3
Transmitting and Receiving Equipment

Contents

A Simple Beacon System

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(From *Proceedings of Microwave Update '91*)

Introduction

Beacon transmitters provide several primary benefits. They expose enhanced propagation, indicating conditions we might otherwise not observe. They produce a constant signal source, helping assure that receiving systems are functioning properly. On some bands, beacons mark the narrowband/weak-signal portion of the band, discouraging encroachment of wide-band activities. Band plans and a beacon list are published in the ARRL *Repeater Directory*. The beacon system described in this paper is proving beneficial on all counts.

Perceiving operational and technical obstacles, few are willing to equip stations for relatively unoccupied bands. Many amateurs are unfamiliar with weak-signal VHF/UHF/microwave propagation characteristics, believing the "line-of-sight" myth. Demonstrations of beacon reception to potential VHF/UHF/microwave enthusiasts can help dispel the myth.

This paper documents progress on the N4MW/B beacon system (Fig 1). Since the system design depends on availability of surplus components, a general description emphasizing integration of diverse equipment is presented, rather than a particular construction plan. I hope that the information provided in this paper assists and encourages others to implement needed beacons on multiple bands and in diverse areas.

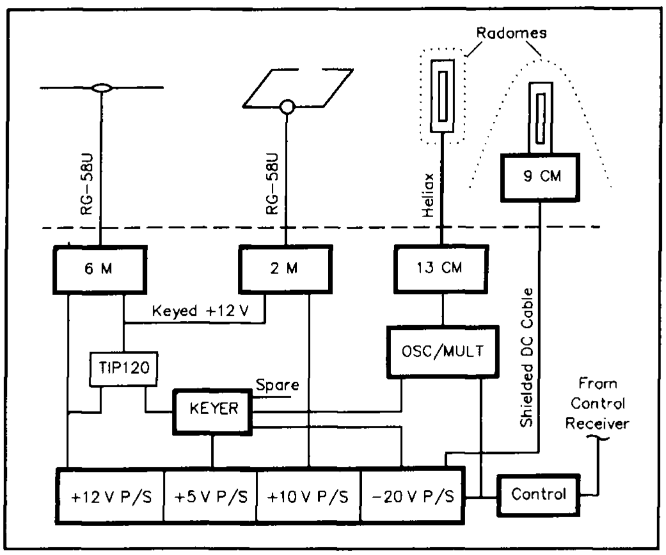


Fig 1—Block diagram of the N4MW Beacon System.

Background

FCC rules permitting beacon operation are found in Part 97 Sections 97.84 and 97.87. Beacon transmissions are defined as one-way transmissions conducted to facilitate measurement of radio-equipment characteristics, observation of propagation or transmission phenomena, or other related experimental activities. Rules for unattended beacons are stricter than for attended beacons. Here is a summary of the rules pertaining to unattended (automatically controlled) beacons:

- Operation is limited to one concurrently operating transmitter per band.
- Appropriate control provisions/procedures are required.
- If the FCC orders, operation must cease until problems are resolved.
- Operation is permitted on the following frequencies:
  - 28.2-28.3 MHz
  - 50.06-50.08 MHz
  - 144.275-144.3 MHz
  - 222.05-222.06 MHz
  - 432.3-432.4 MHz

Operation is permitted on any authorized (non-specific) frequency above 450 MHz.

- Below 450 MHz, permitted emissions are N0N A1A, F1B or J2A. Above 450 MHz, any authorized emission type is permitted. For F1B and J2A emissions, maximum radio- or audio-frequency shift is 1000 Hz.
- A Technician-class or higher license is required.

Power output is limited to the minimum necessary, but may not exceed 100 watts.

- For Morse identification, use the /BCN or /B call-sign suffix.
- The maximum interval between identifications is one minute.

ARRL has adopted beacon subband recommendations in addition to those specified in the FCC Rules. These subbands are:

902.2-902.4 MHz
1296.3-1296.4 MHz
2304.3-2304.4 MHz
3456.3-3456.4 MHz
5760.3-5760.4 MHz
10368.3-10368.4 MHz

Design

All transmitters are commercial or military surplus, keyed simultaneously using a single keyer. Optoisolators are used instead of mechanical keying relays, providing a flexible and reliable means of keying different types of transmitters with different keying requirements from a common keyer. Common power supplies provide power to multiple transmitters where needed. Equipment is located inside a controlled access room, mounted in a locked EIA cabinet rack. On-off control of primary power is provided by a 23-cm FM link and a DTMF (Touch-Tone) control-code sequence.

Keyer

The beacon code keyer is the standard W4RFR EPROM design with the relay eliminated (Fig 2). Four 4N30 Darlington-output optoisolators provide outputs to key the various transmitters. Ideally, the transmitters would all be keyed by switching a logic-level input to ground. More often, current must be switched at the power supply input to the transmitter amplifier stages. The 4N30 will not handle enough current, so a TIP20 Darlington transistor is added as a current amplifier.

The keyer is constructed on a Radio Shack project board. The four optoisolators are mounted on the keyer board. The board is mounted in an aluminum box. A barrier terminal strip allows for connection to a +5-volt supply and to the various keyed devices. An externally visible LED flashes in sync with the keyer.

Programming the EPROM is simple, but requires an EPROM "burner." The 2716 EPROM is a 2k × 8-bit device. The beacon keyer uses only 2 of the 8 bits: one for the code sequence and the other for resetting at the end of the message. Each bit represents one element of a Morse character. Dots are 1 bit; dashes are 3 bits. One bit separates dots/dashes within characters. Character spaces are 3 bits and word spaces are 7 or 9 bits. Long dashes are 50 or more bits long. The EPROM is programmed to contain byte values of 1 or 0 to represent Morse elements and element spaces. A string of 2s forces reset to the beginning of the message. The rest of the EPROM con-

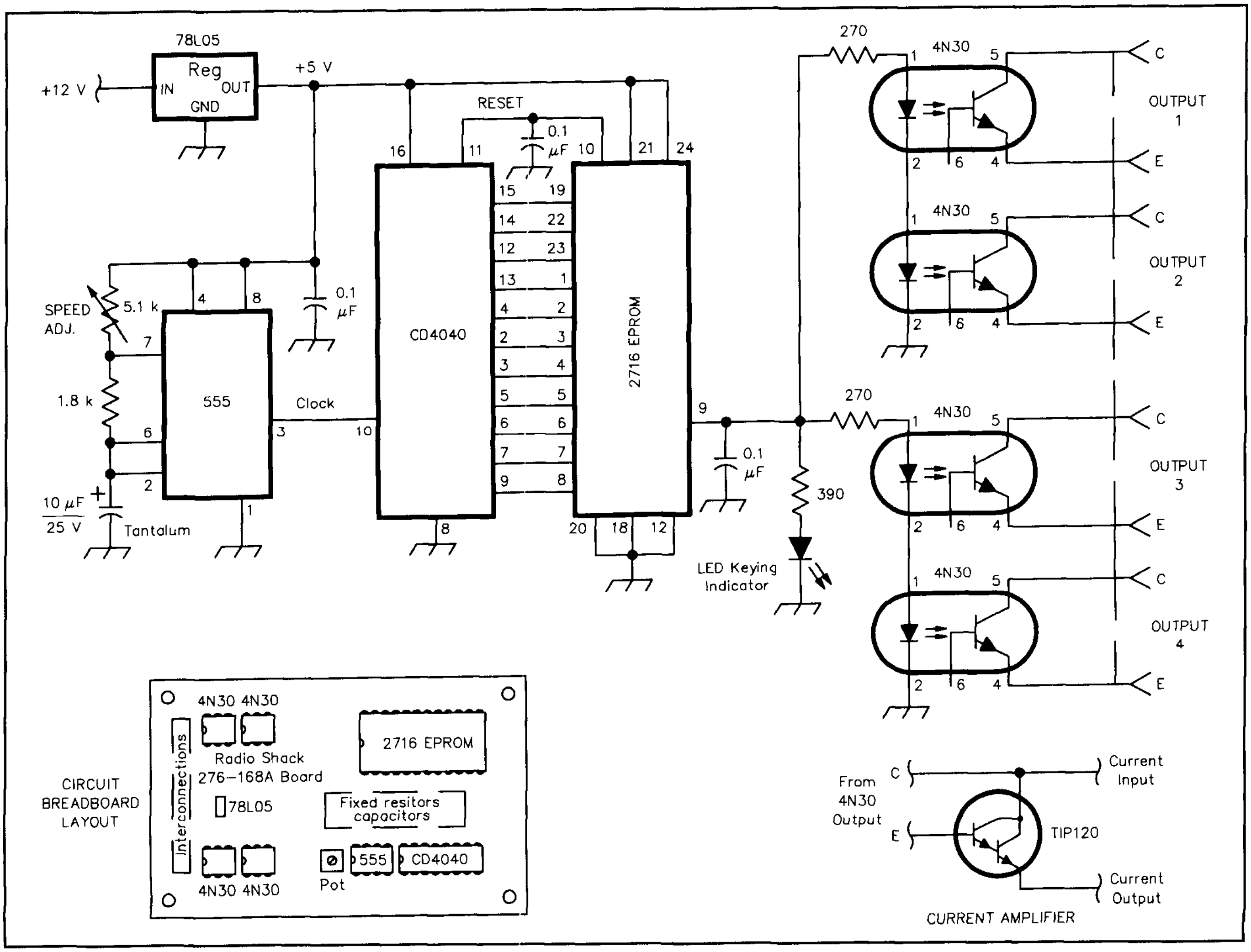


Fig 2—Schematic and layout of the modified W4RFR Beacon Keyer.

tents do not matter. The EPROM message can be seen in the following example, which uses 215 of 2048 bytes, or about 11% of the potential message capacity. For the code sequence:

DE*(WORD SPACE)*N4MW/B*(WORD SPACE)*EM55*(LONG DASH)(SHORT PAUSE)(RESET)*

Program the EPROM as follows:

111010100010000000011101000101010101110001110111000101110111
00011101010111010001110101010000000010001110111000101010101000
1010101010001111111111111111111111111111111111111111111111
1111100000000000000000000000022222

The message content must be arranged to provide at least one identification each minute. This can be done by keeping the message short, or by interspersing two or more identification sequences within the message string. Keying speeds between 10 and 15 WPM are reasonable. A long dash is desirable for equipment adjustment and strength measurement.

Equipment

With 70- and 23-cm beacons already operating locally, 2 meters seemed both the easiest and most immediately useful band to implement. I mounted an exciter board from an RF Communications Corp. RF-403 FM transmitter in an aluminum box. A BNC connector and 2 feedthrough/bypass capacitors provide all necessary connections. A constant +10 V supplies the oscillator and low-level transmitter stages. A keyed +12-V line to the high-level stages produces a chirp-free 4-W signal on 144.280 MHz. A simple gamma-matched "squalo" built from copper pipe produces reasonably omnidirectional, horizontally polarized radiation.

The next band I added was 13 cm. I retuned a Motorola transmit strip to provide 72-MHz output. A homebrew multiplier produces 288-MHz drive and is keyed by one of the keyer optoisolators. A surplus AN/GRC-144 telemetry transmitter provided a multiplier/amplifier assembly and diode doubler capable of transforming the 288-MHz input to 4 watts at 2304 MHz. I constructed an Alford slot from copper pipe, and fed it through 10 meters of ½-inch Heliax. I made a weatherproof cover for the antenna from a length of PVC pipe, capped at the top. The cable end and antenna slips inside the PVC cover, which is clamped to the top of a mast, so that the antenna clears the mast.

The 6-meter transmitter, a General Electric exciter board, requires a constant +12 V and keyed +12 V. To minimize current switched by the primary-keyed +12-V source, I installed another optoisolator and Darlington transistor inside the transmitter box, to switch the constant +12 V. The antenna is a dipole oriented east/west.

The 9-cm beacon uses a Frequency West PLO "brick" source producing 50 mW to an Alford slot. Source and antenna are mounted on a mast pipe under a weatherproof cover made from a TVRO feed cover. dc voltage is supplied through a shielded cable. Power supply current must be continuous to maintain frequency lock. An expedient keying method is to frequency-shift key the source, by slightly varying the negative 20-V supply. One of the keyer optoisolators modulates the dc supply by about 0.1 V, producing a distinctive frequency shift. The supply voltage is modulated by one of the keyer optoisolators, which switches a resistor in the power-supply feedback circuit.

The 13- and 9-cm Alford-slot antennas are based on the G3JVL 2304-MHz Alford slot (Table 1 and Fig 3).[1] Tube wall thickness and length beyond the slot should not be critical. Coaxial 4:1 balun length is for PTFE dielectric line (0.141-in. diameter semirigid coax). Where feasible, make the balun from the feed line, thus eliminating connectors.

TABLE 1
Alford Slot Antenna Scaling Dimensions

Dimensions (Inches)	*FREQUENCY (MHz)* 902.3	1296.3	2304	3456	5760	10368
Tube Diameter	1.86	1.29	0.73	0.49	0.29	0.16
Tube Length	24.23	16.86	9.49	6.33	3.8	2.11
Slot Width	0.26	0.18	0.1	0.07	0.04	0.02
Slot Length	23.02	16.02	9.02	6.01	3.61	2
Balun Length	2.61	1.82	1.02	0.68	0.41	0.23

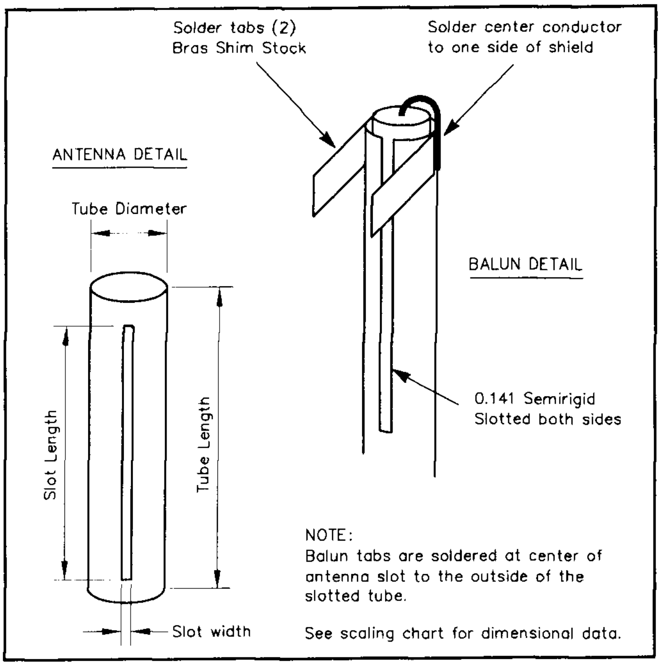


Fig 3—Alford Slot Antenna Construction.

[1]M. Walters, G3JVL, "An Alford slot antenna for 2.3 GHz," *Radio Communication*, June 1983, p 527.

Results

Reception reports have been received from several distant stations. Paul Wilson, W4HHK has used the 13-cm beacon to help diagnose a failed preamp. Rex Turner, W5RCI, receives the 13-cm and lower-frequency beacons consistently over a 100-km (60-mile) path. Al Ward, WB5LUA, has received the 13-cm beacon 20 dB above the noise over a 672-km path. At my home, 14-km from the beacon site, the 13-cm beacon is received with no antenna connected to the converter! It is interesting to monitor the 13- and 9-cm beacons while mobile in motion—the Doppler and multipath effects resulting from vehicle motion are extremely pronounced, sometimes sounding like aurora. I receive the 9-cm beacon using a TVRO LNA as both antenna and preamplifier inside my shack. W4HHK receives it at ground level, over a 28-km obstructed path.

Conclusion and Acknowledgments

I hope this paper encourages additional beacon operation. Beacon operators everywhere appreciate reception reports, which help to establish normal performance expectations, as well as record unusual propagation trends.

I acknowledge and thank Ray Escue, K4RDK, for securing the beacon site, Ken Schildt, N4VSD, for locating the site, James Butler, KB4LJV, for programming the keyer EPROM, Bill Dearing, N4HKS for the logistics, Paul Wilson, W4HHK for inspiration and assistance, and my wife Cissy, N4ZRW for her understanding and support.

A 432-MHz Transverter

By Greg McIntire, AA5C

I recently have become interested in VHF/UHF and microwave activities. Not having built any equipment that operates on the higher frequencies I decided to pursue the construction of a 432-MHz transverter as a means to learn about the design and construction techniques for UHF. A survey of Amateur Radio publications did not yield any "modern" transverter articles. The few that I did find used tubes.

A block diagram of the station is shown in Fig 1. The IF for the project was to be a Kenwood TS-820S HF transceiver. This radio has transverter input and output ports. The transmit port provides 0 dBm and the receive port is switched between the transverter input and the primary SO-239 connector. Eventually, I plan to add a high-power amplifier to the 432 station. The TR relay to connect the transverter input and output to the antenna is external to the transverter so that it can be moved to the amplifier at a later time.

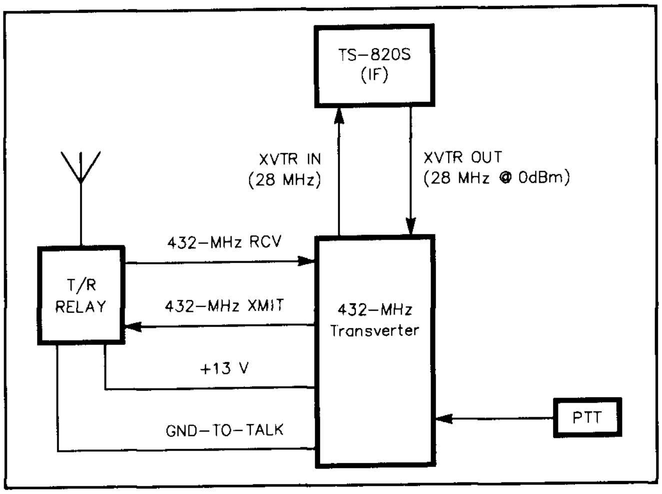


Fig 1—Block diagram of the AA5C 432-MHz station.

Transverter Block Diagram

The block diagram of the transverter is shown in Fig 2. Other than the LO and IF frequencies, there is nothing specific to the 432-MHz band in the block diagram, and this approach is applicable to other frequencies. The basic transverter is a four-port system with the TR relay external to the main unit. This permits additional transmitter power amplifiers to be added to the transmit chain with the TR relay downstream.

The local oscillator multiplies the 101-MHz oscillator by four to 404 MHz. Two times-two multiplier stages are used. Separate mixers are used for transmit and receive converters. A power splitter provides independent outputs for the transmit and receive mixers.

The receive converter chain consists of a low-noise RF amplifier followed by a 432-MHz filter. (The tuned circuits of the RF amplifier serve to limit some of the out-of-band energy.) The two-pole filter provides attenuation of the image frequency ahead of the mixer. The mixer is followed by a two-pole IF filter and an IF amplifier. The IF-amplifier gain can be adjusted to control signal level passed to the HF rig used as the 28-MHz IF.

The transmit mixer inputs are the 28.1 MHz IF (at about 1 mW) and the 404 MHz LO. The transmit mixer is followed by a two-pole filter to attenuate the 376 MHz signal from the output and pass the desired 432-MHz signal. (This should be done right after the mixer before any amplification, else the undesired products will be amplified by the untuned MMIC chain.) The MMIC chain amplifies the 432-MHz signal to the maximum MMIC power levels: Transistors then take over. Impedance-matching circuits in the transistor stages serve to limit out-of-band signals. The final stages bring the transmit signal up to about 7 W.

Construction Techniques

Double sided G10 circuit board material was chosen because the second side of the board makes a good ground plane. This ground plane is a necessity for microstrip circuits, which are used in the power amplifier of this transverter.

I started with a 5- × 18-in. board (Fig 2). The 5-in. dimension was determined by a heat sink I had on hand. The maximum width of the transverter would match that of the TS-820S because I wanted to place the unit on top of the TS-820S.

General construction guidelines that I followed in the unit include:

1) Placing all active devices and circuits on the top side of the board

2) Using 1000-pF ceramic feed-through capacitors to get to the bottom side of the board. I used the solder-in type and soldered around the entire circumference of the feed-through on both sides of the board. This connects the two ground planes together.

