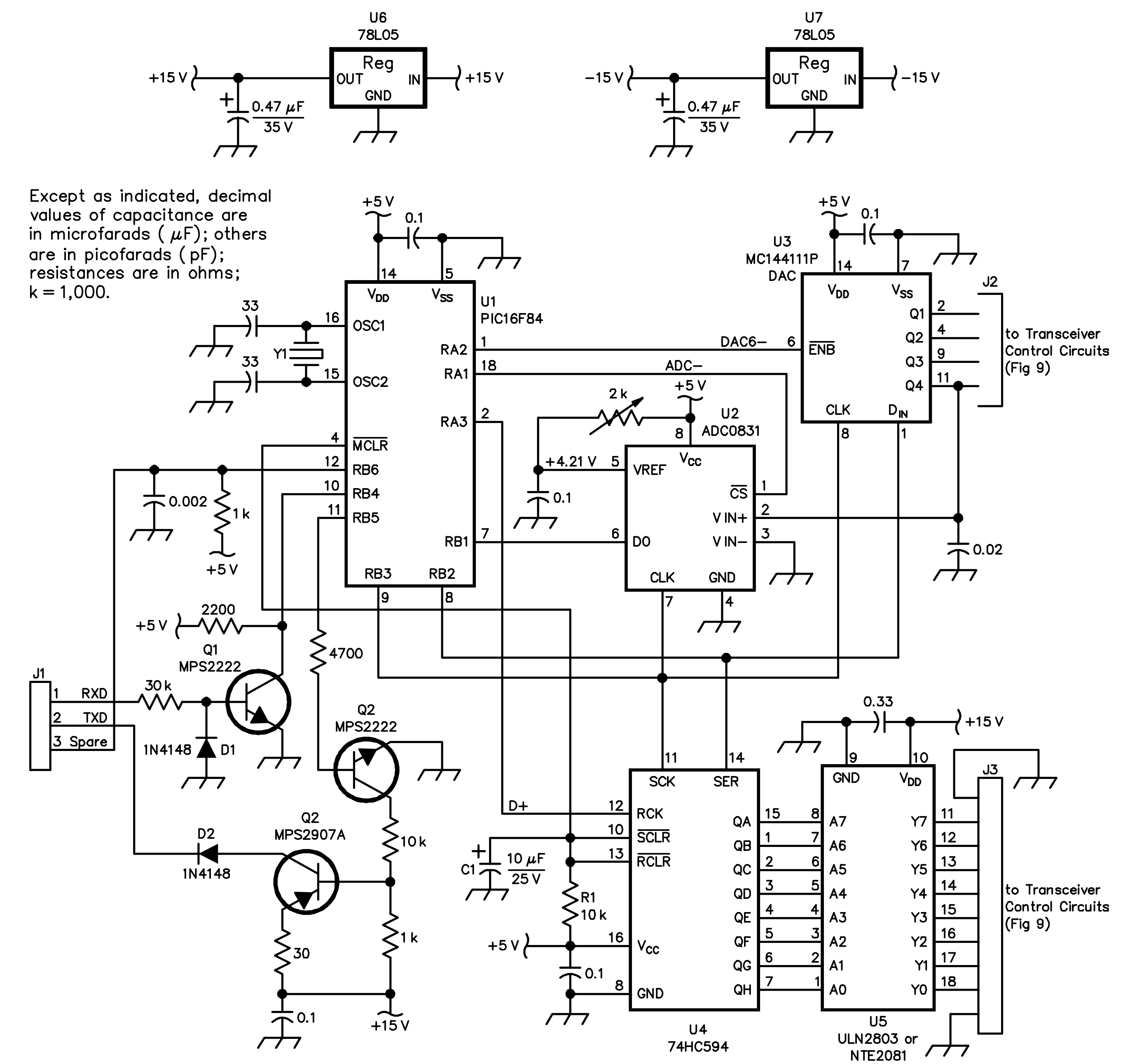


Fig 8—PC interface schematic diagram. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors.

J1-J3—Molex plug, 0.1-inch spacing in line.

Y1—7.3728 MHz crystal, 20 pF parallel resonant.

Z1. This signal level should result in IMD of –36 dBm or less; the audio level is usually much lower and results in less IMD. Z1 has a minimum LO/RF isolation of 50 dB and provides adequate carrier suppression without adjustment. FL1 suppresses the unwanted sideband before speech processing.

Fig 7—AF amplifier/AGC schematic diagram. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors.

Audio Filtering

Audio filters may be used more effectively here than in other designs because the IF-amplifier gain is relatively low, the audio image has been stripped by the tail-end filter and the main AGC loop uses an audio detector. The narrow-band crystal filters have 4:1 shape factors and 90 dB ultimate attenuation, so relatively little filtering is needed. The main purpose is to remove IF amplifier noise outside the crystal-filter passband and increase ultimate attenuation.

The filters use 88-mH toroidal inductors[4] with capacitive coupling to minimize the type and number of inductors. (See Fig 5.) The PSK filter is 270 Hz wide at –3 dB and is centered on 1000 Hz. The RTTY filter is 460 Hz wide at –3 dB and centered on 1360 Hz for use with a demodulator using 1275/1445 Hz tones. Fig 6 shows both the PSK and RTTY filter characteristics. When the filters are not in use, a 1200-Ω resistor provides 6 dB of attenuation to compensate for the lower loss of the SSB filters in the IF.

AF Amplifier and AGC

Fig 7 shows the AF amplifier and AGC circuitry. U1 amplifies the received audio signal by 17 dB to present the proper levels to the variable-gain audio amplifier. Connected to the output of U1 is U3, an Analog Devices AD307 logarithmic amplifier. U3 and U4A are used as the AF AGC detector. R2 sets the slope to –100 mV/dB and R1 sets the power level at which the output is zero. This level is set to –107 dBm or 1 µV RMS at the antenna terminals with the noise blanker bypassed. This is 5 dB above the minimum noise level. The upper end of the AGC range is 1 V RMS. The IF AGC detector was designed to give a slightly lower AGC voltage with the same input signal level, so the amount of adjustment required by the slower audio-derived AGC is minimal. However, the audio AGC provides the final, more-accurate control that is reflected by the S-meter. The AD8307 and AD603 are each accurate to ±1 dB over the operating temperature range. The meter reading is accurate to ±2 dB or 1/3 S-unit.

U4B and D1 form a gate that charges the AGC filters to the peak negative level of the detector output. Two AGC filters are provided. A slow filter drives the IF amplifier AGC line. The IF amplifier gain cannot be adjusted rapidly because of delay in the crystal filter preceding the product detector. A fast filter is used to develop AGC for the variable-gain AF amplifier. The fast and slow AGC voltages are compared and the most negative voltage is applied to the amplifier. The fast AGC voltage compensates for excess IF amplifier gain during transient conditions by decreasing AF amplifier gain.

The circuit is somewhat complex. R3 provides a constant-current discharge path for the slow and fast filter capacitors, C1 and C2. The resulting decay times produce a gain increase of 20 dB/s for the slow AGC voltage and 4 dB/ms for the fast AGC voltage. This allows tracking of fading signals. The fast AGC voltage is offset by +0.65 V from the slow AGC voltage by D2 through D4, so that audio peaks must be 6.5 dB higher than the average level to affect audio gain. This eliminates excess pumping of the AGC by audio peaks, but allows fast response to transients at the beginning of a transmission. D5 clamps the no-signal voltage to +0.4 V. Attack times are set by R4 and R5 to 10 ms for the slow AGC voltage and 50 µs for the fast AGC voltage. The attack and release times are somewhat critical for good SSB reception. I spent two days tuning these time constants for rejection of impulse noise and for minimum audio distortion.

U5A is a voltage follower to isolate the holding capacitor. U5B inverts the AGC voltage for application to the IF amplifier. U6, D6 and D7 comprise a gate to select either the fast or slow AGC voltage for application to U2, which provides an adjustable gain of –10 to +30 dB. Note that R6 and C3 provide 20 dB of attenuation and transform the 100-Ω input impedance of U2 to 1000 Ω. R7 and R8 convert the 0-4 V AGC signal to 0-0.97 V to control audio gain. A 40-dB increase in signal at the AGC detector results in a 37-dB reduction in gain. The 3-dB increase is left to compensate for variations in gain slope and logarithmic-amplifier accuracy to ensure that an AGC-detector signal increase never results in a decrease in audio level.

U7 provides the "hang" AGC function that causes rapid gain increase if a signal is lost completely for more than the "hold" time. U7A discharges C4 through D7 whenever the audio output is greater than 100 mV. This causes the output of U7B to go negative. C4 is charged through R9 so that after 500 ms, U7B's output will go positive. This causes C1 to be discharged through D8 and R10, increasing the gain at a rate of 400 dB/s. This allows full gain to be achieved within a short time after the disappearance of very strong signals, rather than having to wait up to six seconds at the normal 20 dB/s rate. It is not really needed on signals below S9, but it helps when strong local signals are present.