3) I bypassed each feedthrough with a 0.01-µF ceramic ca-

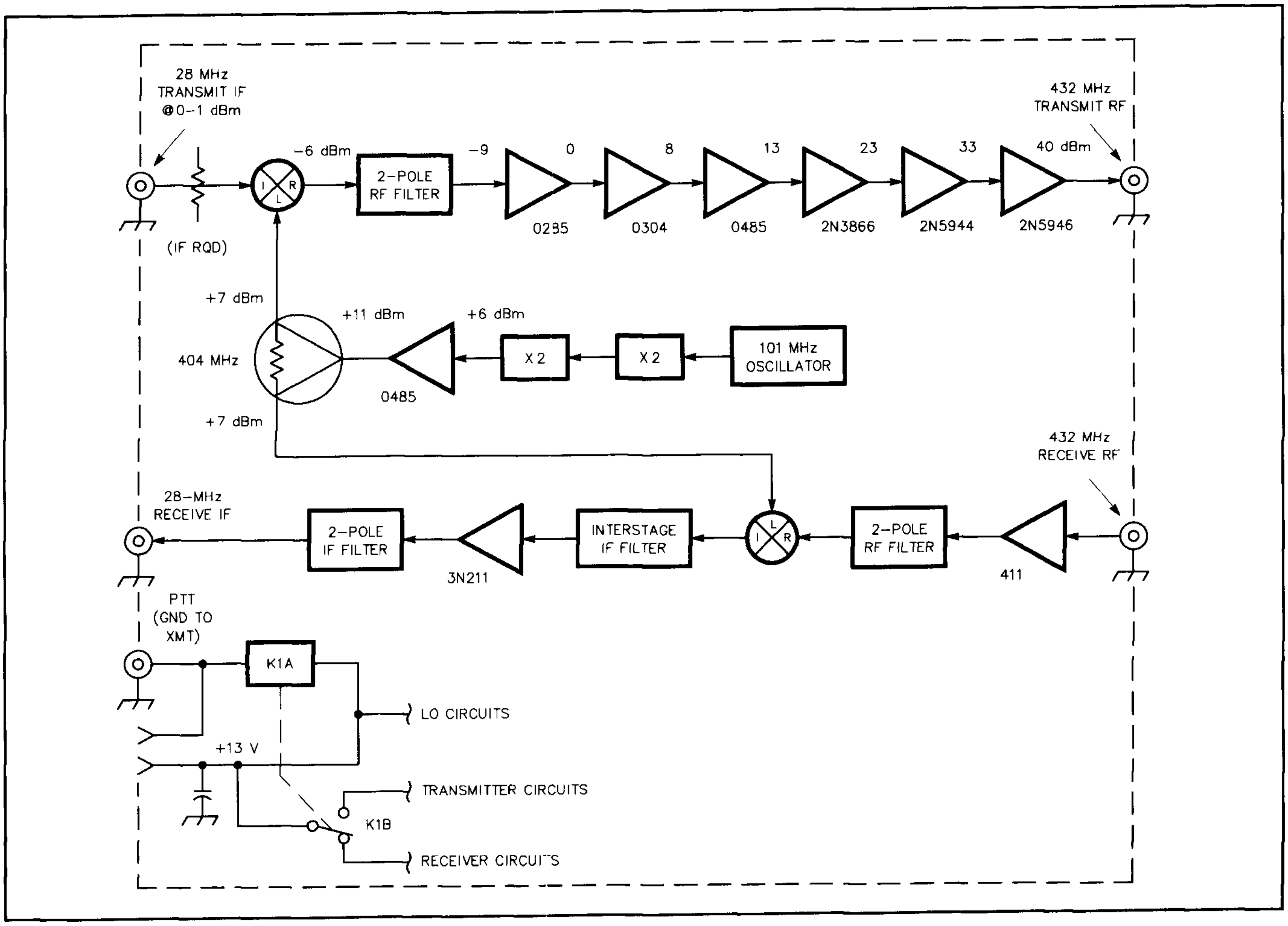


Fig 2—Block diagram of the transverter.

pacitor on the bottom side of the board as additional decoupling to prevent any HF oscillations through the supply leads.

4) All supply leads and biasing circuitry are located on the bottom side of the board.

5) Shielding was included as shown in Fig 2. The shields are placed to divide either major functions or circuits that could couple easily: high-gain, high-power or high-impedance stages, for example. The receiver RF amplifier MMIC chain and the power amplifier are thus appropriately shielded. Most of the system is at the 50-Ω impedance level.

I didn't have a metal chassis, so I tried building the chassis and shields out of PC board material. One-sixteenth-inch thick material is easily worked with a good pair of tin snips and a 60-W soldering iron. The chassis covers are secured by means of nuts soldered into the chassis corners and screws through the covers. Small pieces of scrap PC board are also handy: I used ⅜-inch squares for terminals on the bias side of the transverter. To attach a square to the board, I tinned one side of the square and the area of the board where I planned to locate it. With a soldering iron in contact with both pieces, I lowered each square into position. Be sure not to slap the squares down, as solder splashes can result.

The transverter grew around the local LO. Placing the LO between the receiver and transmitter sections seemed ideal to minimize the need for coaxial interconnects. The LO was fully tested first. A shield running the depth of the board separates the LO from the receiver section, which was built next. Once complete, the board was cut and one edge of the "chassis" was installed. A shield was added to the other side of the LO and the transmitter front end was added. A partition was added to separate the power amplifier from the transmitter low-level stages. After completing the PA the board was cutoff at the other end and that end of the chassis was installed. Small partitions of PC material were cut to match the height of the heat sink. These permit a cover to be installed over the bias side of the board, which then became the top of the chassis. The heat sink was exposed and placed on top for ventilation. Fig 3 shows the layout I used for the transverter.

Local Oscillator

A good, stable LO is critical to any UHF or microwave transverter. I adapted the 384-MHz stage of an 1152-MHz LO designed by Al Ward, WB5LUA, for my 404-MHz LO. The original circuit had the crystal in the emitter-collector circuit of the oscillator with no buffer amplifier following. This circuit drives the crystal pretty hard, and drift and "FMing" on current peaks were problems—even with the circuit supply Zener regulated to 9 V. Al was having drift problems with a 2160 LO that used the same circuit. He redesigned the oscillator using a Butler configuration and a buffer stage. That is the circuit presented in Fig 4. The parts list is in Table 1.

The oscillator, Q1, uses a 101-MHz, fifth-overtone, 0.001%-tolerance crystal. The 1-10 pF trimmer in series with the crystal can be used to adjust the LO frequency. An emitter follower, Q2,

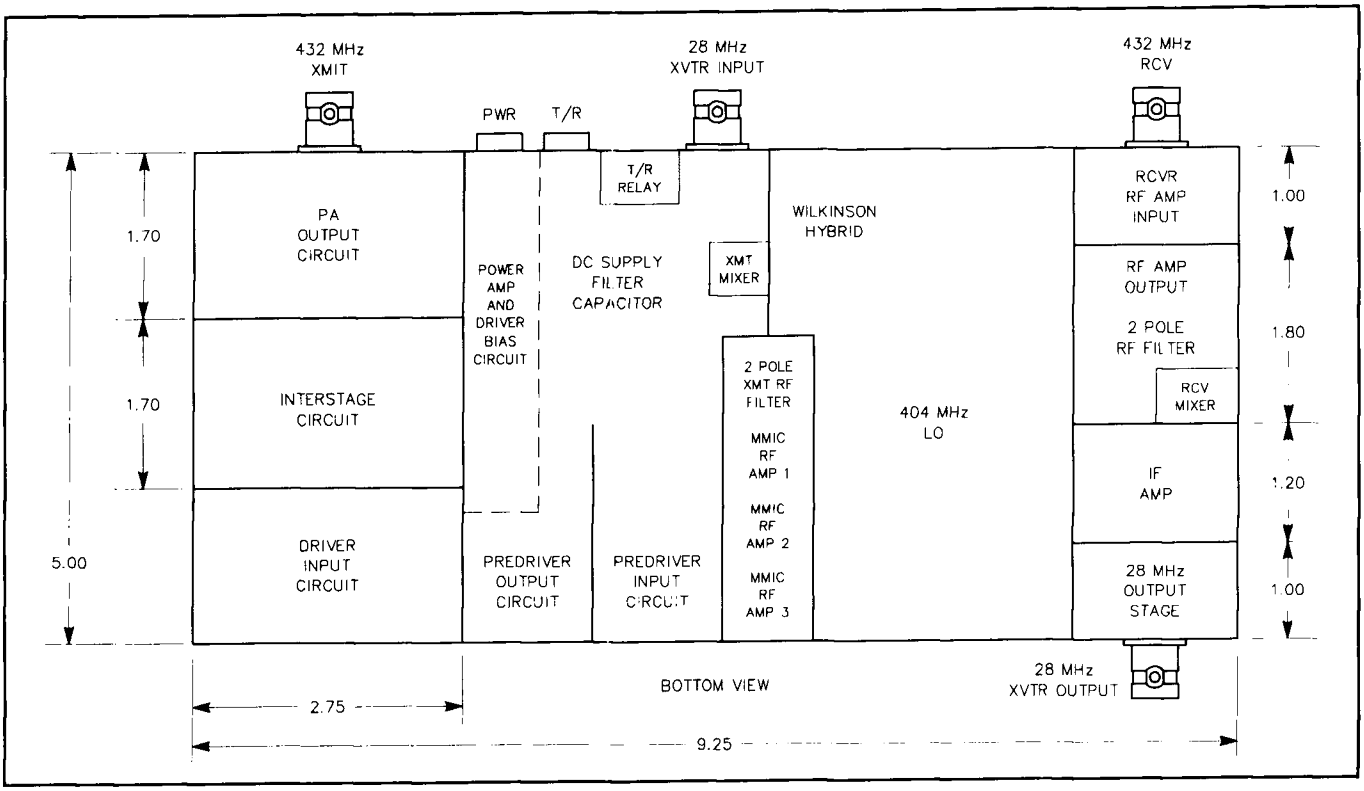


Fig 3—Layout of the transverter.

buffers the oscillator from the first doubler, Q3. The second doubler, Q4, produces the desired 404-MHz signal at +3 dBm. L6 is tapped at the 50-Ω point (a half turn) and the signal is fed into a '0485 MMIC. The '0485 amplifies the signal to about +10 dBm. A Wilkinson hybrid power splitter then produces two 50-Ω ports with +7 dBm signal levels. (This LO signal level is ideal for the SRA-1 transmit and receive mixers.) These signals are routed to the receive and transmit mixers.

The LO is best tuned with the aid of a spectrum analyzer. I was able to tune up the LO using a diode RF detector, and the results were verified on a spectrum analyzer. Access to a spectrum analyzer obviously makes the job easier.

In order to tune the LO with an RF detector, first connect the detector to the emitter of Q and tune C4 and L1 for maximum signal. The turns of L1 can be compressed or spread for adjustment. The positions of L1 and C4 affect the startup characteristics of the oscillator, and they should be positioned for consistent start up when power is applied to the circuit.

It is advisable to use a frequency counter to verify that the oscillator circuit is indeed tuned to 101 MHz before proceeding. A wide variation in circuit parameters could result in the crystal resonating on an incorrect overtone.

Move the RF detector to the collector of Q3 and tune the coupling circuit between Q2 and Q3 for maximum signal. The first doubler should be tuned for maximum signal at the emitter of Q4. Initial indications can be obtained across the 330-Ω base resistor of Q4. The 404-MHz output should be peaked by examining the signal level at each port of the power splitter, with the other port terminated in 50 Ω.

Since the final doubler circuit tuning is somewhat touchy, it may be easier to initially leave the MMIC input disconnected and terminate the output side of the 100-pF silver-mica capacitor that taps L6 in 50 Ω. Then tune for maximum signal across the resistor. Finally, tune the trimmer on the input side of the power divider for maximum signal at both output ports.

Receive Converter

The receive section of the transverter converts incoming 432-MHz RF to a 28-MHz IF. This circuit is shown in Fig 5, and the parts list is in Table 2. The front-end circuit uses an NEC 41137 GaAsFET. This circuit effectively sets the noise figure for the receiver. The 41137 is a high-gain, low-noise device, and is ideally suited for weak-signal receiver front ends. The specified noise figure of the 41137 is less than 0.8 dB at 500 MHz. I adapted a preamplifier circuit designed by Kent Britain, WA5VJB, for this application. This circuit provides a high output impedance for driving the two pole RF filter instead of a 50-Ω output for driving coax. A 1-kΩ resistor is placed inside inductor L2 to "de-Q" the drain circuit. The 41137 has about 20 dB gain and the circuit will easily break into oscillation without the resistor. A two-pole 432-MHz filter attenuates the image frequency prior to the mixer. Inductor L4 is tapped at the 50-Ω point for connection to the receive mixer. All three ports of the double balance SRA-1 have 50-Ω input or output impedances. This signal is fed to the RF port of the mixer. The LO port connects to one leg of the LO power divider.

An excellent discussion of double balanced mixers starts on page 304 of the 1976 *Radio Amateur's Handbook*. This article, by K1AGB, discusses the merits of proper termination of a DBM. One instinctively designs a mixer circuit to provide a proper termination for the desired signal. This article discusses the merits of properly terminating the DBM so that reflections from the undesired mixer products do not interfere with the desired signal. I applied the circuit included in this article to my transverter. The IF amplifier should be a good low noise device such as a 3N204 or

Fig 4—Schematic of the local oscillator.

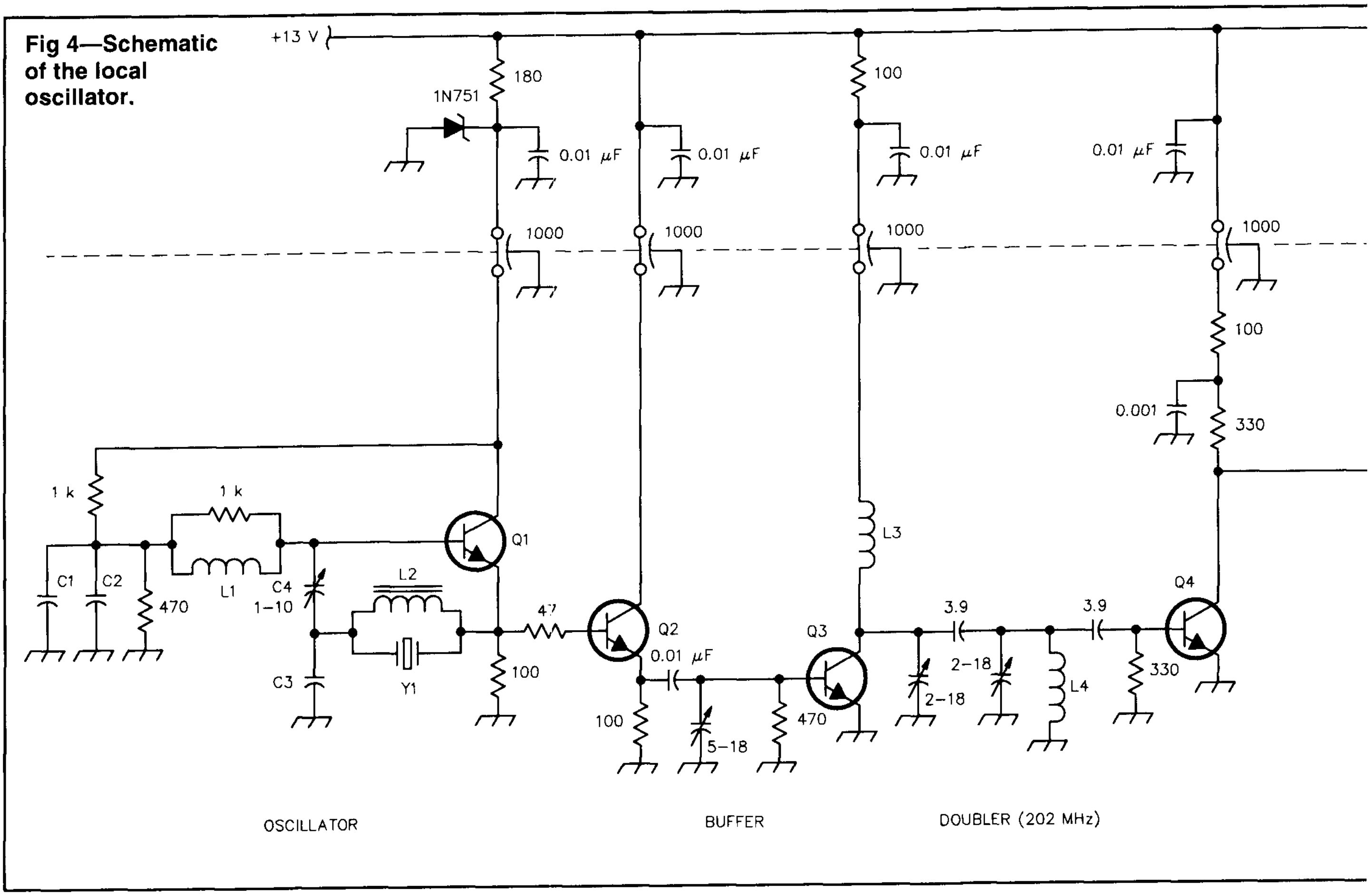
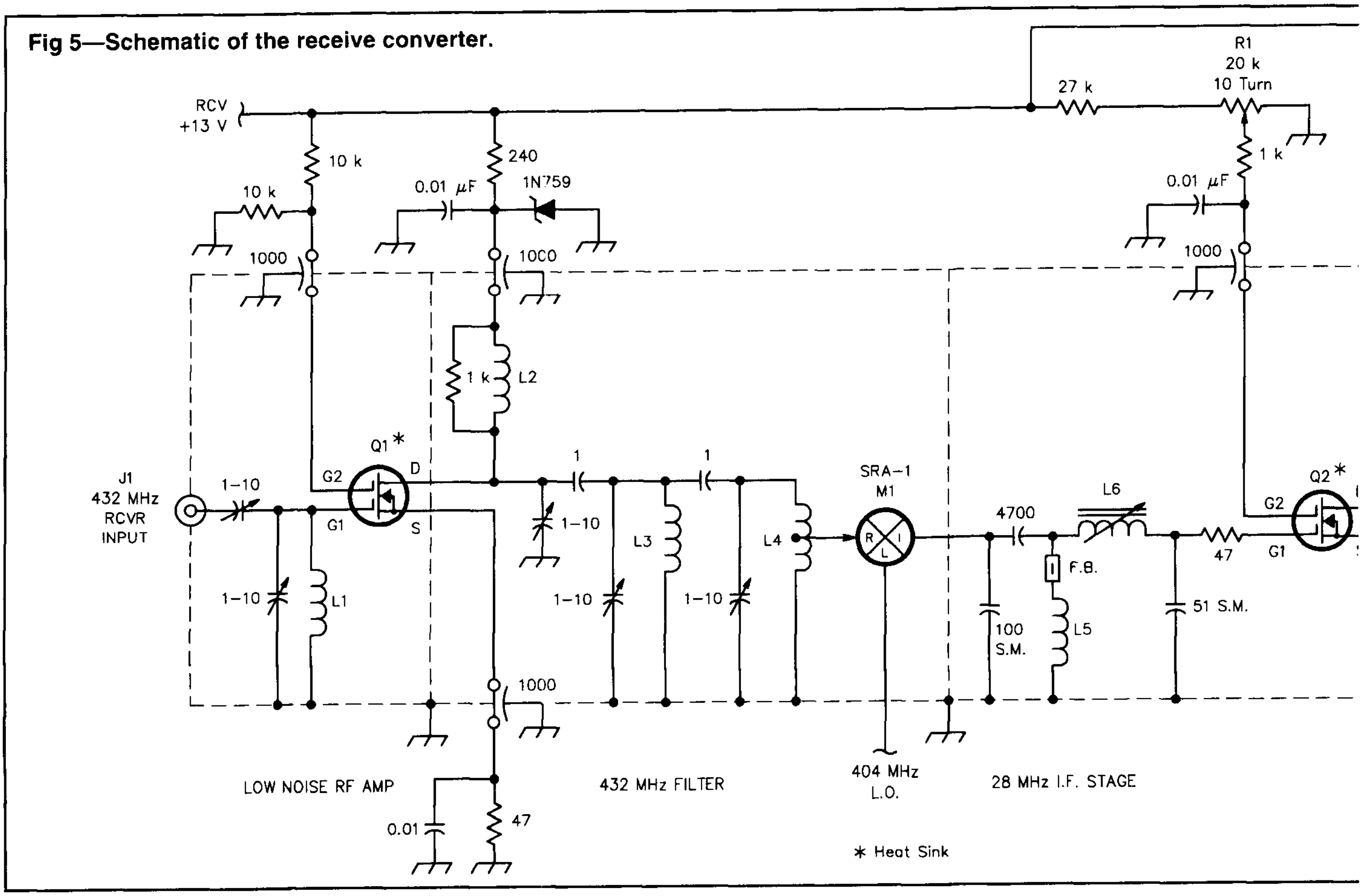
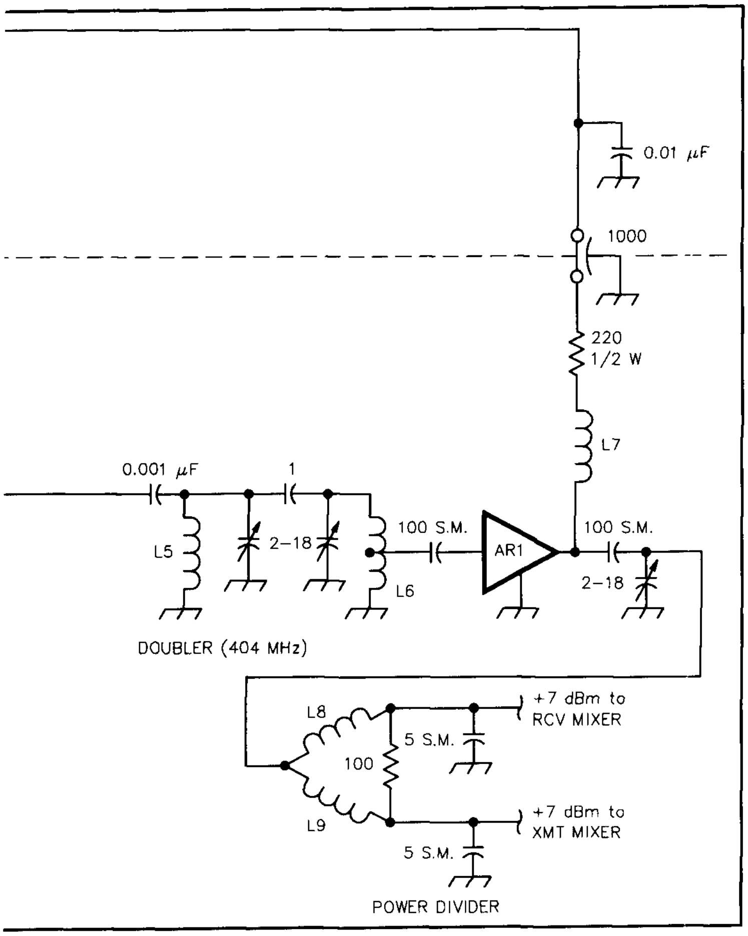


Fig 5—Schematic of the receive converter.