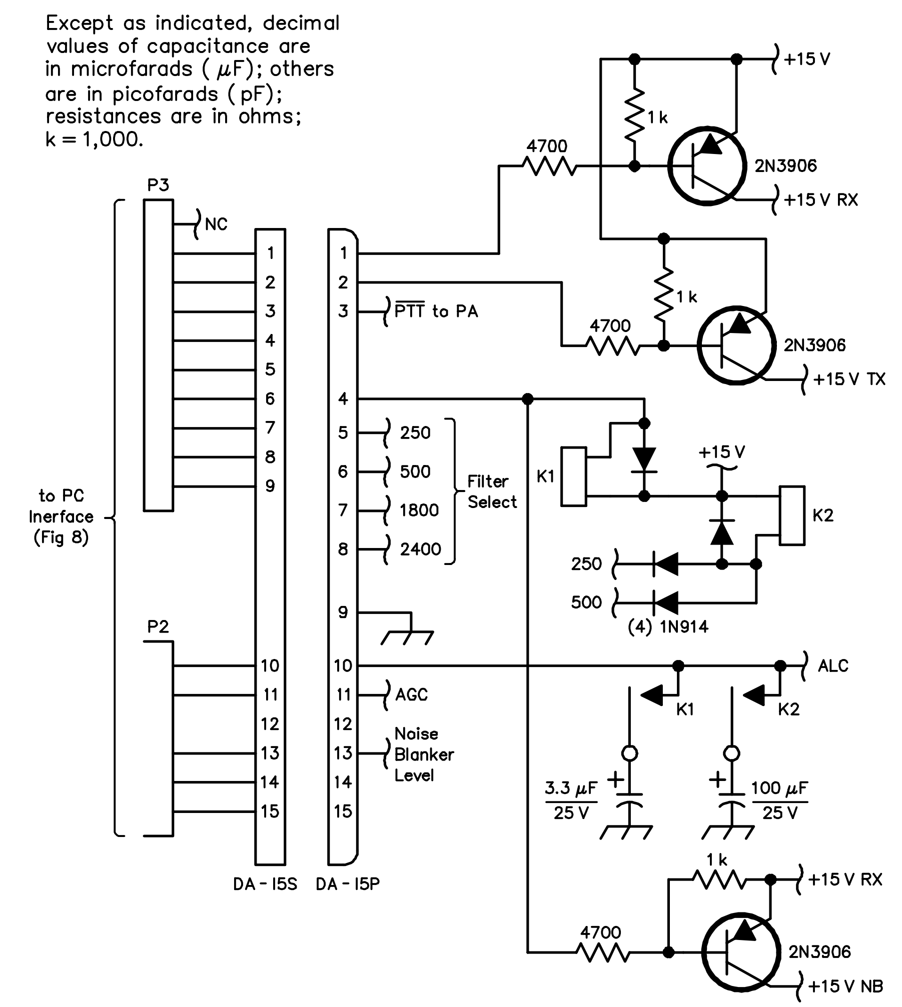


Fig 9—Transceiver control circuits. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors.

K1, K2—Reed relay, SPST, 12 V dc coil.

P2, P3—Molex plug, 0.1-inch spacing in line.

PC Interface

The entire radio is controlled from a personal computer. The method of controlling the DDS was described in Part 1, but there are several functions still left: filter switching, transmit RF compressor control, IF gain control and TR switching. These are controlled by a second microprocessor (MCU) as shown in Fig 8.

The MCU, U1, is another PIC16F84 with the same UART software as described in Part 1. Q1, D1 and associated resistors form the EIA-232 receiver, whose input is wired in parallel with the receiver in the DDS circuit. Both MCUs receive the same commands but only one executes each command. U1 recognizes only the commands in Fig 9.

Unlike the DDS-control MCU, this MCU can also send responses to the PC. Q2, Q3, D2 and associated resistors form the EIA-232 driver. This driver is designed to have its output wired in parallel with other similar units to allow sharing of one PC serial port among multiple radios. The driver sources 20 mA when sending a zero and is completely inert when sending a one. To pull the line to a negative voltage (one) when no MCU is sending data, the output is terminated near the PC with a 1500-Ω pull-down resistor to a –12 V supply. The diode, D2, in combination with the base-collector junction of Q3, ensures that the driver will not source or sink current when power is removed. This allows one or more radios to be turned off while controlling others from the PC.