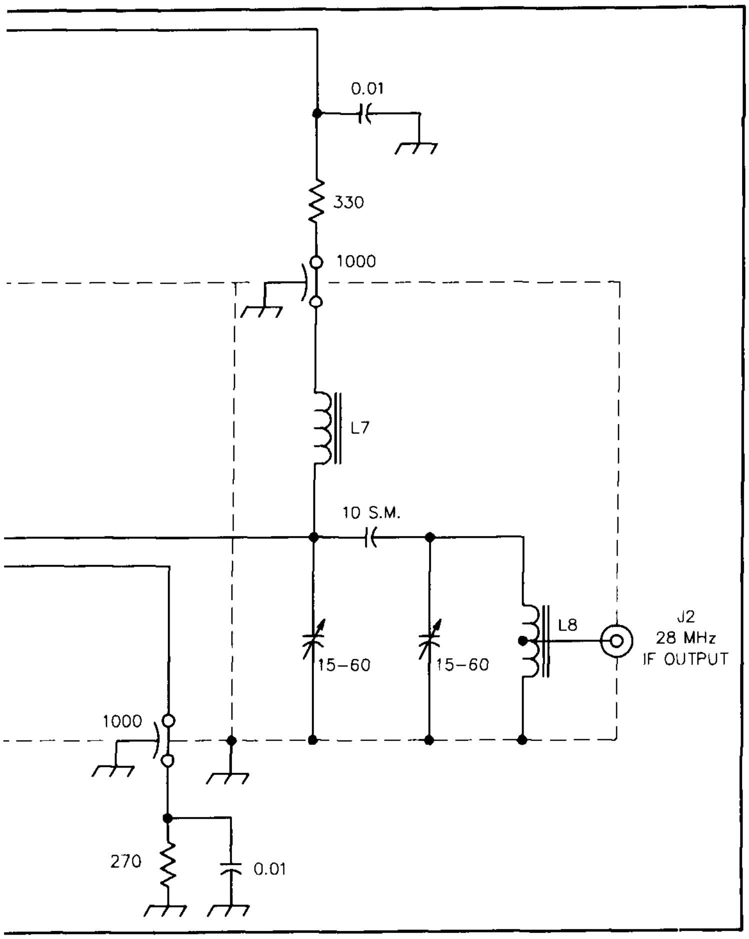


Table 1
Local Oscillator Parts

C1—0.001-μF ceramic.
C2—0.1-μF ceramic.
C3—33-pF NPO.
C4—1-10 pF trimmer.
L1—7 t no. 26, 0.25 ID, 1-kΩ resistor inside.
L2—39-μH molded choke.
L3, L4—3 t no. 16 0.25 ID.
L5—1 t no. 16, 0.25 ID.
L6—1 t no. 16, 0.25 ID, tapped ½ t.
L7—6 t no. 20 on ½-W 1 MW resistor as a form.
L8, L9—1 t no. 20, 0.25 ID.
Y1—101-MHz 5th overtone crystal, HC-18/U case.

Notes
All resistor values ¼-W unless otherwise noted.
All capacitor values in pF unless otherwise noted.

Table 2
Receive Converter Parts

L1—½ t no. 20 0.25 ID, 0.5 in. long.
L2—7 t no. 20, 0.25 ID, 1-kΩ resistor inside.
L3—1 t no. 20 20, 0.25 ID.
L4—1 t no. 20, 0.25 ID, tapped ½ t.
L5—56-μH molded choke.
L6—19 t no. 26 on 0.25 ID coil form, per slug.
L7—18 t no. 28 on T-30-6 core.
L8—18 t no. 28 on T-30-6 core tapped 4 t from cold end.
M1—SRA-1 Minicircuits DBM.
Q1—NEC 41137 GaAsFET.
Q2—3N204, 3N211.

Notes
All resistors ¼ W unless otherwise noted.
All capacitor values in pF unless otherwise noted.

3N211. A 40673 could be used but is not recommended. The drain circuit of the IF amplifier is connected to a doubly tuned output circuit. A dual-gate FET is used to provide stage gain adjustment. The adjustment is effected by using potentiometer R1 to control the voltage on G2 of the FET. A preamplifier or transverter will inject additional noise into the front end of the rig being used either as a baseband rig or an IF. The result is an S-meter reading higher than if the gain stage preceding the rig was turned off. R1 permits setting the S-meter reading with the transverter turned on to the same value as if the transverter were turned off. S-meter readings can thus be exchanged. Good weak-signal reporting practice should still include reports of signal strength over the noise floor.

A two-pole 28-MHz filter follows the IF amplifier. I used the same inductors that were used in the "Rochester converters" shown in the 1976 *Radio Amateurs Handbook*. In that application, the circuit was stagger tuned for broad bandwidth and no tuning adjustments, other than taking turns off the toroids, were provided for. I replaced the fixed capacitors with trimmers. Since most activity is at 432.100 MHz, I tuned both sections of the filter for the corresponding IF frequency of 28.100 MHz. Tune up of the

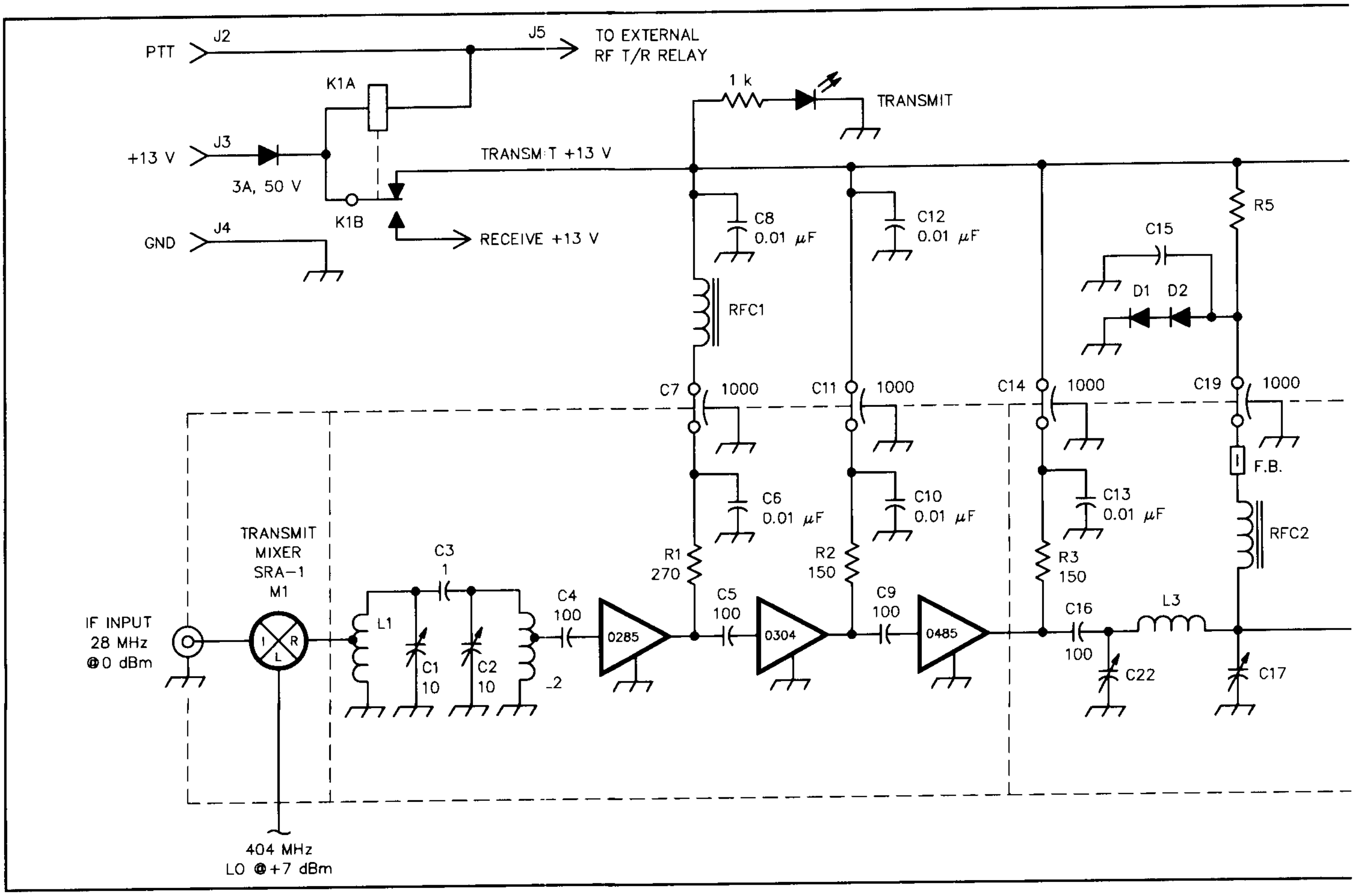


Fig 6—Schematic of the transmitter front end.

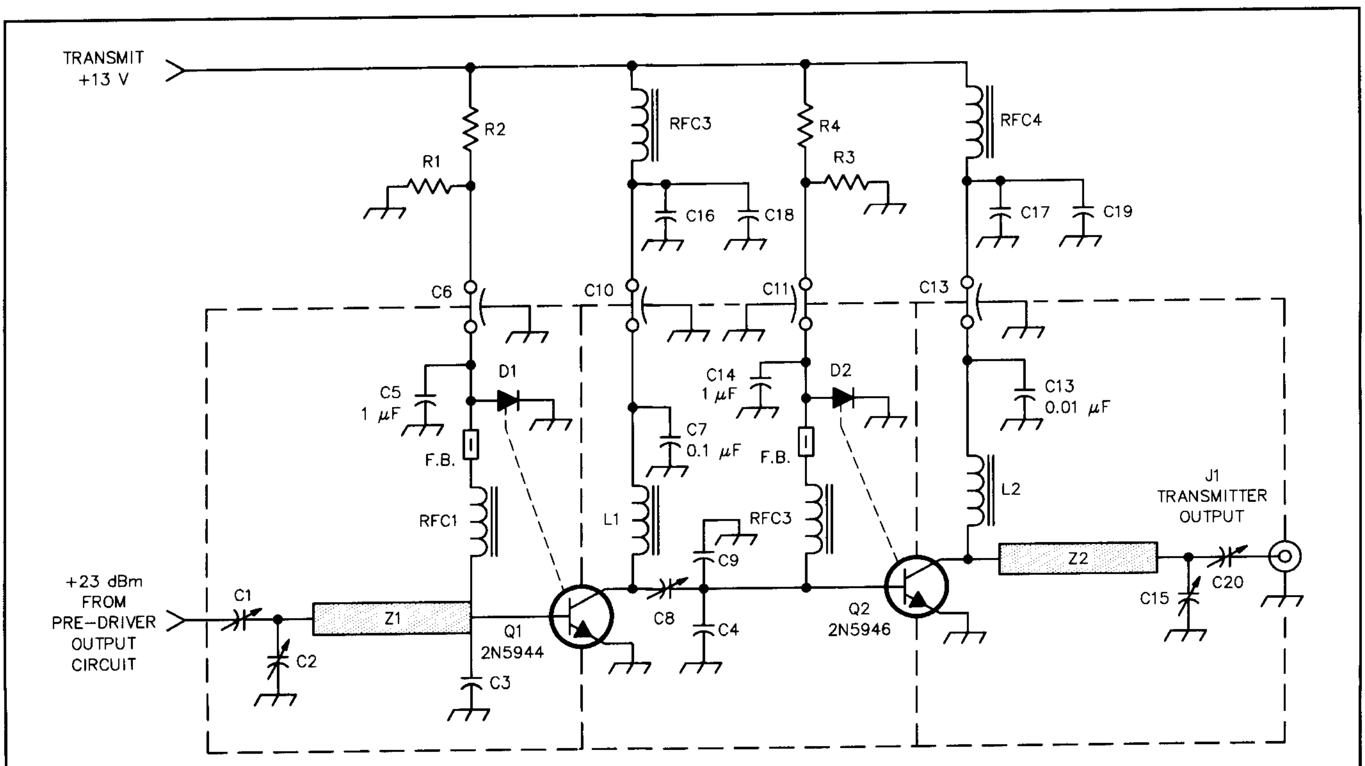


Fig 7—Schematic of the transmitter power amplifier.

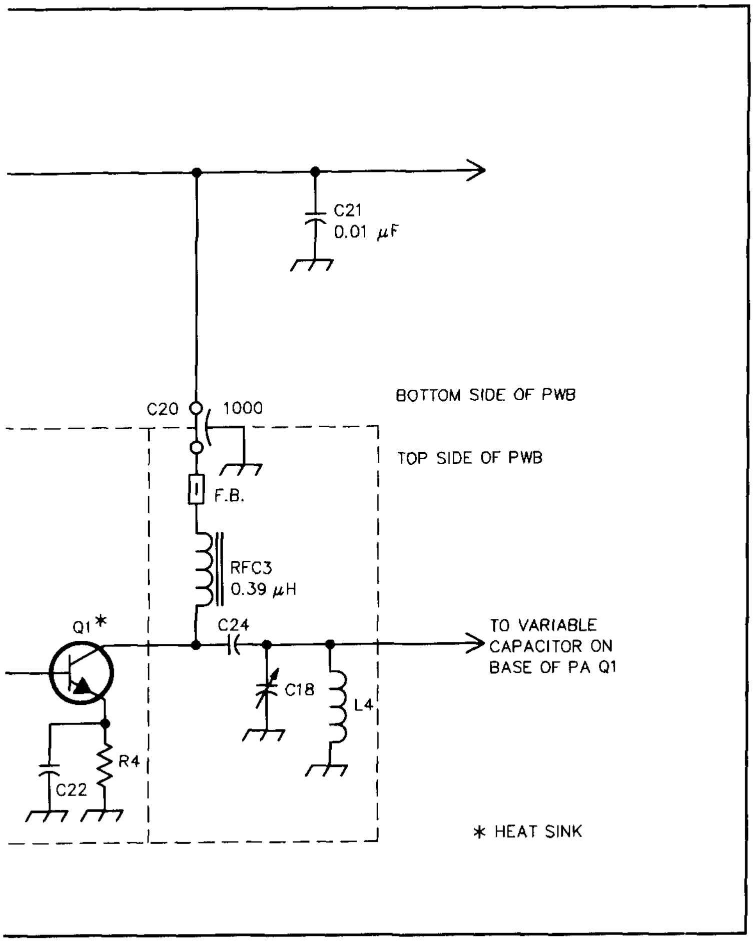


Table 3

Transmitter Front End Parts

Q1—MMIC, 0285 or equiv.
Q2—MMIC, 0304 or equiv.
Q3—MMIC, 0485 or equiv.
C1, C2, C22—miniature trimmer, 1-10 pF.
C3—Ceramic, 1 pF.
C4, C5, C9, C16—silver mica, 100 pF.
C6, C8, C10, C12, C13, C15, C21, C23—ceramic, 0.01 μF.
C7, C11, C14, C19, C20—feedthrough, 1000 pF.
C17, C18—miniature trimmer, 5-25 pF.
C24—0.01-μF ceramic.
D1, D2—silicon diode, 50 PIV, 1N4001 or equiv.
K1—DPDT relay, 5-A contacts, 12-V dc coil.
L1, L2—1 t no. 20 enam, ¼-in. diam, tapped at ½ t.
L3—2t no. 16 enam, ¼-in. diam.
L4—1 t no. 20 enam, ¼-in. diam.
Q1—2N3866 or equiv.
R1—270-Ω, ¼-W carbon composition.
R2, R3—150-Ω, ½-W carbon composition.
R4—15-Ω, ½-W carbon composition.
R5—270-Ω, ½-W carbon composition.
RFC1—molded choke, 1 μH.
RFC2, RFC3—molded choke, 0.39 μH.

circuit can be done several ways. A 432-MHz signal source with variable attenuation is necessary. The rig being used as the IF serves well as the detector. With the LO connected to the mixer, apply a strong 432-MHz signal (−30 dBm) to the 432-MHz input. Make several passes though the stages, tuning for maximum indication on the S-meter. Decrease the signal level as the circuits come into alignment, so that the S-meter reading stays below S9. You should be able to reduce the signal level to less than −120 dBm and still hear a beat note. The circuit could be split if tuning the full receiver proves difficult.

In this case, break the circuit at the input to the mixer and terminate the connection from L4 to the mixer with a 50-Ω resistor. Then, tune the 41137 circuit and the filter for maximum signal strength. Keep the input signal level low enough that the 41137 is not running into compression; for example, below −30 dBm. On the output side of the mixer, drive the circuit connected to the IF port of the mixer with a 28-MHz signal and tune L6 and the two 15-60 pF trimmers for maximum signal. The radio can still be used as the detector for this section of the circuit. A good initial value for R1 should result in about 1.5 volts on G2 of the IF amp. Finally, connect everything back up and make final adjustments for best overall signal conversion gain. Kent Britain discusses tune up of preamplifiers in the 1986 *Handbook for Radio Amateurs*. As noted there, adjustment of the two trimmers on the input of the 41137 are not necessarily the same for best gain or best noise figure. After the receiver section is tuned up enough to be connected to an antenna, find a friend who can transmit a weak signal and tune up on it for best intelligible signal. I had the opportunity to put the transverter on a Hewlett-Packard 8970 noise-figure meter. The overall receiver noise figure for the transverter was measured to be 1.60 dB. Not perfect, but very acceptable. I have assumed 13 dB gain in the 41137 stage. Although more than 20 dB of gain is available with the 41137, the damping to stabilize the stage probably results in about 13 dB of stage gain.

Transmit Converter

This portion of the transverter converts a 1-mW 28-MHz IF input to a +40-dBm (10-watt) 432-MHz output. Refer to the transverter block diagram for signal level and flow. These are described below. The transmitter front end schematic is shown in Fig 6 and the transmitter power amplifier in Fig 7. Associated parts are listed in Tables 3 and 4 respectively. The 28-MHz IF signal is input directly into the IF port of the SRA-1 mixer. This is the specified input level. Attenuators or amplifiers may be necessary if the 28-MHz signal level available is more than a few dB different from 0 dBm. 404-MHz LO energy from the power splitter is input to the LO port of the SRA-1.

The mixer output is taken from the RF port. The desired 432-MHz signal level is about −6 dBm at this point. A two-pole 432-MHz filter follows the mixer. This passes the desired 432-MHz signal and attenuates the 376 MHz lower mixer product. A DBM is useful here, as the 404-MHz and 28-MHz signals are greatly attenuated, due to the inherent nature of a DBM. Filter loss was measured at 3 dB and the resulting signal at the output of the filter is −9 dBm. The −9 dBm signal is input to a series of three MMICs for amplification. The upper limit of signal level that commonly available silicon MMICs can work with is about +13 dBm, for the 04 series of parts.

The data sheet specifications are 12 dB for the 0285 and 0304 MMICs, and 8 dB for the 0485 device. Overall three stage gain would be 32 dB. I got 22 dB overall. I don't find this disturbing, as the specification sheet values are based on tests where the devices

Table 4

Power Amplifier Parts

C1, C2—compression trimmer, 4-20 pF; ARCO 402 or equiv.
C3—UNELCO, 10 pF.
C4—UNELCO, 25 pF.
C5, C14—ceramic, 1 μF.
C6, C10, C11, C12—feedthrough, 1000 pF.
C7, C13—ceramic, 0.1 μF.
C8—compression trimmer, 55-250 pF; ARCO 426 or equiv.
C15—Johanson trimmer, 1-10 pF.
C16, C17—dipped tantalum, 1 μF.
C18, C19—aluminum electrolytic, 47 μF, 25 V.
C20—compression trimmer, 8-45 pF, ARCO 403 or equiv.
D1, D2—silicon diode, 50 PIV, 1N4001 or equiv.
J1—BNC connector, chassis mount.
L1—6 t no. 20 enam, ¼-in. diam.
L2—5 t no. 20, ¼-in. diam.
R1, R3—10-Ω, ½-W carbon composition.
R2, R4—150-Ω, 1-W carbon composition.
RFC1—molded choke, 0.1 μH.
RFC2—molded choke, 0.39 μH.
RFC3, RFC4— 2½ t no. 20 ferrite core.
Z1—0.95 × 0.25 microstrip; see text.
Z2—see text.

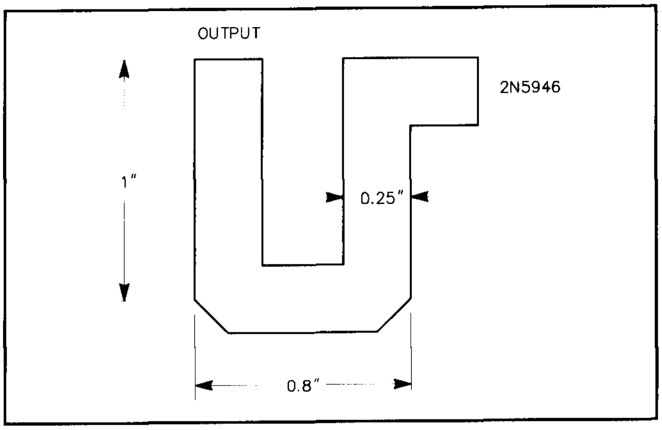


Fig 8—This copper microstrip is used in the 2N5946 output circuit.

are well matched. Mismatch losses between stages will decrease the available gain. The +13 dBm output of the 0485 is input to a 2N3866 transistor. This stage produces 10-dB gain and outputs the signal at +23 dBm to the next stage. L networks are used for input and output impedance matching. Both input and output circuits are impedance matched to 50 Ω. This permits testing of the stage independently, if desired. The quiescent collector current is set up by controlling the current into D1. The quiescent collector current is set to 6 mA, about 5% of the expected maximum collector current. This puts the device into the linear AB mode of operation. Next, a 2N5944 amplifies the +23 dBm level to +33 dBm (2 watts). This is followed by a 2N5946, which provides up to 10 watts output. Several circuits in the Motorola *Power Devices* catalog served as the model for the circuits shown here. A capacitive L and a 0.25-inch wide by 0.9-inch long section of microstrip is used to match the 50-Ω input impedance to the base impedance of the 2N5944. A "U" shaped section of microstrip shown in Fig 8, and a mirrored capacitive L transform the low collector impedance of the 2N5946 back to 50 Ω. This section of microstrip was shaped this way to conserve board space. There was enough inductance in the ARCO trimmer capacitor in the interstage circuit that a microstrip section was not necessary. The biasing technique for the driver and final is similar to that used for the 64020. The maximum collector current for the 2N5944 is 400 mA and the quiescent collector current was set to 20 mA. Maximum collector current for the 2N5946 is 2 A and the quiescent current was adjusted to 100 mA.

This current adjustment is again made by varying the current into a forward-biased diode across the base. It is important to thermally couple the bias diode to the transistor for these higher power devices. I put the bias diode directly across the lid of the power transistors and applied heat-sink grease, to effect a close thermal coupling. Experimentation may be required in the decoupling circuits to achieve stable operation. I initially started out with the decoupling circuits shown in the Motorola catalog. The end result was somewhat different. Overall, this portion of the transverter has required the most trial and error. Stabilization was achieved by varying the placement of the chokes and ferrite beads and by changing choke values. The impedance should be low. I used values in the 200- to 500-Ω range. Larger values may permit VHF oscillations in the stage. Chokes of 0.1 and 0.18 μH worked well.

I have achieved 7 watts output. Tests indicate that the driver is loaded up to the full 400 mA collector current. Key-down collector current of the 2N5946 measured 1.2 A. This indicates that the match to the input of the 2N5946 isn't perfect. This is an area for improvement. Testing the transmitter section requires some means of measuring 432-MHz energy from about −10 dBm to +10 watts. A calibrated diode detector gave me satisfactory results for the lower-level stages. A microwave power meter, such as an HP 430, 431 or 432 is better. If you do use one of these instruments, always put enough attenuation in front of the thermistor to prevent burning out the thermocouple if the stage breaks into oscillation. (I learned the hard way.)

A Bird 43 or other suitable wattmeter is needed to measure the output power of the final stage. Since the interstage impedances are mostly 50 Ω, the transmitter can be brought up as it is constructed. That is, you can measure power at the output of the first second and third MMICs, the 2N3866 and the 2N5946. The interstage circuit between the 2N5944 and 2N5946 is not at 50 Ω so these two stages must be tuned at the same time. A quick check to see if the MMIC is working is to measure the voltage on the output lead of the device. It should be about 5 volts for the 0285 and 0304, and about 6 volts for the 0485. Bias current is a better check. The bias current for the 0285 should be about 30 mA, with 40 mA for the 0304 and 50 mA for the 0485. Proper operation of the 2N5944 and 2N5946 can be determined by measuring the collector current in a key-down condition. It should be close to 400 mA for the 2N5944 and 2 A for the 2N5946.

A Note On The MMICs

The last two digits of the MMIC part numbers denote the package type. A 0404 will perform essentially the same as a 0485. These part numbers are Avantek numbers. Mini-Circuits equivalent parts are MAR-1, MAR-2, MAR-3 and MAR-4 (corresponding to the Avantek 0185, 0285, 0385 and 0485 devices.)

A Single-Board, No-Tune 902-MHz Transverter

By Rick Campbell, KK7B

(This article, which originally appeared in *Microwave Update '89*, is reprinted from *QST*, July 1991.)

The sole obstacle many amateurs face to getting on the UHF and microwave bands is a lack of equipment. It's not that the equipment and antennas for these bands are expensive or unobtainable; it's more that many would-be microwavers feel intimidated by the prospect of building their own gear for these bands. If you have a 2-meter multimode transceiver, you're already well along in getting on the 902-MHz and higher bands. The complete 902-MHz transverter I'll describe here is printed on a single 5- × 7-inch G10 circuit board and costs less than $150 to build. You'll soon see that you don't need lots of money or much building skill to put a 33-cm station on the air!

This transverter, unlike similar designs for higher-frequency bands,[1-4] has its local oscillator (LO) on the main PC board; only a 12-V supply, 144- and 902-MHz signals enter and exit the board. All the gain stages employ low-cost plastic MMIC (monolithic microwave integrated circuit) amplifiers. All the band-pass filters, traditionally the trickiest part of UHF-equipment design, are hairpin, third-order Chebyshev types printed on the PC board. The filters require no adjustments, like those of the higher-frequency transverters. Kits of parts and assembled transverters are available from Down East Microwave.[5]

Introduction

This transverter, shown in block-diagram form in Fig 1,

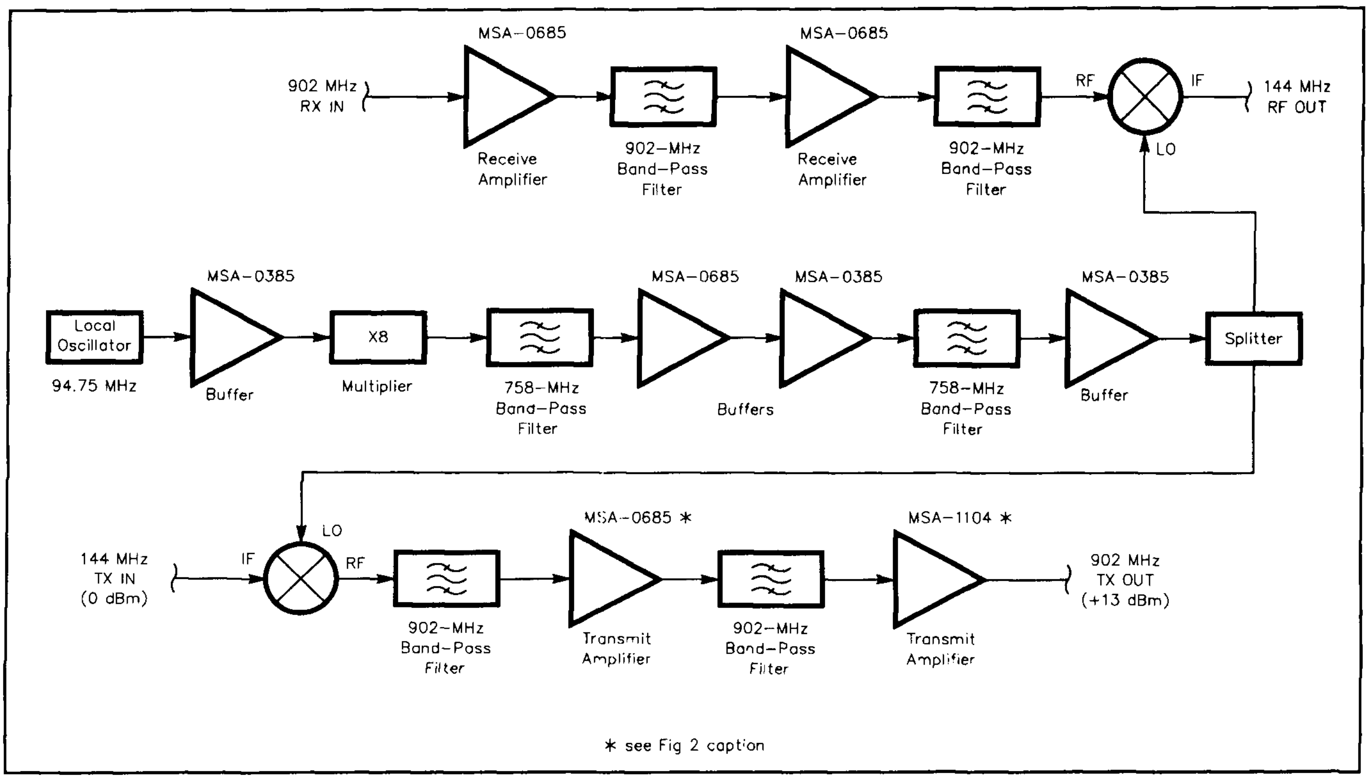
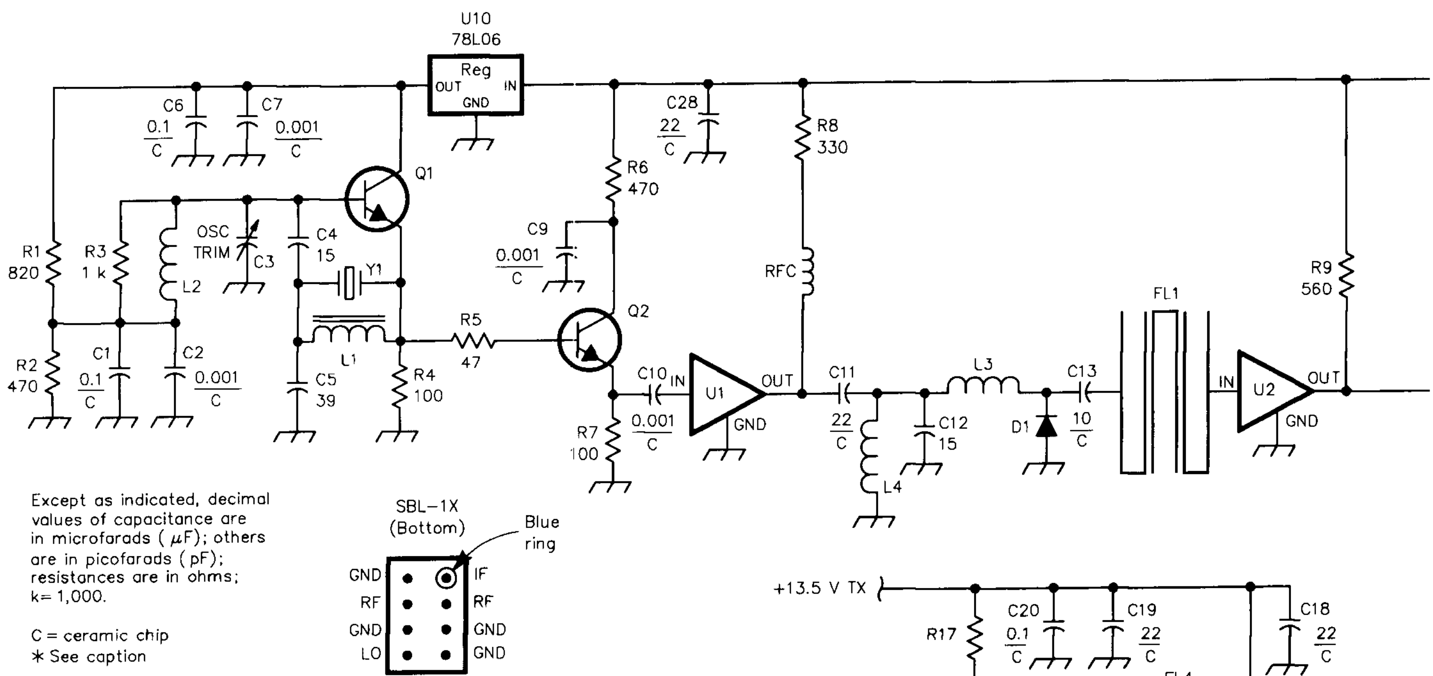


Fig 1—Block diagram of the single-board 902-MHz transverter. Unlike similar designs for the higher bands, this unit has an on-board local oscillator and discrete mixers. The only external connections are for the IF transceiver, antenna and power supply.



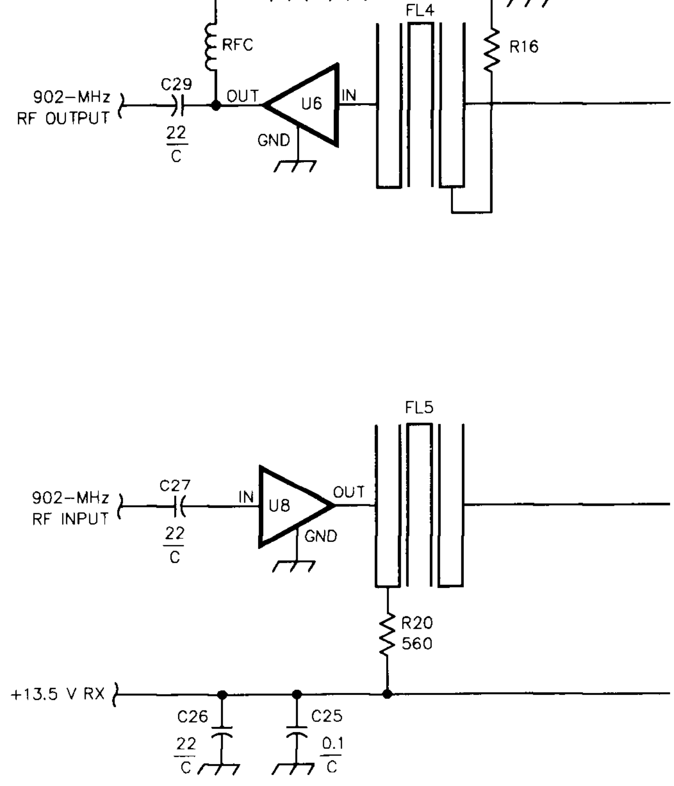
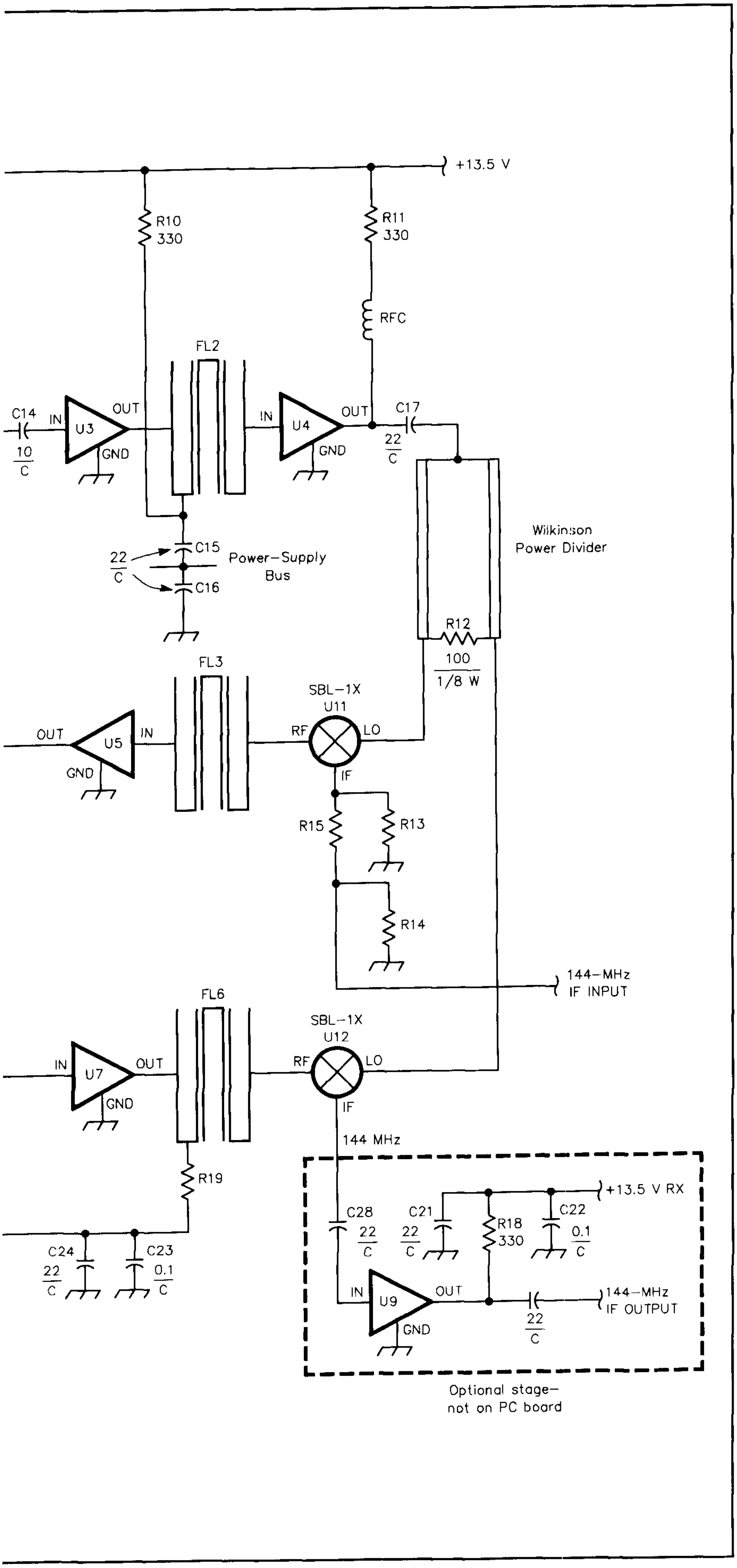


Fig 2—Schematic of the 902-MHz transverter. Heed the manufacturer's specifications for MMIC connections. The boxed part of the circuit isn't on the PC board, but may be needed to drive your 144-MHz IF receiver. A pi-network attenuator can be used here, if appropriate (see text and Fig 3). Filters FL1-FL6 are etched band-pass units. All resistors are ¼-W carbon-composition units unless specified otherwise. See the parts list for differences between 0-dBm-output and 13-dBm-output versions.

C3—3- to 10-pF ceramic or piston trimmer.
C4, C12—15 pF, miniature ceramic, NP0.
C5—39 pF, miniature ceramic, NP0.
C11, C14-19, C21, C24, C26-29—22 pF, ceramic chip, NP0.
C13—10 pF, ceramic chip, NP0.
D1—Multiplier diode; HP2800, HP2835 or 1N5711.
L1—10 turns no. 28 enam wire on T-25-6 toroid core.
L2—7 turns no. 24 enam wire, 0.1 inch diam, closewound.
L3, L4—8 turns no. 24 enam wire, 0.1 inch diam, closewound.
Q1, Q2—AT42085, NE021, BFX89 or equiv.
RFC—8 turns no. 24 enam wire, 0.1-in. diam, closewound.
R13, R14—150 Ω, ¼ W.
R15—33 Ω, ¼ W.
R16—360 Ω for low-power version; 560 Ω for high-power version.
R17—½ W; 330 Ω for low-power version; 130 Ω for high-power version.
U1—Avantek MSA-0385 or Mini-Circuits MAR-3.
U2, U7, U8—Avantek MSA-0685 or Mini-Circuits MAR-6.
U3, U4—Avantek MSA-0385 or Mini-Circuits MAR-3.
U5—Avantek MSA-0685 or Mini-Circuits MAR-6 for high-power version; MSA-0285 or MAR-2 for low-power version.
U6—Avantek MSA-0385 or Mini-Circuits MAR-3 for low-power version; MSA-1104 for high-power version.
Y1—94.75-MHz crystal. See text.



was a natural development following my 1.3-GHz transverter[6] and 576-MHz local-oscillator (LO) board.[7] The signal filters were scaled to pass 900-940 MHz and the LO filters were scaled to pass 740-780 MHz. The LO is a Butler type operating in the 100-MHz range, followed by a broad-tuned Schottky-diode multiplier and printed bandpass filter to reject all but the desired 758-MHz harmonic.

The 1.3-GHz transverter uses etched quadrature hybrid mixers. Scaled for the 758-MHz LO needed to convert a 144-MHz IF signal to 902 MHz, these mixers are too large to fit on a reasonably small PC board. A number of reasonably priced, discrete double-balanced mixers (DBMs) are available for frequencies below 1 GHz, so a pair of such mixers are used in this transverter. Prepackaged discrete DBMs need far less board space than etched mixers.

The transverter prototypes were built using Mini-Circuits SRA-11H and SRA-2CM mixers. The second-generation design shown here uses less-expensive Mini-Circuits SBL-1X mixers. Fig 2 shows the schematic diagram.

The Local Oscillator

Using prepackaged mixers freed up enough circuit-board space for an on-board local oscillator. The LO circuit is nearly identical to the 540- to 580-MHz version described in July 1989 *QST*.[8] For 758-MHz output, the Butler oscillator's frequency can be 94.75 MHz (x8), 108.286 MHz (x7) or 126.333 MHz (x6). As it turns out, 94.75 MHz is the best choice, as the oscillator is easiest to adjust and the undesired oscillator harmonics at 853 and 948 MHz are farther (and therefore easier to filter) from the RF-amplifier and transmit-amplifier passbands. The prototypes used 126.333-MHz seventh-overtone crystals. The first prototype built at KK7B uses SRA-11H mixers, which are specified for +17 dBm LO injection. A few decibels more output from the harmonic generator is obtained by biasing the diode as described in "A Clean Microwave Local Oscillator."[9] An 820-Ω bias resistor was used for this.

With an MSA-1104 LO amplifier and careful tweaking of the inductors in the harmonic generator, +18.5 dBm was available before the Wilkinson power divider that splits the LO for the receive and transmit mixers. This is sufficient for good performance with SRA-11H mixers. For the low-level SRA-2CM and SBL-1X mixers, +10 dBm output from the LO is optimum, which is easily obtained with the parts indicated in Fig 2.