Bits 1, 2 and 3 of MCU port A are used to select the peripheral chips being read or written to by the MCU. Data are written to the selected chip by shifting the data using bit 3 of MCU port B as the clock, and bit 2 as data-output pin. Data are read by shifting them into the MCU using bit 3 as the clock and bit 1 as the data-input pin.

U2 is an 8-bit analog-to-digital converter (ADC) that is used to sample the AGC input to the IF amplifier. R2 sets the reference voltage so that the conversion slope is 1 bit/0.5 dB of receiver gain. U3 contains four 6-bit digital-to-analog converters (DACs) that are used to control IF gain, RF clipping level and noise-blanker gain. One DAC is unused. J2 connects the DACs and ADC to the rest of the transceiver.

U4 is a CMOS shift register and latch used as a parallel output port. Eight Darlington transistors, contained in U5, buffer the output of U4. The outputs are open-collector and are connected via J3 to the circuit shown in Fig 9 to control power to the noise blanker, transmit and receive sections of the transceiver and the filter-selection relays.

The commands used to control the receiver are shown in Fig 10. They have the structure defined in Part 1. In the case of commands involving the DACs, the command contains a data byte with a six-bit value that is output to the appropriate device. The filter-selection command uses four bits to control the filter-selection relays. Other commands have no data field. The response to an AGC Request command contains the eight-bit digitized value of the IF AGC line.

Note that the transmit/receive timing is controlled by the MCU. When a transmit command is received, the MCU mutes the receiver, switches out the receive audio stages and the receive IF amplifiers and enables the TR relay. After a time delay for the TR relay to settle, it switches in the transmit IF amplifier and compressor, waits for transients die out, then ramps the IF amplifier gain up to the desired clipping level. The reverse sequence is executed on receipt of the command to go to the receive mode. Table 2 lists other internal transceiver control lines.

Test Results: AGC System

Several performance parameters of the transceiver were tested on the bench as described below. Dynamic AGC response was tested using a pulsed signal at various levels from –60 to +10 dBm with a 10-s period. There was no overshoot on the AGC voltage applied to the IF amplifier. Overshoot of the audio output never exceeded 7 dB of the final value and undershoot was about 3 dB maximum. Stabilization of the output level occurred within 30 ms.

The AGC system was also tested for linearity. Table 3 shows the measured response of the receiver to a CW carrier with the IF preamplifier out of circuit. The signal levels decrease by 11 dB when it is switched in. However, the software displaying the signal strength can easily compensate for this.

The AGC detector slope and intercept were adjusted using 1-µV and 1-V signals from a HP 8640B signal generator. The generator's output is accurate to within ±1.5 dB and the voltmeter is

Table 3—IF Amplifier AGC Response

Signal (µV RMS)	Signal (dBm)	AF AGC (V dc)	IF AGC (V dc)	AF AGC (dB)	IF AGC (dB)	AF–IF AGC (dB)
1,000,000	+13	3.99	1.92	–78.8	–76.8	2.0
316,000	+3	3.62	1.68	–72.4	–67.2	5.2
100,000	–7	3.28	1.44	–65.6	–57.6	8.0
31,600	–17	2.96	1.22	–59.2	–48.8	10.4
10,000	–27	2.63	0.98	–52.6	–39.2	13.4
3,160	–37	2.29	0.75	–45.8	–30.0	15.8
1,000	–47	1.99	0.52	–39.8	–20.8	19.0
316	–57	1.64	0.31	–32.8	–12.4	20.4
100	–67	1.31	0.10	–26.2	–4.0	22.2
32	–77	0.98	0.01	–19.6	–0.4	19.2
10	–87	0.65	0.00	–13.0	0.0	13.0
3	–97	0.34	0.00	–6.8	0.0	6.8
1	–107	0.00	0.00	0.0	0.0	0.0

Table 2—Transceiver Control Lines

Pin	Digital	Pin	Analog
1	Qa RX	9	Ground
2	Qb TX	10	Q4 Compressor
3	Qc TR Relay	11	Q3 AGC (I/O)
4	Qd ALC Select	12	(No Connection)
5	Qe 250	13	Q2 Blanker
6	Qf 500	14	Q1
7	Qg 1800	15	Ground
8	Qh 2400		

Table 4—RF Compressor Characteristics

Input Level (dBm)	ALC (V)	Output Level (dBm)
–10	4.00	+2.5
–20	3.21	+0.5
–30	2.39	–1.0
–40	1.55	–2.5
–50	0.72	–4.0
–58	0.00	–6.0

Table 5—RF Compressor Characteristics

ALC Filter Capacitor (µF)	Time Constant (dB/ms)	IMD (dBc)
47	0.1	<–40
6.8	0.7	–35
3.3	1.5	–30
1.0	5	–21
none	1000	–12

accurate to within 0.5% or ±0.6 dB. The transceiver was then operated for three days and the response measured. The maximum deviation of the AGC voltage from the ideal response of 33 mV/dB was +0.01/–0.05 V or +0.3/–1.5 dB. This is within the error range of the measurement equipment.