All spurious LO outputs below 1.3 GHz are more than 40 dB below the desired output, but the LO filters have some spurious responses in the 1.4-GHz range. The spurs in this range may be stronger, depending on drive level to the MMICs

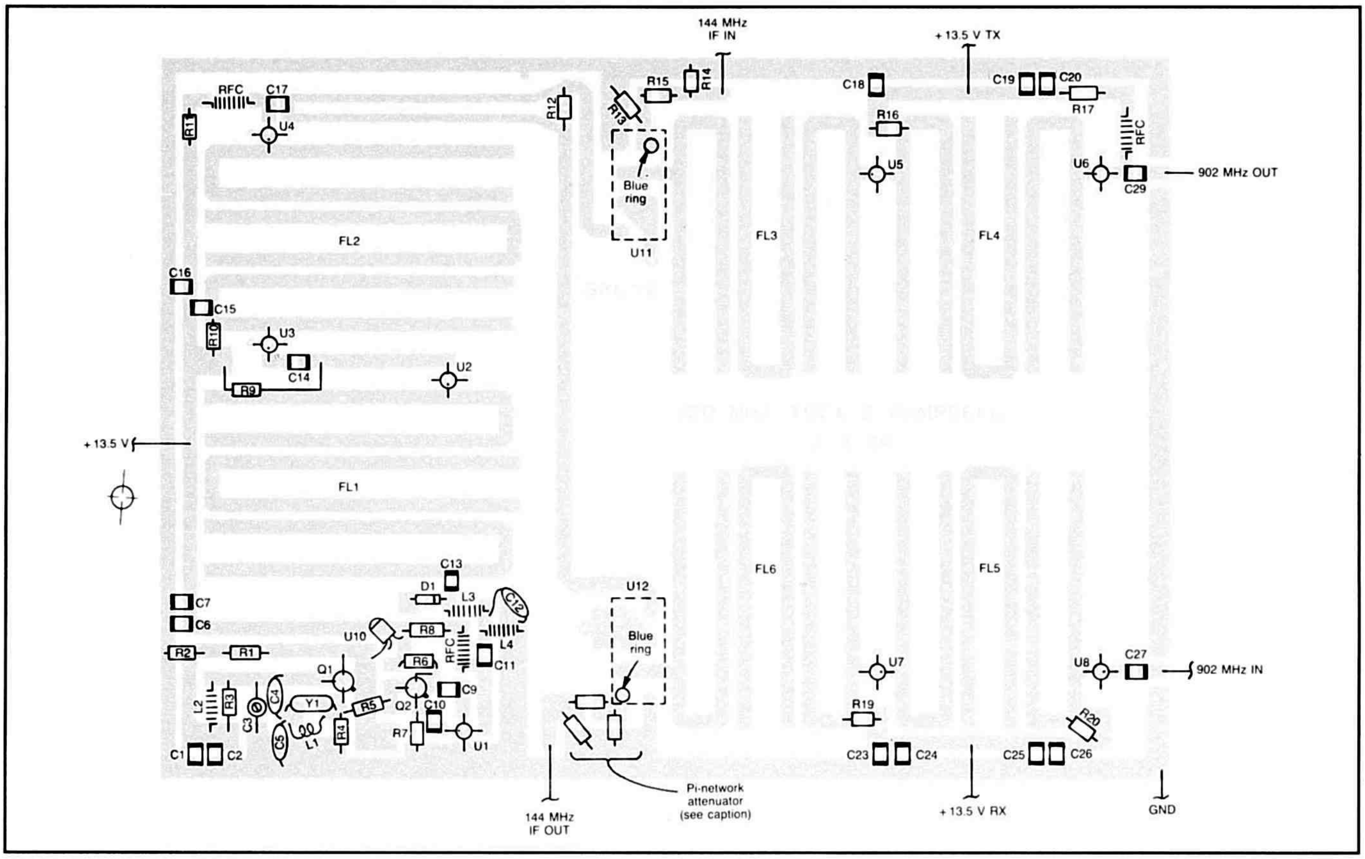


Fig 3—Part-placement diagram for the transverter. Although either MCL or Avantek MMICs are acceptable for use in this project, the MMICs shown here are marked like MCL parts (the colored dot signifies the input lead).

in the LO chain. The worst-case spurs in any of the prototype LOs were 20 dB below the desired output. Adding a low-pass filter after the LO would reduce the high-frequency LO spurs; I didn't do this, because these low-level, high-frequency spurs aren't visible in the RF-output spectrum and don't degrade receiver performance.

RF Amplifiers

The transmit and receive amplifiers are similar to those in the 1.3-GHz transverter, except that the filters are somewhat closer to being ideal at 902 MHz, so only two MMICs per chain are needed to obtain the required gain. MSA-0685s are used for both receive stages, providing a noise figure (NF) under 4 dB, more than 20 dB of conversion gain (including filter and mixer loss), and a stable 50-Ω termination for an external GaAsFET preamplifier. The first prototypes used an MSA-0685 IF amplifier after the receiver mixer, but total gain was excessive.

The SBL-1X version uses a 3-dB resistive pad at the receiver IF port to keep the receive-converter output to an appropriate level. The transmit amplifier provides 13 dBm (20 mW) output at the 1-dB compression point. This is appropriate for driving a discrete amplifier chain, or for direct connection to the antenna for local, line-of-sight or hilltop operation. Most hybrid linear-amplifier modules suitable for 33-cm use, such as the MC5874 and PF0011,[10] require a lower drive level. Fig 2 and the parts list specify components for versions with 13 dBm (20 mW) output and with 0 dBm (1 mW) output to best suit your requirements.

Construction, Tuning and Operation

Because of the tight dimensional tolerances required for etched microwave filters that require no adjustments, and because there are many variables in the printing process, a PC-board etching pattern is not included with this article. If you want to make a board for your own use, send an SASE to the ARRL Technical Department Secretary for a dimensioned copy of the artwork.[11] PC boards, parts and kits are available, however, as mentioned earlier.[12] Follow the construction guidelines discussed in Jim Davey's 1989 article,[13] and use only high-quality, microwave-rated porcelain chip capacitors in building the circuit. Avantek and Mini-Circuits (MCL) make MMICs suitable for use in this project. See the parts list in the Fig 2 caption for equivalent parts. There is some variation in the packaging of MMICs. Some use a colored dot at the bevel-cut input lead; others use a raised dot at the output lead. Be sure to install the MMICs in the correct orientation. Fig 3 shows placement of parts on the transverter PC board. The 144-MHz IF output comes off the board via a pad adjacent to U12. If necessary, you can use the traces in that area for a pi-network attenuator, or as a take-off point for the receiver signal or for a subsequent amplifier stage, such as the boxed section in the lower right corner of Fig 2.

Once the PC board is populated, the only adjustment this

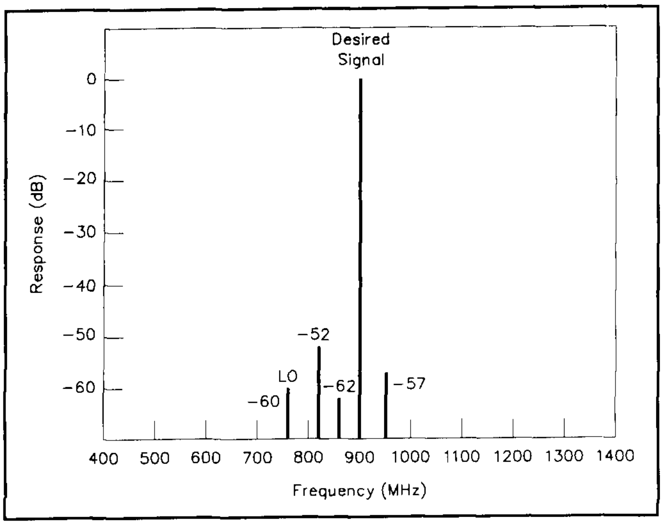


Fig 4—The transverter's spectral output. Vertical divisions are 10 dB; horizontal divisions are 100 MHz. All spurious outputs are at least 50 dB down from the carrier (–50 dBc). This plot shows that it doesn't take expensive or complex circuitry to generate clean signals—even above 900 MHz.

transverter requires is tweaking of its LO trimmer (C3) to ensure reliable oscillator starting. To do this, apply 13.5 V to the LO. If you can, observe the LO signal at 758 MHz and adjust the trimmer until the oscillator restarts every time power is removed and reapplied. If you like, you can also use either the 902-MHz transmit or receive section to verify this. The 94.75-MHz LO is also audible on standard FM-broadcast receivers. In operation, apply 13.5 V to the LO and the transmit or receive section, depending on which is in use. (It's good practice to remove power from the unused stage.)

Performance

When driven with 1 mW of 144-MHz RF, this transverter provides a clean, low-power 902-MHz signal. All spurious outputs are more than 50 dB below the desired output, as shown in Fig 4. The transmit-converter output is suitable for direct connection to a linear amplifier without additional filtering.

On receive, the transverter's under-4-dB noise figure and unconditionally stable 50-Ω input termination are hard to beat. Image rejection is more than 70 dB. No input filtering is done before the first amplifier stage, so the input stage is susceptible to overload in high-RF environments. For use in such environments, there are several options:

- Replace the receive amplifiers with MSA-1104s. This increases the noise figure by about 1 dB and increases dynamic range by about 10 dB. This is usually not enough of an improvement to cure overload, though.
- Omit the first amplifier stage. This increases the noise figure to 7 or 8 dB, but will probably cure the problem.
- The best alternative: Use an external, low-loss filter. If this or any other transverter is to be used around other transmitters, *it is good practice* to use a low-loss cavity filter before the first receive-amplifier stage.

Conclusions

When used with a suitable outboard linear amplifier and a GaAsFET preamp, this transverter easily outperforms older designs—at a fraction of the cost. The performance advantages gained by the use of printed band-pass filters, combined with the elimination of all microwave adjustments and the need for a spectrum analyzer, are significant advances in the state of the art that are common to this family of microwave transverters.When this transverter is coupled with the right antenna system, you can work tremendous DX on the 902-MHz band. Dave (WA3JUF) Mascaro's *QST* article, "A High-Performance UHF and Microwave System Primer," which begins on page 30 of May 1991 *QST*, has antenna ideas and lots of other useful information on how to get the most out of your UHF station.

Acknowledgments

Thanks to Don Hilliard, WØPW, for building and testing a prototype, and for preparing the schematic.

Notes

[1] R. Campbell, "A Single Board No-Tuning 23 cm Transverter," *Proceedings of the 23rd Conference of the Central States VHF Society* (Newington: ARRL, 1989), pp 44-52. (Reprinted in this book.)

[2] J. Davey, "No-Tune Transverter for 2304 MHz," Proceedings of Microwave Update '89 (Newington: ARRL, 1989), pp 30-34. (Reprinted in this book.)

[3] J. Davey, "A No-Tune Transverter for 3456 MHz," *QST*, Jun 1989, pp 21-26. Also see Feedback, Oct 1990 *QST*, p 31. (Reprinted in this book.)

[4] R. Campbell, "A Single-Board Bilateral 5760-MHz Transverter," *QST*, Oct 1990, pp 27-31. This article lists, in its end notes, sources for lots of information on microwave antennas and other subjects of general interest. (Reprinted in this book.)

[5] Etched PC boards, crystals, kits of parts, assembled boards and complete transverters are available for this rig and the projects referenced in notes 1-4 and 7, from Down East Microwave, RR 1 Box 2310, Troy, ME 04987, tel 207-948-3741, fax 207-948-5157. Catalog available.

[6] See note 1.

[7] R. Campbell, "A Clean, Low Cost Microwave Local Oscillator," *QST*, Jul 1989, pp 18-23.

[8] See note 7.

[9] R. Campbell, "A Clean Microwave Local Oscillator," *Proceedings of the 1296 and 2304 Conference*, reprinted in *Proceedings of the 21st Conference of the Central States VHF Society* (Newington: ARRL, 1987), pp 51-57.

[10] Down East Microwave (see note 5) carries these hybrid amplifier modules.

[11] Send a no. 10 SAE with one unit of First-Class postage to the ARRL Technical Department Secretary; request the July 1991 *QST* 902-MHz transverter template package. The PC-board artwork is copyrighted by Down East Microwave. Feel free to use it to make boards for your personal, noncommercial use, but not for commercial purposes.

[12] See note 5.

[13] See note 3.

Reflections on the KK7B 903- and 1296-MHz No-Tune Transverters

By Ron Neyens, NØCIH

Introduction

I have built both the 903- and 1296-MHz versions of the KK7B No-Tune Transverters designed by Rick Campbell. Both are complete and ready to operate even if only at the QRP-output level. These transverters are about the fastest, easiest way of getting active on 903- and 1296-MHz. They're easy to assemble and, fortunately, don't require any of those hard-to-find, one-of-kind components or parts.

A few hours with the soldering iron is all that's required to put these transverters on the air. All the frequency-sensitive inductors and bandpass filters are etched directly on the circuit board. This eliminates the need for fancy test equipment and the tedious tuning process usually required with home-brew transverters. I found that I spent more time putting the assembled board into a chassis than soldering the parts on the circuit board and testing it.

As with other projects, someone always looks for ways to improve the design or layout. Hopefully this effort results in an improvement on the original design. What I am presenting here are some of the components I've used and the layout changes I've made. These changes came about during the process of building and testing the transverters. Hopefully this information will help others that are using these transverters now or are thinking of building equipment for 903 and/or 1296 MHz. This paper refers to the 903-MHz transverter but the information presented here applies just as well to the 1296-MHz version.

Background

To begin with, I want to supply a little background information on the 903-MHz transverter. I purchased the etched circuit board from Down East Microwave while I was attending the 1989 Central States VHF Conference in Chicago. The quality of the G-10 circuit board and the etched pattern on it was excellent. A *902K Preliminary Instruction Manual* was included with the circuit board and it contained a description of operation, a parts list and several component layout diagrams. The only error with the preliminary manual was that the component layout diagrams were for an earlier version of the transverter layout. These were only minor errors and I had several articles on the transverter to refer to. The most recent release of the *902K Assembly And Operational Guide* has corrected some of these errors. The ones that have not been corrected are pretty obvious.

The recommended parts list in the preliminary manual provided excellent information about the parts needed in the transverter. All parts are easily obtained through a variety of mailorder suppliers and local vendors. If you don't feel like collecting the individual parts, a packaged kit for the transverter is available from Down East Microwave. Going a bit further in cost, Down East Microwave also sells a complete ready-to-run transverter with an optional aluminum housing and IF switching board.

In my case, I decided to build the transverter from scratch, in the process saving a few dollars and at the same time learning something. The local oscillator, receive downconverter, and transmit upconverter are contained on one 5" × 7" circuit board (Note: The 1296-MHz version requires two circuit boards for a complete transverter.)

The local oscillator (LO) circuit is simple and straightforward. It contains a crystal controlled Butler oscillator, X8 multiplier, two triple stage stripline bandpass filters, and a stripline power divider. A 94.75000-MHz, fifth-overtone crystal is multiplied by 8 to provide a final output frequency of 758-MHz at approximately +13 dBm. The crystal (Y1) was purchased from International Crystal (ICM) but is also available directly from Down East Microwave. I used a Motorola 78L05 for the +5-volt regulator and a Motorola 2N5179 for transistors Q1 and Q2. For the ×8 multiplier diode I used a 1N5711 Schottky diode in place of the recommended HP 2835 diode. I installed the diode so that the anode lead extended through the circuit-board pad and was soldered on both the top and bottom of the circuit board.

A combination of one Avantek MSA-0685 monolithic microwave integrated circuits (MMIC) and four MSA-1104s are used in the ×8 multiplier/bandpass filter circuit. These provide the required LO level for two double-balanced mixers. I purchased all of the MMICs from Microwave Components of Michigan. The double-balanced mixers I used are Mini-Circuits SBL-1Xs. These also can be purchased from Microwave Components of Michigan.

The receive downconverter circuit uses a pair of Avantek

MSA-0685s MMICs. This combination is presently the best to use as far as gain, low noise figure, and device cost. A Mini-Circuits MAR-6 can be used as a replacement for the Avantek MSA-0685s without any changes to the bias resistor values.

In the final receiver configuration of the transverter, I have a plate-line preamp using an MGF-1302 in front of the receiver. This was added to improve the overall noise figure of the transverter. The design of the preamp is similar to one described by Al Ward, WB5LUA, for 1296 MHz. Without the preamp the noise figure of the transverter is between 3.8 to 4.0 dB with a gain of approximately 35 dB.

The transmit upconverter circuit uses a single Avantek MSA-0685 driving an Avantek MSA-1104. This combination is also the best set up with what MMICs are presently available. One-mW drive at 145 MHz provides approximately +13 dBm (20-mW) output at 903 MHz. A Mini-Circuits MAV-11 can be used as a replacement for the Avantek MSA-1104 without any changes to the bias resistor value. If a 0-dBm (1-mW) power output is desired, it can be achieved by using two MSA-0685s.

Several RF chokes are used on the circuit board. These were hand wound on a 3/32" drill bit using #24 AWG enameled copper wire. All are 8 turns, close wound, with the exception of a 7-turn, close-wound choke across the base/collector of transistor Q1 and a 0.33-µH molded RF choke across crystal Y1. If the oscillator fails to start properly or the 903-MHz signal is distorted, spread or compress the 7-turn choke slightly.

I followed the recommended parts list in the preliminary manual, but made a change to the capacitors used. Instead of a mixture of ceramic disk caps and chip caps, I decided to use 0.110" × 0.110" ceramic chip caps throughout (including the local oscillator). I also made a change in the values of bypass capacitors by using three (22 pF, 0.01 µF and 0.001 µF) instead of the recommended two (22 pF and 0.01 µF).

That about sums it up. The ceramic chip caps are the only real deviation I made from the original Down East Microwave parts list. Overall, the placement of components was per the assembly instructions supplied by Down East Microwave. I recommend that you refer to the notes at the end of this article for more information.

A Housing to Put It in

I cut the circuit board down to 5" × 7" to expose the copper-clad ground plane along all the edges. After installing the parts, I formed a box around the circuit board using 0.32" thick × 1.00" wide brass strips. I had access to a sheet metal brake and this allowed me to bend the strips to fit the circuit board, and form the corners of the box. The component side of the circuit board was placed ½-inch up from the bottom edge of the strip. The walls were then soldered to the ground plane on both the top and the bottom sides of the board to provide a continuous RF ground.

To complete the RF-tight assembly, I needed top and bottom covers. Refer to Fig 1. The covers were made by using two pieces of 0.060" sheet aluminum cut to 5⅛" × 7⅛". I "sandwiched" the transverter between the two covers and used 1" long, 4-40 threaded, hex standoffs on each corner to maintain separation. Two-meter input/output and 903-MHz input/output connections were made using SMA chassis connectors soldered to the brass walls. The completed unit makes a nice RF-tight assembly that can be easily mounted to a chassis.

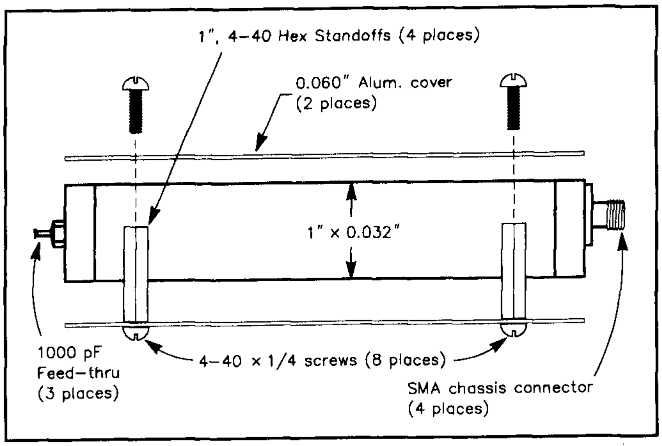


Fig 1

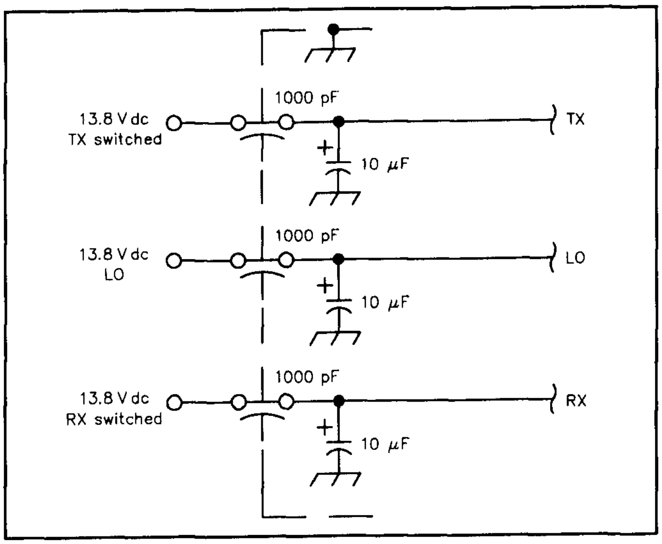


Fig 2

Supplying dc Power

Refer to Fig 2. All power is routed through 1000-pF, 50-V feedthrough capacitors. These are soldered directly to the brass walls of the box to insure a low inductance RF path. I used three feedthrough capacitors: one for the local oscillator, one for the transmitter, and one for the receiver. The 10-µF, 35-V Tantalum capacitors are used for additional bypassing of the dc supplies. I routed all of the wiring on the unetched side of the circuit board. Placing it on this side helps prevent any problems with RF floating on the dc bus.

Final Testing and Modifications

While testing the transverter on a spectrum analyzer, I found I was getting excessive spurious outputs from the transmitter. I tried squeezing and expanding the turns of L5 and L6 but this didn't get me any closer to a cleaner output. The only other thing I could do was move the bypass capacitors closer to the MMIC bias resistors. The easiest stage to work on was the final output MMIC. I moved the bypass capacitors as close to the bias resistor as possible and rechecked the output on the spectrum analyzer. I found that the gain had increased slightly and the spurs had decreased.

I heated up the soldering iron and unsoldered the remaining bypass capacitors in the transmitter chain. Not wanting to do half a job, I also moved the other bypass capacitors on the board closer to their intended source. Checking the output again showed an increase of approximately 3 mW over that previously obtained. This brought the transverter's output to approximately 18 to 19 mW, where it should be. The spectrum analyzer also revealed that the spurs had decreased considerably. Some of the spurs had dropped 10-20 dB. With a signal generator for an input signal, I checked the receiver for any improvements but saw no real evidence of any change. [KH6CP/1 comments: The chip caps may have caused the spurious signals. They are "better" than disk ceramics, resulting in sharper parallel resonances that prevent good bypassing. Jim Davey has recommended series resistors to help reduce these resonances. Moving them around varies the inductance of the parallel resonant circuit. See *Microwave Update 93*, p 58.]

Summary

I hope I have helped someone with these ideas. If you're just considering building one of the KK7B transverters, spend a little money and use chip caps throughout the transverter. Save those disk ceramics for something on HF or dc. If you're getting insufficient output, low gain, or a large spurious output from your transverter it might help to check the bypass capacitors and reposition them.

The "sandwiched" case idea has been used on a variety of projects around my shack. It has been ideal for packaging preamps, low-level power amps, and converter subassemblies. Whenever I need a housing for a project I get the size I need since the circuit board determines the dimensions. Besides this, it doesn't cost much in material when compared to a diecast aluminum box.

Even though the output from this transverter is only about 18 mW (10 mW for the 1296-MHz version) it's enough to start with. With the variety of RF brick modules on the market, it's possible to have up to a 20-W transverter very easily. With that amount of output, band openings should become interesting.

Notes

[1] *SHF Systems 902K Preliminary Instruction Manual*, Down East Microwave, 1989.

[2] R. Campbell, "A Clean, Low-Cost Microwave Local Oscillator," *QST*, July 1989, pp 18-23.

[3] R. Campbell, "A Single-Board No-Tuning 23-Cm Transverter," *Proceedings of the Central States VHF Society 1989*. (Reprinted in this book.)

[4] R. Campbell and D. Hilliard, "A Single-Board 900-MHz Transverter With Printed Bandpass Filters," *Proceedings of Microwave Update '89*.

[5] J. Hinshaw, "Modular Transmit And Receive Converters For 902-MHz," *Ham Radio*, March 1987, pp 17-20, 23-26.

[6] *SHF Systems 902K Assembly and Operational Guide*, Down East Microwave, 1990.

A Single Board No-Tuning 23-cm Transverter

By Rick Campbell, KK7B

(This article, which originally appeared in the *Proceedings of the Central States VHF Society 1989,* is reprinted from *The ARRL Handbook for Radio Amateurs.*)

This transverter uses bandpass filtering with printed hairpin filters. It will work with any MMICs capable of operating at 1.3 GHz at the appropriate input and output levels. New MMIC types may be easily substituted simply by changing the MMIC and its bias resistor.

System Block Diagram

Fig 1 is the transverter block diagram and Fig 2 is the schematic diagram. Fig 3 shows the layout of the 5- × 7-inch G-10 board with all functional blocks labeled. In the transmit path, FL1 removes the image, LO (local oscillator) and higher-order spurious outputs. FL3 selects the desired LO harmonic. The LO signal is filtered again in FL4 before the LO is split and applied to the transmit and receive mixers.

The received signal passes through FL5, which attenuates out-of-band signals. After amplification, image noise added to the received signal is attenuated by FL6.

One noteworthy feature of this transverter is that the choice of LO drive frequency is left to the builder. The filters are narrow enough to permit use of outputs at 1152, 576, 384, 288 and 230.4 MHz. Fig 3 shows component values for 576-MHz drive from the LO board described in note 1. Other

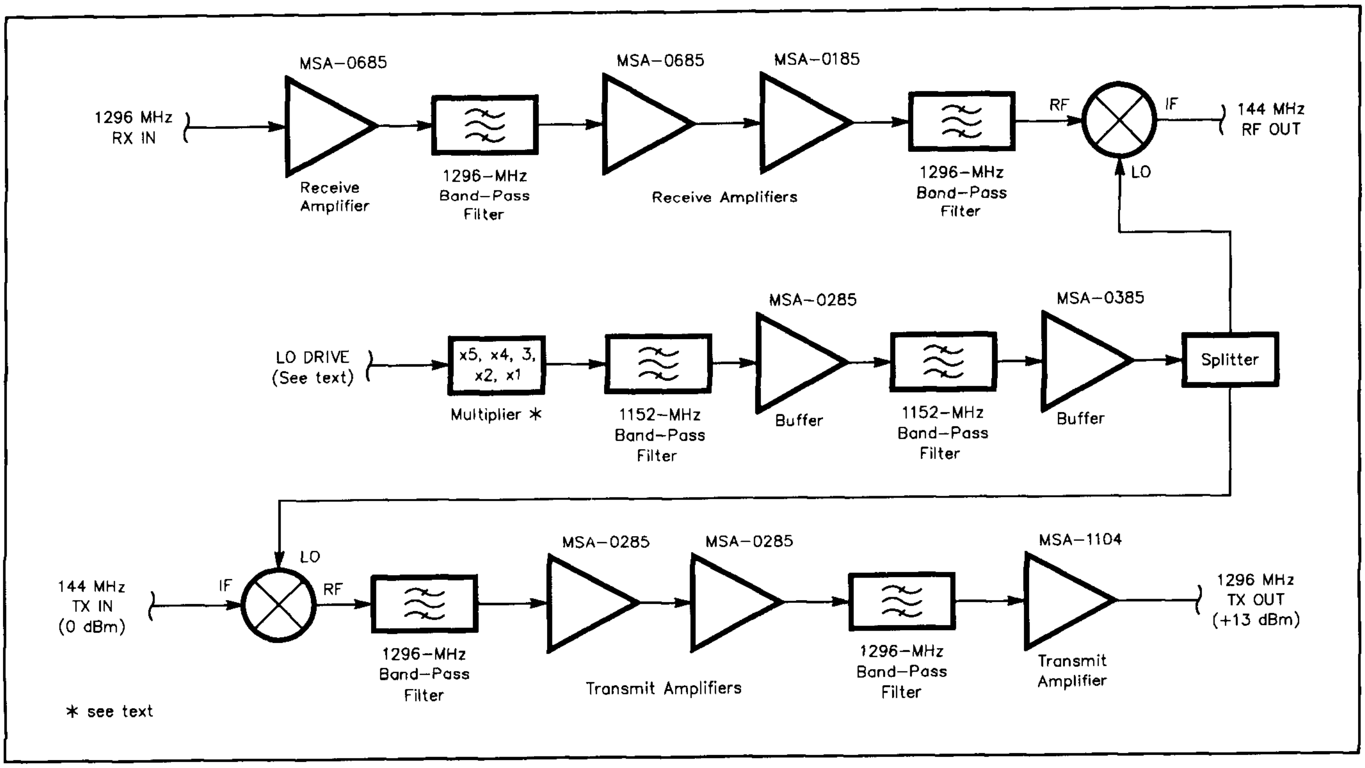


Fig 1—Block diagram of the single-board transverter. Extensive use of printed-circuit band-pass filters simplifies construction and alignment.

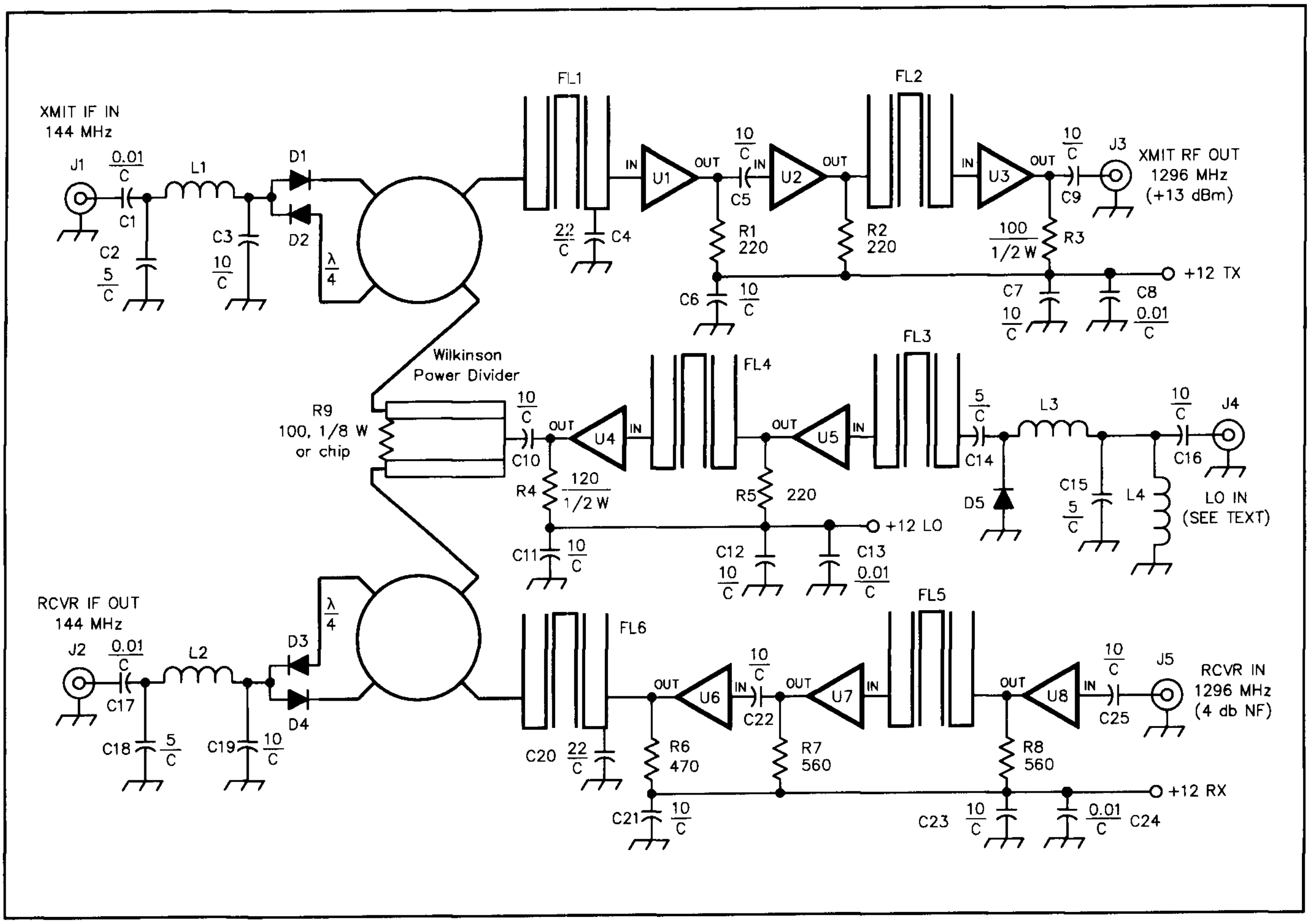


Fig 2—Schematic of the 1296-MHz transverter. All resistors and capacitors are chip types except as noted. Diodes D1 through D4 are HP 5082-2835. FL1 through FL6, the mixer tuned circuits and the Wilkinson power divider are etched on the circuit board.

LO frequencies may require minor component changes.

Circuit Performance

Careful measurements show a mixer conversion loss of 5.5 dB. The output for 1-dB compression is 0 dBm. LO suppression is about 25 dB. The mixers compare favorably with more-expensive packaged units.

Transmit output is about 13 dBm, which is suitable for some applications without further amplification. A major advantage of the MMIC output stage is that it offers an unconditionally stable, near-50-ohm source to the following stage or antenna.

Receive noise figure is less than 4 dB. The unconditionally stable 50-ohm load presented by the input MMIC is useful for direct connection to an antenna, and ideal as the stage following an external GaAsFET preamp.

The filters allow coverage of the entire 1240- to 1300-MHz band with a 144-MHz IF and appropriate LO drive. The bandpass filters used, although not mechanically critical, remove significant spurious responses. The passband response of one RF and one LO filter section is shown in Fig 4.

Construction

All components are surface mounted on the etched side of the board. The most tedious step is drilling the holes next to the MMIC mounting pads and soldering the "C" wires to provide a low-impedance ground. Plated-through holes would eliminate this step. After preparing the board with its "C" wires, mount the chip capacitors, inductors and connectors. Finally, install the MMICs and bias resistors.

The bottom of the hairpin U is a low-impedance point for signals in the filter passband. A bypass capacitor at that point has little effect on the filter passband but provides a ground path for VHF signals. Although 22-pF capacitors are specified, values from 12 to 30 pF gave good results in the prototypes.

The multiplier works best with drive levels of 0 to 10 dBm, without a bias resistor. For higher drive levels, the bias circuit described in Note 2 will provide more output. With 10-dBm drive at 230.4 MHz, output from FL3 is about –20 dBm, so about 33 dB of additional gain is needed. With 10-dBm drive at 576-MHz, output from FL3 is –3 dBm, so only 16 dB of gain is required.

Pads are included on the board for up to three MMICs. Choose the appropriate number of MMICs based on available LO drive level and frequency. FL3 and FL4 must be isolated. If only one MMIC is used it should be between them. If sufficient LO drive is already available at 1152 MHz, use a

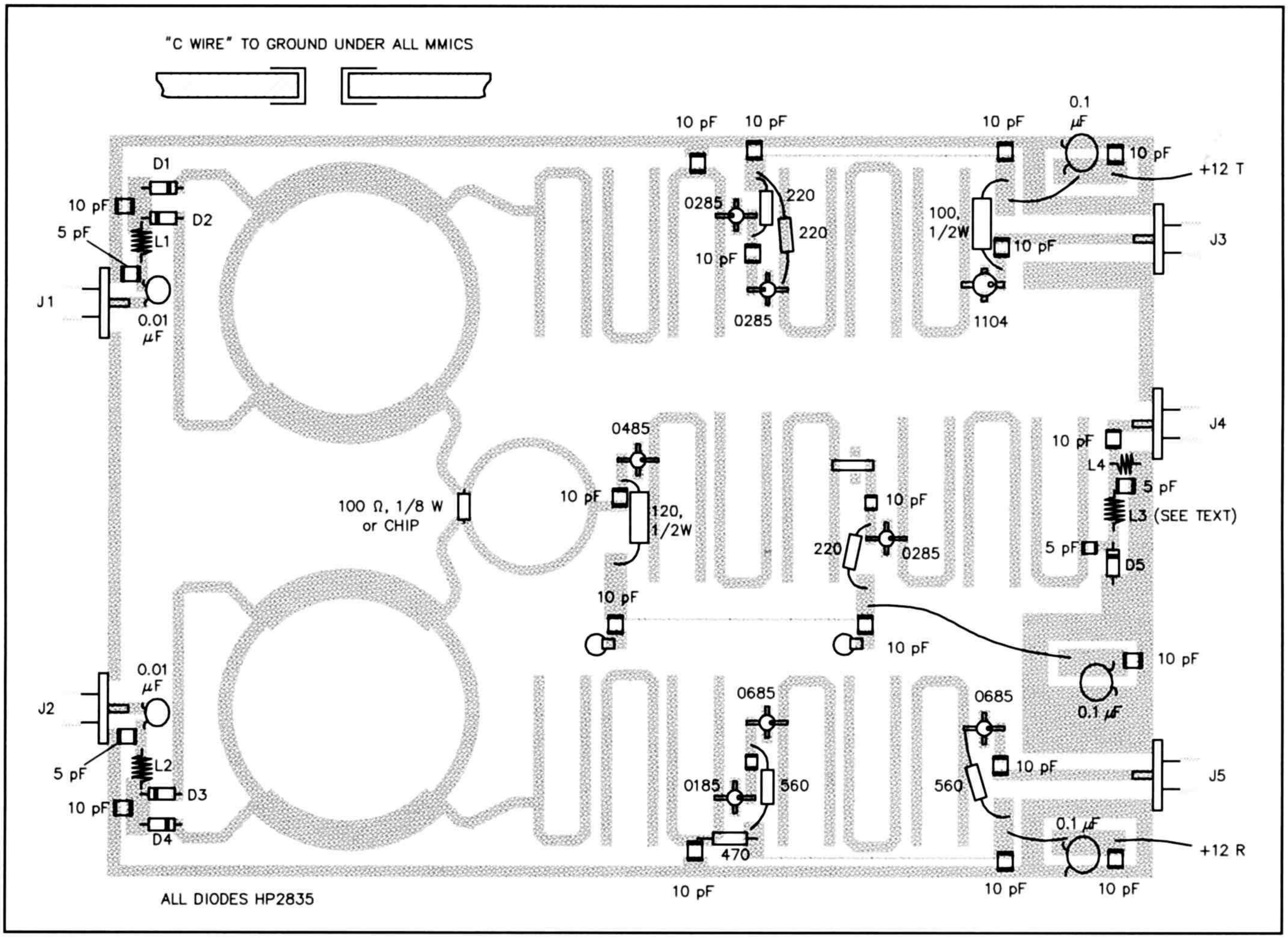


Fig 3—Part-placement diagram for the transverter. The "C" wires under each MMIC connect the component pad to the ground plane on the other side of the board.

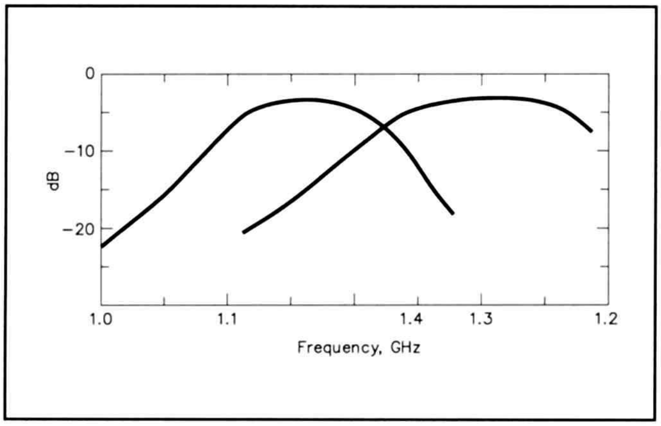


Fig 4—Passband response of one LO filter (A) and one RF filter (B) in the transverter. With appropriate choice of LO frequency, the transverter is suitable for use across the entire 23-cm band.

6-dB attenuator pad between FL3 and FL4.

Alignment

The tuning components in the LO multiplier are quite broad: network Q is about 1. The 576-MHz network will also work for lower frequency inputs, but the component values can be optimized for each drive frequency. L3 and L4 may be increased to 3 turns for 384 MHz, 4 turns for 288 MHz and 5 turns for 230.4 MHz if necessary. If LO drive is sufficient to saturate U4, however, mixer performance will be relatively unaffected by small changes in LO level, and tuning is unnecessary. A high-level MMIC is recommended for the LO-output stage.

Parts and Circuit Boards

Components are available from Microwave Components of Michigan, whose address appears elsewhere in this chapter. Etched and plated FR-4 circuit boards, complete kits and assembled transverters are available from Down East Microwave, whose address also appears elsewhere in this chapter.

The author acknowledges the assistance of Jim Davey, WA8NLC, John Miller of Michigan Technological University, Bob Dryden, W4OJK and John Molnar, WA3ETD.

Notes

[1] R. Campbell, "A Single-Board No-Tuning 23 cm Transverter," *Proceedings of the 23rd Conference of the Central States VHF Society* (Newington: ARRL, 1989), pp 44-52.

[2] R. Campbell, "A Clean Microwave Local Oscillator," *Proceedings of the 21st Conference of the Central States VHF Society* (Newington: ARRL, 1987), pp 51-57.

A No-Tune Transverter for 2304 MHz

By Jim Davey, WA8NLC

4664 Jefferson Township Place
Marietta, GA 30066

(From December 1992 *QST*)

Introduction

The circuit, layout and performance of the 2304-MHz transverter are similar to those of the 3456-MHz design described in the article that follows this one. It has the following features:

- 10-mW transmit output
- Receiver noise figure of 3.5-4.0 dB
- Same 2-meter IF requirements (1-mW drive level)
- Same size circuit board
- Same no-tune local oscillator
- Even easier to assemble than the 3456-MHz version

For construction details and background information, refer to the 3456-MHz article. Only the major features of the 2304-MHz transverter are covered in this article. A preliminary version of this transverter was described in *Proceedings of Microwave Update '89*.