When the AGC value was read from the PC, the displayed value was accurate within ±1 dB for all but the extreme upper end of the range, above 0.5 V of input signal. The maximum excursion was 2 dB at 1 V of RF input.

RF Compression

The RF compression range was first tested by injecting a 9-MHz CW signal into the IF amplifier module and measuring the output level on a HP 8555A/8552B spectrum analyzer. The results are shown in Table 4. They met expectations.

Distortion of the transmitted signal was also measured with various time constants for the ALC. A two-tone audio signal generator (700 and 1700 Hz tones at 0 dBm/600 Ω per tone) drove the balanced modulator and the output was checked with a spectrum analyzer. The results are shown in Table 5.

Only one IMD product (2700 Hz) passes through the final crystal filter, so the IMD levels are measured by referencing that tone to the 1700 Hz tone. No other IMD products could be found above –60 dBc, which was the limit of measurement.

LO Phase Noise

LO phase noise was measured by connecting a low-noise crystal oscillator (described in Part 1) at the antenna terminals and measuring the amplitude of noise sidebands surrounding it on the digital S-meter in the transceiver control program. Measurements were made with 2400-Hz and 250-Hz bandwidths on both sides of the test signal and the results are shown in Table 6. The possible measurement error in the test setup is ±3 dB.

No spurs were noted at ±36 kHz from the signal, so the reference frequency component of the PLL error voltage is adequately suppressed. The spurious components at ±10 kHz and ±20 kHz are less than 50 Hz in width and are probably generated by other equipment in close proximity rather than the transceiver under test. The phase noise is within predicted limits except for the area below 1500 Hz, where it is at least 2-3 dB higher than anticipated, but certainly acceptable for normal use.

Receiver Dynamic Range

Receiver dynamic range was evaluated by measuring the minimum discernable signal (MDS) and the third-order intercept (IP3). An HP 8640 signal generator was used to generate a signal for measuring the MDS and the level required for a 3-dB increase of audio output in a 2.4 kHz bandwidth was measured. In addition, 9 MHz and 32.2 MHz signals were generated to measure IF and image rejection.

Two of the low-noise crystal oscillators described in Part 1 were used with a hybrid combiner for IMD testing. The output from the combiner was 2.5 dBm/tone and a 25-kHz spacing was used. The amplitude of spurious responses was read from the digital S-meter in the transceiver control program. Readings taken from both sides of the tones were averaged.

The dynamic range looks slightly greater than expected because the noise figure of the IF preamplifier is lower than expected in one case, the IP3 higher than expected in another case. However, these measurements are only accurate to ±2 dB. To compare the third-order dynamic range with measurements made by the ARRL standard method, a small correction[5] must be applied. IF and image rejection are more than adequate and LO leakage to the antenna is essentially nonexistent. A measurement summary is presented in Table 7.

Conclusions

Almost all low-level stages of the transceiver worked as anticipated when completed. Two areas, however, required changes during testing. The original design for this transceiver used RF clipping, but on-the-air tests yielded reports of excessive audio distortion when clipping exceeded 5 dB. Consequently, clipping was abandoned as a method for RF compression, which resulted in much cleaner audio. From reports by other stations, two types of

Byte 1	*Byte 2*	*Bytes 2*	*Byte 3*	*Byte 4*	*Command*
STX	B	Bits 0-3	ETX	LRC	Select IF filter
STX	G	0-63	ETX	LRC	Set IF Gain
STX	C	0-63	ETX	LRC	Set RF clipping level
STX	Z	0-63	ETX	LRC	Set noise receiver gain (0=off)
STX	S	ETX	LRC		Request AGC voltage
	s	0-255			AGC voltage (response)
STX	X	ETX	LRC		Go to transmit mode
STX	N	ETX	LRC		Go to receive mode

Fig 10—Transceiver control command strings.