Construction

The entire transverter can be assembled in a few hours, including the box. The brass-box type of enclosure is recommended for mechanical and electrical bonding of the connectors. A cover is optional. The 2304 transverter is indifferent to the presence of a cover. A few builders of the 3456-MHz transverter, however, have noticed increased spurious outputs

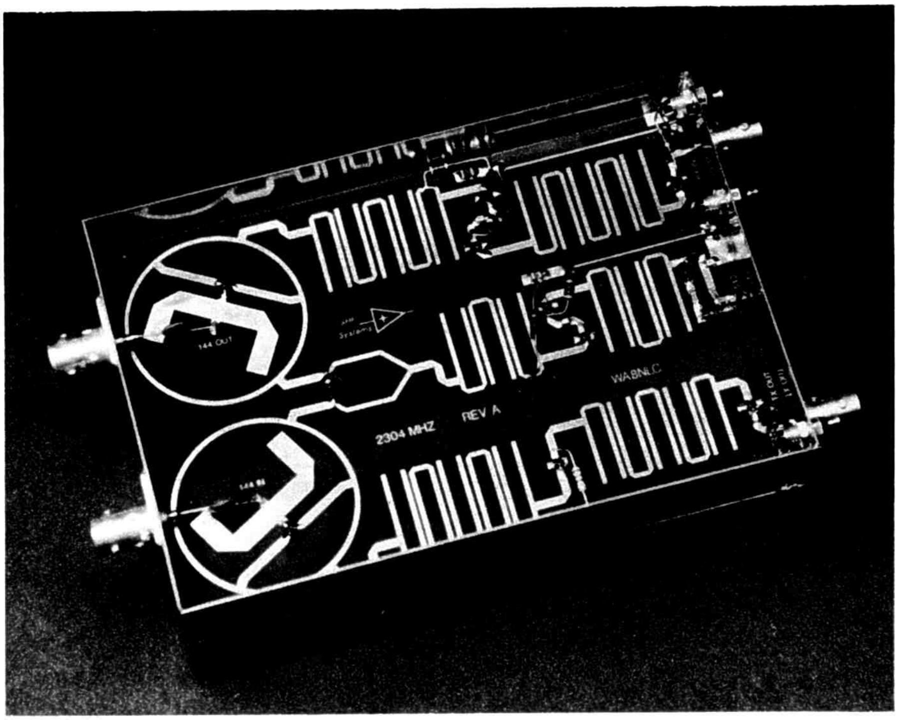
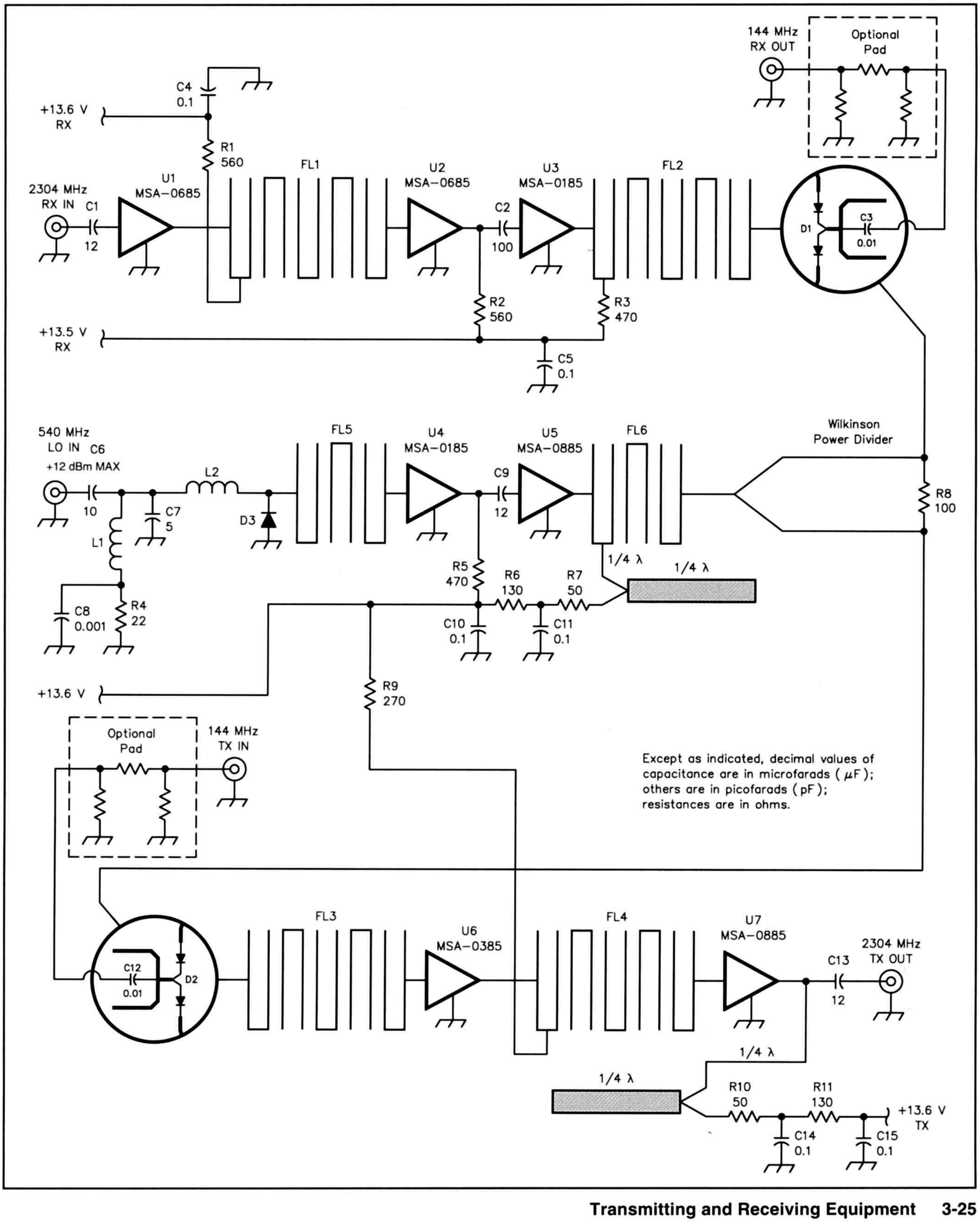


Fig 1—Schematic of the 2304-MHz transverter. See the text for parts sources. When building this transverter, follow good microwave construction techniques, as detailed in the 3456-MHz transverter article that follows this one. The Avantek MSA-0185 is equivalent to the Mini-Circuits MAR-1; the MSA-0685 is equivalent to the MAR-6, etc. (Lead labeling varies, however; follow the manufacturer's instructions.) Bias-resistor values are given for 13.6-V operation. All capacitors in the 2304-MHz signal path are 50- or 100-mil ceramic chips, except for the 0.01-µF coupling capacitors at the first IF input and output. Here, use ceramic discs, encapsulated chips or other low-loss capacitors. Use ¼-W carbon-composition or film resistors.

D1, D2—Hewlett-Packard HSMS-2822 surface-mount diode pair.
D3—Hewlett-Packard 5082-2835 Schottky diode.
FL1-FL6—Etched band-pass filter.
L1—4 turns #28 enameled wire, 0.075 in. ID, closewound.
U1, U2—Avantek MSA-0685 or Mini-Circuits MAR-6.
U3, U4—Avantek MSA-0185 or Mini-Circuits MAR-1.
U5, U7—Avantek MSA-0885 or Mini-Circuits MAR-8.
U6—Avantek MSA-0385 or Mini-Circuits MAR-3.

of the transmitter with a cover in place. When mounting components, keep the bias resistors flat against the board and away from the filter elements. Wrap a small piece of copper foil through the ground holes for the bypass capacitors, under each MMIC and at the ground end of the LO multiplier diode. I used SMA connectors for the 2304-MHz ports, but BNC connectors would be acceptable at this frequency.

As before, no matching is done at the IF ports. This has not been a problem, except if you want to measure the noise figure of the receiver. I have noticed that newer noise-figure



meters, like the HP 8970, have a problem with this arrangement due to the presence of other signals at the IF port, presumably the LO. If you want to check the noise figure, first run the receiver IF through a 2-meter converter to a lower IF, such as 28 MHz. When this is done, the problem with the test equipment should disappear.

Design

The transverter, shown schematically in Fig 1, is built on 0.032-in. woven Teflon board material with a dielectric constant of 2.5 and half-ounce copper foil. The key to the circuit's simplicity lies in the printed microstrip third- and fifth-order Chebyshev band-pass filters. These filters have excellent stopband rejection and low insertion loss. Midband insertion loss is just 2.1 dB, thanks to a low-loss Teflon PC-board substrate.

The center frequency and bandwidth of the receive filters were chosen to also allow receive-only operation on the OSCAR Mode S downlink at 2401 MHz using a 564.25-MHz LO. The transmitter-filter bandwidth is narrower, to help suppress unwanted mixing products. The transverter shouldn't be used for transmitting above 2325 MHz without an external cavity or interdigital filter that attenuates the LO signal by at least 10 dB.

Transmit Converter

Transmitter power output is 10 to 20 mW (10 to 13 dBm) and varies slightly with component variations in the Avantek MSA-0885 MMIC output stage. Although more power would help balance the transverter's capability with that of most home stations, power-amplifier stages are best implemented as external accessories. It's good design practice to limit the amount of gain in a single box to 30 dB or so, unless special efforts are taken to prevent feedback.

Receive Converter

The transverter's receive-converter noise figure (NF) is approximately 4.5 dB. This may seem high by today's standards, but it can be easily dropped to 1 dB or less by a modern GaAsFET preamplifier with a gain of 14 dB or more.

The transverter's first receive stage, an Avantek MSA-0685 MMIC, provides a 50-W input impedance to properly terminate an outboard preamp. The preamp can be mast mounted or connected directly to the transverter input.

Performance

When driven with 1 mW of 144-MHz RF, the transverter produces a clean 2304-MHz signal. The 1-dB compression point is 10 mW. All spurious products are down from the fundamental output by at least 50 dB. For a receive frequency of 2304.1 MHz, the measured image rejection is more than 70 dB and the measured NF is 4.3 dB.

Be careful not to exceed the transmit mixer's maximum drive rating of 0 dBm (1 mW). If you have extra transmit gain

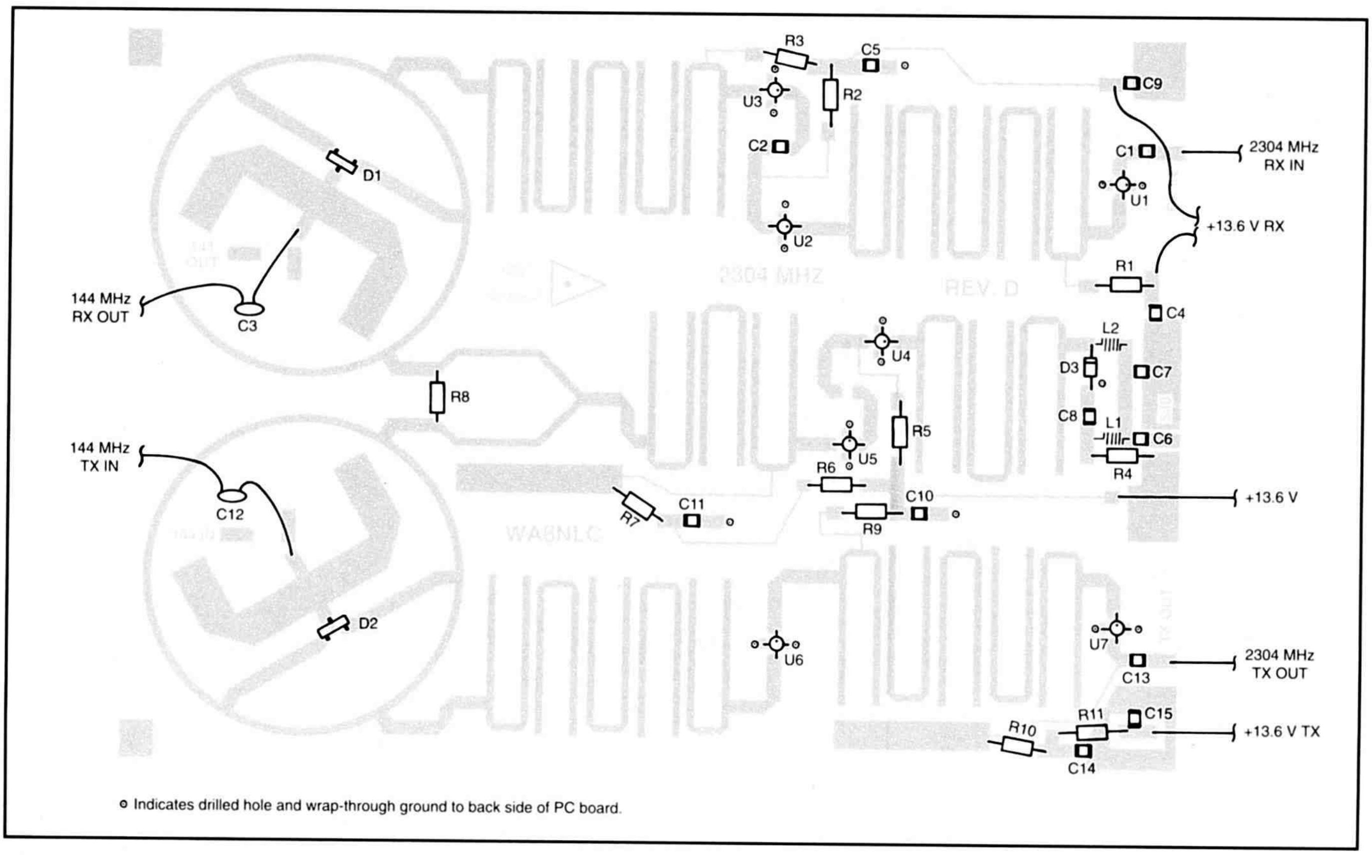


Fig 2—Part-placement guide for the 2304-MHz transverter (not shown actual size). All components mount on the etched side of the board. Feedthrough grounds, indicated by circles, *must* be installed and soldered top and bottom. Follow MMIC manufacturer's lead coding.

in your final configuration, reduce the transmitter drive.

Parts

Fig 2 is a part-placement drawing. Parts are available from Microwave Components of Michigan, PO Box 1697, Taylor, MI 48180; kits and assembled units are available from Down East Microwave, RR1 Box 2310, Troy, ME 04987, tel 207-948-3471, fax 207-948-5157. As in the 3456-MHz transverter, the success of the no-tune approach depends on how acccurately you can duplicate the printed filters. I strongly recommend you purchase the board from Down East Microwave, instead of trying to make your own.

A No-Tune Transverter for 3456 MHz

By Jim Davey, WA8NLC

(Reprinted from June 1989 *QST*.)

The development of easy-to-build equipment for the microwave bands has increased dramatically in the past few years. No longer do you have to have a small fortune invested in test equipment to get the satisfaction of building your own station for the bands above 1 GHz. Nor do you have to rely on the availability of surplus components gathered at hamfests—or be lucky enough to live in a high-tech part of the country where surplus is more plentiful—to home-brew your own equipment.

The recent increase in activity on the microwave bands has been well documented in *QST* and elsewhere. In the past two years alone, amateurs have conquered the difficult EME challenge on 3456, 5760 and 10368 MHz and set impressive distance records on nearly every microwave band through 47 GHz. UHF contest stations now regularly have 2304- and 3456-MHz equipment available. Our knowledge of the propagation characteristics of these bands has also benefited from the increased activity.

Commercially manufactured ham equipment from Europe has been available for a few years for the more popular microwave bands, and this equipment has helped to spawn activity. The 3456-MHz band, however, has not been supported by commercial manufacturers as of this writing. Let's not let the lack of commercial equipment stop us from having a little fun! Besides, you can get a lot of enjoyment and personal satisfaction from building your own equipment. I can personally attest to this: My own station uses no commercially manufactured equipment except for the IF rigs. This article describes a transmitting and receiving conversion module (transverter) for the 3456-MHz band. Any multimode 2-meter transceiver can be used as a tunable IF. The transverter has several features that make it ideal for the newcomer to 3456 MHz and the veteran looking for a simple loaner rig for grid-square expeditions:

- It doesn't require RF alignment or microwave test equipment for proper operation.
- The entire transverter, minus the 552-MHz local oscillator, is contained on one PC board, reducing the need for separate enclosures and expensive RF connectors and cables.
- The transmitter features 10-mW output, and the receiver features a 4-dB noise figure (NF)—performance sufficient for a lot of interesting work on this band.
- Inexpensive diodes and MMIC gain blocks are used to keep the cost low.
- An external receiving preamplifier, transmitter power amplifier and antenna relay can be added to make the unit a high-performance package for fixed or portable operation.
- The transverter can also be used as an IF for the higher microwave bands above 10 GHz.

Background

The straightforward design of this transverter is a result of two fairly recent developments in the amateur microwave field. First, the introduction of cascadable monolithic-microwave integrated circuits (MMICs) several years ago revolutionized the design of microwave amplifiers. Not only are MMICs inexpensive, but most are unconditionally stable and can be cascaded for increased gain or paralleled for greater power output. Al Ward, WB5LUA, authored excellent articles for *QST* concerning the use of these devices.[1] You are encouraged to review these articles for a more thorough treatment of the subject.

The second development contributing to the design of this transverter is my recent work on microstrip band-pass filters that provide the required selectivity for this system without the need for tuning adjustments. These filters were first introduced to amateurs in 1987 at the Microwave Update Conference in Estes Park, Colorado.[2] A later paper reported on further work to improve the SWR of the filters and compared the microstrip filters to other commonly used filters for the microwave bands.[3]

To avoid taxing your wallet or your patience, I used parts that are inexpensive and easy to get. You can get this transverter up and running for less than $200 including the LO. The idea here is to get you on the 3456-MHz band as easily as possible without relying on unique or hard-to-find components.

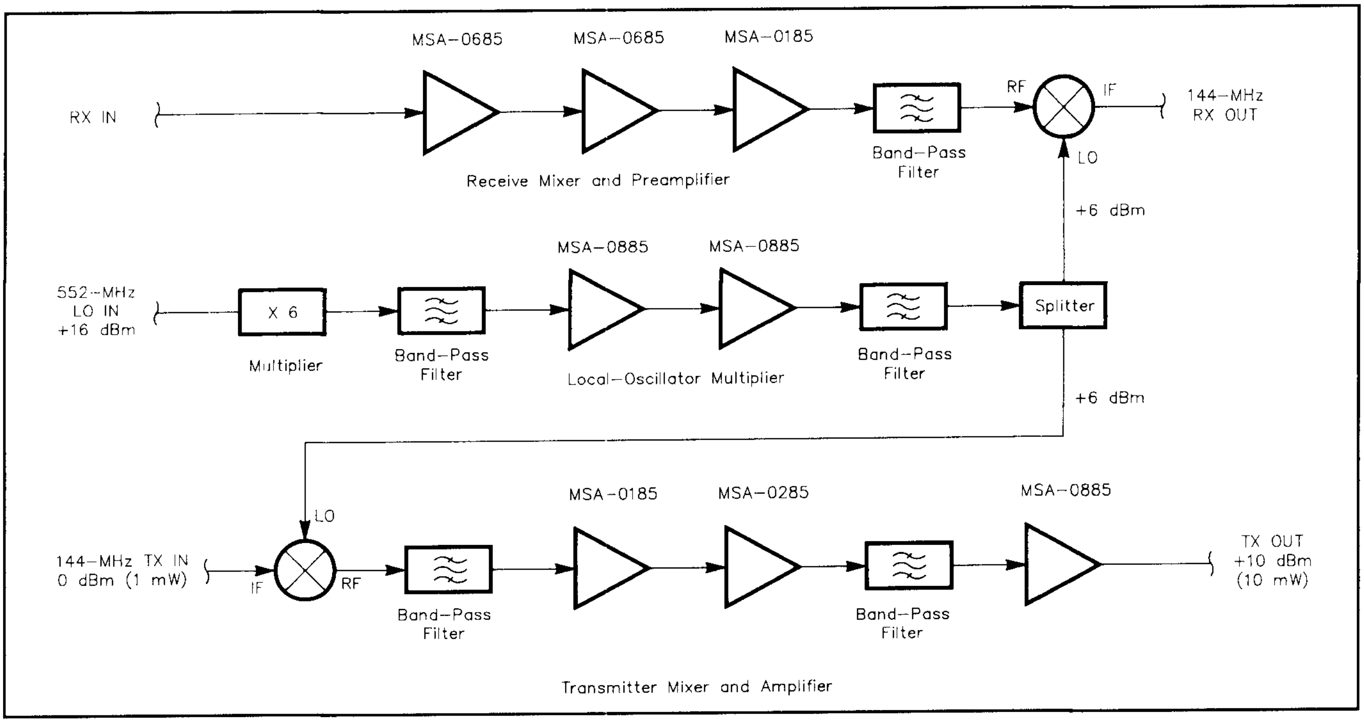


Fig 1—Block diagram of the 3456-MHz transverter.

Circuit Description

The transverter is divided into three basic sections: transmit mixer/amplifier, receive mixer/preamplifier and local-oscillator multiplier. See Fig 1. Each section is described in the following paragraphs.

Local-Oscillator Multiplier

The transverter requires an external 552-MHz local oscillator (LO) signal that is multiplied by six in the transverter to obtain 3312-MHz injection for the mixers. For the multiplier circuit, I used an idea developed by Rick Campbell, KK7B, who demonstrated that a simple diode multiplier and inexpensive MMIC gain stages can produce a clean microwave local oscillator. Although Rick's design placed each stage in a separate box, I was able to implement the entire multiplier on micro strip using my own printed filters. Rick used the printed-filter design idea to produce a clean 552-MHz LO module that I use as a companion to this transverter.[4]

The transverter requires +16 dBm (40 mW) of 552-MHz energy from an external LO. As shown in Fig 2, the first stage of the multiplier is a broadly resonant circuit that drives D1, a Schottky-diode comb generator. Following D1 is FL1, the first of two band-pass filters used to select the desired 3312-MHz output. Two stages of amplification (U1, U2) using Avantek MSA-0885 MMICs bring the level up to +11 dBm. A second filter, FL2, is used to further clean up the LO and reduce broadband noise that is generated in the amplifier stages. Following FL2, a 90° hybrid divider splits the LO signal into two equal outputs for injection into the transmit and receive mixers. The LO output level to each mixer is +6 dBm. Undesired products in the LO output are 35 dB below the carrier.