Table 6—Measured LO Phase Noise

	Phase Noise	
Offset (Hz)	*Expected (dBc/Hz)*	*Measured (dBc/Hz)*
0.5 k	–104	–99
1 k	–113	–109
1.5 k	–117	–116
2 k	–122	–119
3 k	–126	–128
4 k	–131	–137
5 k	–133	–140
7 k	–137	<–140
10 k	–141	–140*
15 k	–145	<–140
20 k	–150	–147*
25 k	–152	<–153

*Narrow-band spur

Table 7—Measured Receiver Characteristics with 2.4 kHz IF bandwidth

Measurement	*IF Preamp. On (0 dB Attn)*	*IF Preamp. On (3 dB Attn)*	*IF Preamp. Off (3 dB Attn)*
Third Order Intercept (IP3)	32.8 dBm	35.0 dBm	37.0 dBm
Minimum Discernable Signal (MDS)	–128 dBm	–125.5 dBm	–122.0 dBm
Noise Figure	12.0 dB	14.5 dB	18.0 dB
Spurious Free Dynamic Range	107.2 dB	107.0 dB	106.0 dB
LO Leakage	< –85.0 dBm	–	< –85.0 dBm
Image Rejection	–104.0 dB	–	–108.0 dB
IF Rejection	–92.0 dB	–	–92.0 dB

compression seem to be desirable. A medium ALC time constant (1.5 dB/ms) for syllabic compression results in low distortion on good paths and better readability than long time constants. Under marginal conditions, the fastest possible ALC time makes signals readable where syllabic compression is ineffective.

The other problem area was the mixer. Although most literature indicates that the IF port must see a broadband 50-Ω load for low IMD, it was found that the RF port is actually more sensitive to termination in a high-level DBM. The RF band-pass filter described in Part 1 and the IF diplexer described here were necessary to cure an initial 15-dB deficit in IP3.

The low noise figure and high dynamic range (with the 3-dB IF attenuator removed) are useful on the 10-meter band where any RF amplifier would degrade the dynamic range. Atmospheric noise on this band still exceeds the noise figure by 6 dB or more.

The next part of this series will cover the linear amplifier, noise blanker and power supply.

Notes

[1]John Stephensen, KD6OZH, "The ATR-2000: A Homemade, High-Performance HF Transceiver, Pt 1" *QEX*, Mar/Apr 2000, pp 3-15.

[2]Ulrich L. Rohde and T. T. N. Bucher, *Communications Receivers—Principles and Design*, McGraw Hill, 1988.

[3]KVG filters are no longer distributed in this country. Similar filters are available from International Radio, 13620 Tyee Rd, Umpqua, OR 97486; tel 541-459-5623 (9 AM-1 PM PDT, Tues-Sat), fax 541-459-5632; e-mail **inrad@rosenet.net**; **http://www.qth.com/inrad/**. Suitable substitutes are part numbers 2301 (250 Hz), 2302 (500 Hz) and 2310 (2400 Hz). In all cases, the extra termination capacitors (30 or 24 pF) must be removed from the circuit.

[4]These were obtained from surplus telephone loading coils. An alternative is to wind an inductor on a ferrite pot core. 155 turns of #28 AWG enameled wire on an Amidon PC-2213-77 core provides 88 mH.

[5]The dynamic range should be about 4.7 dB greater with a 500 Hz bandwidth.

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 homebuilt 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.

Next Issue in QEX/Communications Quarterly

L. B. Cebik continues his series on LPDAs with a look at some slightly larger arrays (164-foot booms). L. B. explores tweaking and optimization of design parameters with as many as 42 elements. Segmentation limi-tations imposed by software and computing power are addressed. This is serious stuff for those of you interested in HF gain that spans more than an octave of frequency. Do you have a few acres lying fallow?

Paul Hewitt, WD7S, gives you something to drive your six- and two-meter arrays: A no-bandswitch, dual-band, legal-limit linear amplifier. This project neatly exploits transmission-line theory to achieve its dual-band aspect. Come take a spin around the impedance chart with us and boost your signals.

Add synthesis and computer control to your HW-101 or other older rig! Rick Peterson, WA6NUT, brings us a PLL "spur eliminator." It's a PLL you can drive with your PC-controlled DDS to reduce or eliminate spurs outside the loop bandwidth. The PLL's VCO output is used to drive a transceiver's BFO or LO input. Remote control also creates some interesting possibilities for the experimenter. In the first part of a series, Sam Ulbing, N4UAU, presents some work he has done to control his rigs over UHF links. He uses off-the-shelf UHF transceivers and explains the difference between Part-97 auxiliary operation and Part-15 operation.