Transmit Mixer and Amplifier

The transmit mixer (Fig 3) is a 3/2 wavelength rat-race balanced design from an article by H. Paul Shuch, N6TX.[5] The original article described a 1296-MHz mixer etched on G-10 PC-board material. I developed the 3456-MHz mixer by resizing the line lengths and widths for the new frequency and for a Teflon-fiberglass substrate. To keep the mixer as efficient as possible and maintain high isolation between ports, I used a matched pair of diodes (D2, D3) in a surface-mount package. Isolation between the mixer's LO and RF ports is greater than 20 dB—about as good as any low-cost commercial mixer.

Conversion loss—about 9 dB—is a little higher (worse) than most commercial mixers. Impedance matching was not done at the IF port and did not appear to be necessary with the ICOM IC-202A transceiver I use as an IF rig.

I used three stages of amplification and two filters in the transmit-amplifier chain to reach the final 10-mW-output level. The mixer RF port drives FL3, the first 5-pole band-pass filter. Centered at 3500 MHz, FL3 strips image energy (3168 MHz) from the mixer output and also contributes about 20 dB of rejection at the LO frequency. Rejection at the LO frequency is important. With an LO injection level of +6 dBm and LO-to-RF-port isolation of 20 dB, the level of the LO signal at the RF port of the mixer is –14 dBm. For comparison, the desired signal (the sum frequency at 3456 MHz) is injected at 0 dBm (1 mW) and encounters a 9-dB loss through the mixer to emerge at –9 dBm, only 5 dB above the oscillator feedthrough.

Following the first filter are two amplifier stages (U3, U4) using an MSA-0185/0285 combination. With prototype versions of this transverter, I found that correct MMIC choice for these amplifier stages is critical because band-pass filters

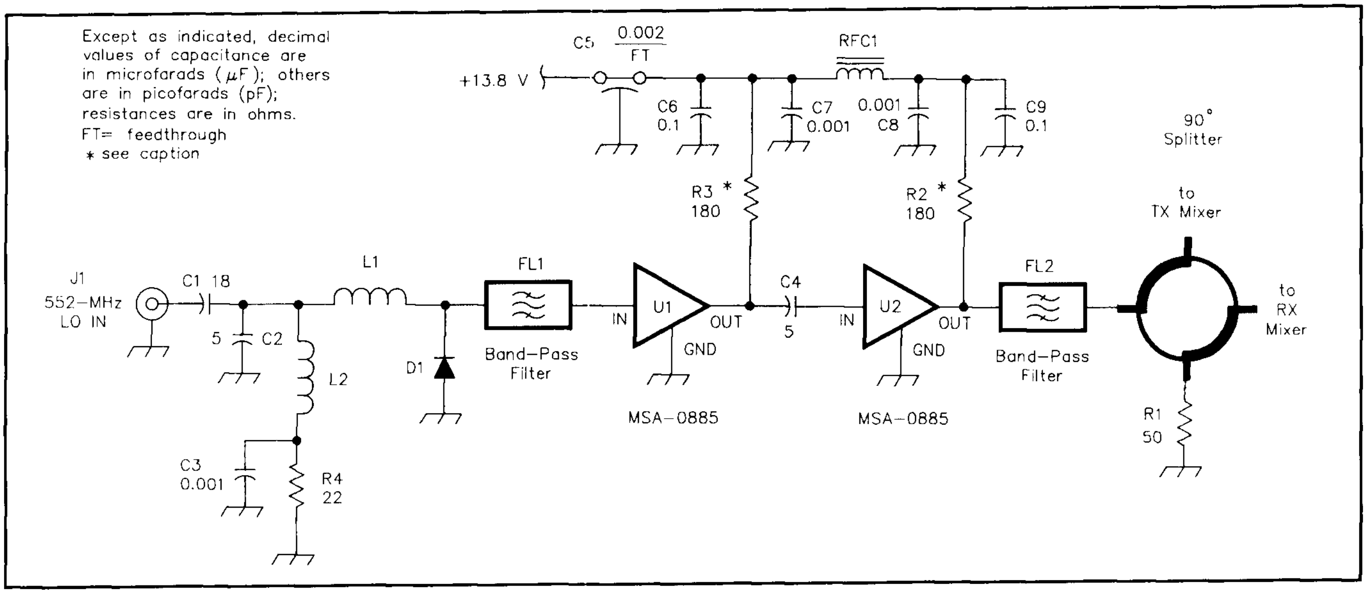


Fig 2—Schematic of the 3456-MHz transverter LO multiplier/amplifier.

C1—18-pF disc-ceramic or silver-mica capacitor.
C2, C4—5-pF porcelain chip capacitor.
C3, C7, C8—0.001-μF ceramic chip (preferred) or disc ceramic capacitor.
C5—0.002-μF feedthrough capacitor.
C6, C9—0.1-μF ceramic chip (preferred) or disc-ceramic capacitor.
D1—Hewlett-Packard 5082-2835 Schottky diode.
FL1, FL2—Band-pass filters printed on PC board.
J1—Female chassis-mount SMA connector.
L1—Inductor printed on PC board.
L2—3 turns no. 28 enam wire, 0.078-in. ID, close-wound.
R1—50-Ω chip resistor.
R2, R3—180-Ω, ¼- or ½-W carbon-film resistor. Note: This value is for 13.8-V operation. See text for information on operation at other voltages.
R4—22-Ω, 1/4-W carbon-film resistor.
RFC1—270 mH subminiature molded RF choke. A suitable alternative is 24 turns no. 28 enam wire on an FT-37-72 toroid core.
U1, U2—Avantek MSA-0885 or Mini-Circuits MAR-8 MMIC.

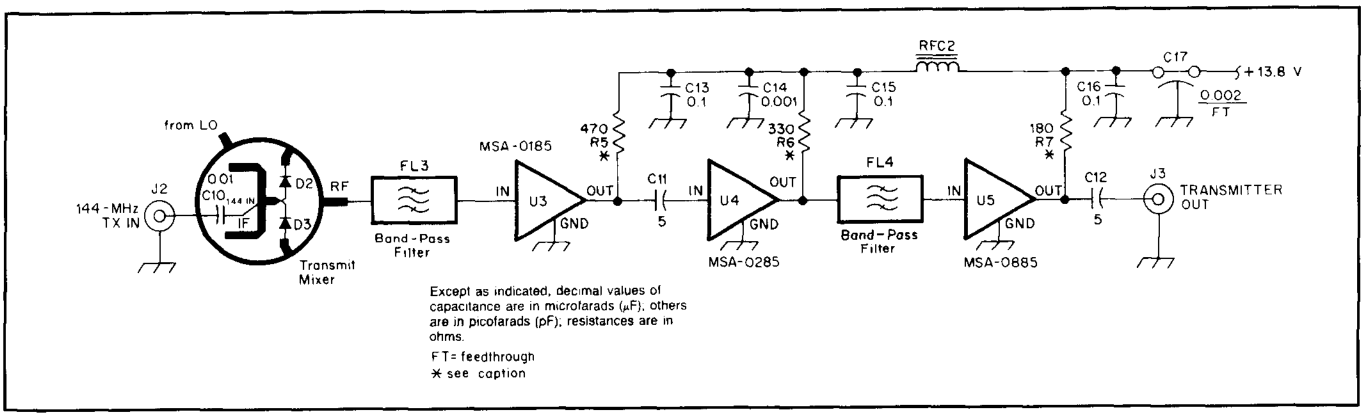


Fig 3—Schematic of the 3456-MHz transverter transmit mixer/amplifier.

C10—0.01-μF disc-ceramic capacitor.
C11, C12—5-pF porcelain chip capacitor.
C13, C15, C16—0.1-μF ceramic chip (preferred) or disc-ceramic capacitor.
C14—0.001-μF ceramic chip (preferred) or disc-ceramic capacitor.
C17—0.002-μF feedthrough capacitor.
D2, D3—Hewlett-Packard HSMS-2822 surface-mount diode pair. See text.
FL3, FL4—Band-pass filters printed on PC board.
J2—Female chassis-mount connector, builder's choice.
J3—Female chassis-mount SMA connector.
R5—470-Ω, ¼-W carbon-film resistor. Note: This value is for 13.8-V operation. See text for information on operation at other voltages.
R6—330-Ω, ¼-W carbon-film resistor. Note: This value is for 13.8-V operation. See text for information on operation at other voltages.
R7—180-Ω, ¼- or ½-W carbon-film resistor. Note: This value is for 13.8-V operation. See text for information on operation at other voltages.
RFC2—270 mH subminiature molded RF choke. A suitable alternative is 24 turns no. 28 enam wire on an FT772 toroid core.
U3—Avantek MSA-0185 or Mini-Circuits MAR-1 MMIC.
U4—Avantek MSA-0285 or Mini-Circuits MAR-2 MMIC.
U5—Avantek MSA-0885 or Mini-Circuits MAR-8 MMIC.

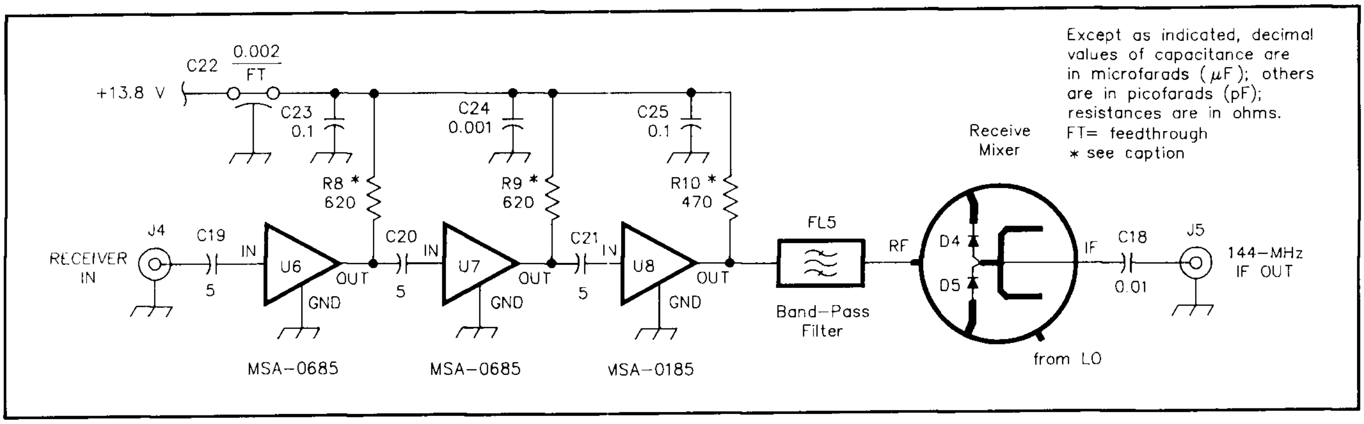


Fig 4—Schematic of the 3456-MHz transverter receive mixer/amplifier.

C18—0.01-μF disc-ceramic capacitor.
C19-C21—1-pF porcelain chip capacitor.
C22—0.002-μF feedthrough capacitor.
C23, C25—0.1-μF ceramic chip (preferred) or disc-ceramic capacitor.
C24—0.001-μF ceramic chip (preferred) or disc ceramic capacitor.
D4, D5—Hewlett-Packard HSMS-2822 surface-mount diode pair or matched pair of HP 5082-2835 Schottky diodes. See text.
FL5—Band-pass filter printed on PC board.
J4—Female chassis-mount connector builder's choice.
J5—Female chassis-mount SMA connector,
R8, R9,—620-Ω, ¼-W carbon-film resistor. Note: This value is for 13.8-V operation. See text for information on operation at other voltages.
R10—470-Ω, ¼-W carbon-film resistor. Note: This value is for 13.8-V operation. See text for information on operation at other voltages.
U8—Avantek MSA-0685 or Mini-Circuits MAR-1 MMIC.
U6, U7—Avantek MSA-0185 or Mini-Circuits MAR-6 MMIC.

(FL3, FL4) are present at the input and output of the amplifier strip. The filters are highly reactive out of band where their return loss is very low (in the range of 0 to 3 dB), so the MMICs used in conjunction with the filters must be unconditionally stable. Another consideration is that the microstrip filters were developed with good 50-ohm terminations on each end. The SWR at the input and output ports of the MMIC should be close to 50 ohms within the passband of the filters to maintain flat filter response and low insertion loss. An examination of the S-parameter data for the MSA-0185 and '0285 shows that they are excellent choices for this application.

FL4 was added in later prototypes to reduce mixing products above 3456 MHz that were present in the output. This filter, a modified version of the filters used in the LO multiplier chain, is centered at 3312 MHz. The combination of FL3 centered at 3500 MHz and FL4 centered at 3312 MHz creates a narrow window at 3456 MHz where the two response curves overlap, giving the effect of a much narrower filter. A single narrow-band, higher-Q filter could be used to accomplish this, but the stop-band rejection for one filter would be inferior to that of two filters. Technical references indicate that the stop-band rejection of a single microstrip filter is only about 40 dB because of surface-wave effects; two separate filters can achieve a total stop-band attenuation of 80 dB. FL3 and FL4 in tandem do a good job of cleaning up the output spectrum.

The final transmit amplifier is an Avantek MSA-0885, chosen for its gain and +10-dBm power-output capability at 3456 MHz. One disadvantage of the '0885 is that it is only conditionally stable: It can oscillate if supply line decoupling is inadequate or if it is terminated in highly reactive loads. U5 should be stable as long as the transmitter-output port is terminated in an impedance close to 50 ohms. If, however, the transverter is used "barefoot" in conjunction with an external antenna relay, I recommend that the dc power to the transmit amplifier section be removed during receive periods so the absence of an antenna load does not cause U5 to oscillate. My prototype units were stable without an output termination, but variations from device to device may cause U5 to oscillate in some transverters. Result: lots of noise in your receiver, making it impossible to hear weak signals!

Receive Mixer and Preamplifiers

The receive mixer (Fig 4) is identical to the transmit mixer. It is preceded by a 5-pole image-stripping filter, FL5, which is necessary to keep the noise energy at the image frequency from being converted to the IF and degrading receiver sensitivity. In a receive application, 20 to 25 dB of image rejection is all that is necessary to ensure that this does not happen. A 5-pole microstrip filter easily provides this much rejection at the 3168-MHz image frequency.

The receiver-preamplifier stages were chosen to have good SWR and low noise figure. Preceding FL5 is an MSA-0185 (U8) that terminates the filter well and has a noise figure of approximately 6.5 dB at 3456 MHz. The front end consists of two MSA-0685 stages (U6, U7). The overall receiver noise figure is about 4.4 dB. An outboard preamplifier can be added to make a state-of-the-art setup. Performance of the transverter is summarized in Table 1.

Construction Hints

The transverter is constructed on 0.031-inch-thick woven Teflon-glass substrate with a dielectric constant of 2.50. The board is double clad with ½-ounce copper. I get my board

Table 1

3456-MHz Transverter Performance

General

- Frequency range: 3456-3460 MHz
- IF range: 144-148 MHz
- Local oscillator: +16 dBm (40 mW) required at 552 MHz
- Power required: 13.8 V dc at 250 mA

Transmitter

- Output power: 10 mW (+10 dBm)
- Spurious rejection: –40 dBc
- IF drive level: 0 dBm max

Receiver

- Noise figure: 4.5 dB
- Image rejection: 25 dB
- Net gain: 15 dB

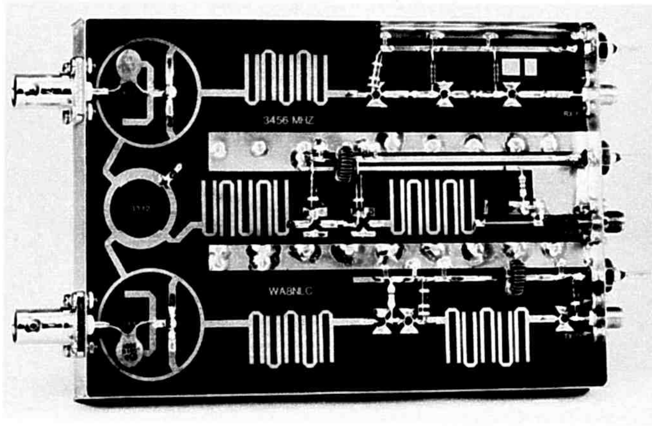


Fig 6—Photograph of a finished 3456-MHz transverter. This unit was built by Zack Lau, KH6CP, in the ARRL lab.

material from Taconics Plastics, Ltd, Petersburg, NY 12138, part number TLX-9-0310-R5/R5. A word of caution: The filters require that dimensional tolerances of +0.001 inch or better be maintained in the fabrication of the board. This is the price you must pay for microwave filters that require no adjustments. Because of the critical tolerances necessary and the many variables involved in the printing process, an etching pattern is not included in this article. You can purchase an etched board from Down East Microwave.[6]

Construction is as simple as populating the board with the components and mounting the connectors. A parts layout guide is shown in Fig 5. Lead dress can be seen in the photograph of the completed transverter (Fig 6). All of the parts needed to complete the project are available from amateur suppliers, and kits are available as well.

Be sure to use high-quality porcelain-chip capacitors for coupling between stages. The remaining capacitors can be less expensive ceramic chips or disc ceramics. I used chip capacitors everywhere I could because they make for a neater layout and cost about the same as the equivalent disc-ceramic capacitors.

Mini-Circuits offers a line of MMICs equivalent to the specified Avantek parts. The corresponding part numbers are easy to determine. The Mini-Circuits MAR-8 is equivalent to the MSA-0885, the MAR-6 to the MSA-0685, and so on.

I strongly recommend that you enclose the board in a brass box to support the connectors and provide ground continuity to the top side of the board. I used strips of 0.032-inch-thick, ¾-inch-wide brass available at most hobby stores. The entire perimeter of the inside walls is soldered to the top and bottom of the board. This provides a ground connection to the component side in several places, as well as a ground for the input and output connectors.

The board material is soft and bends easily. It is quite tolerant of heat and rework if you misplace a part. I use a 27-watt pencil soldering iron with good results.

Through-the-board wires are required to tie components to ground, as shown on the parts layout. Tiny brass rivets can be used if you have them available, but I used no. 18 bus wire with good results. The ground leads of the MMICs are grounded to a small piece of copper foil under each lead. The foil is wrapped through a clearance hole under each MMIC lead and soldered to the ground plane below (see Fig 7). The cold end of the 50-ohm chip resistor (R1) on the LO power divider should be grounded in a similar fashion.

Keep the bias resistors and

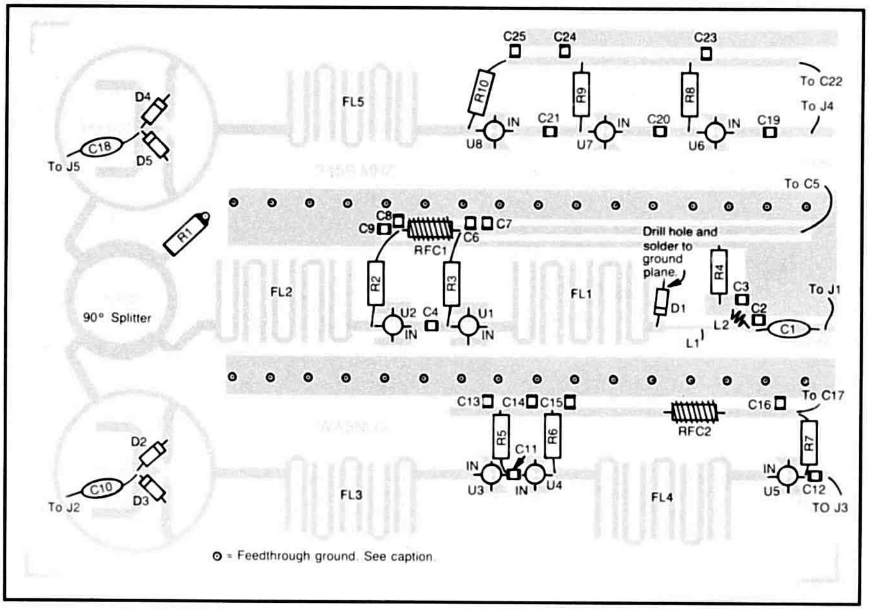


Fig 5—Parts-placement guide for the 3456-MHz transverter (not shown actual size). All components mount on the etched side of the board. Feedthrough grounds, indicated by circles, must be installed and soldered top and bottom; see text. See Fig 7 for mounting details for U1-U8.

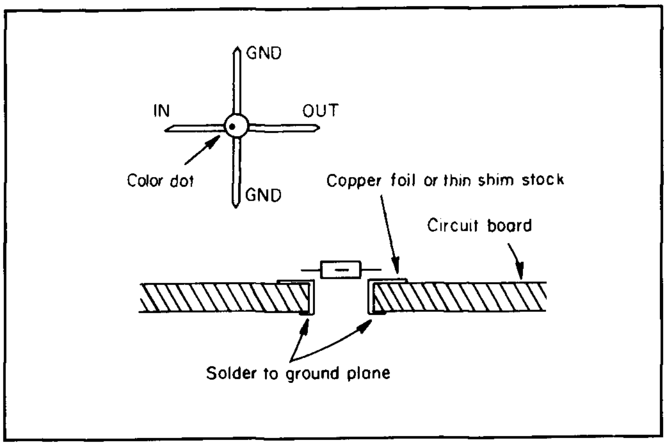


Fig 7—Mounting details for the MMICs.

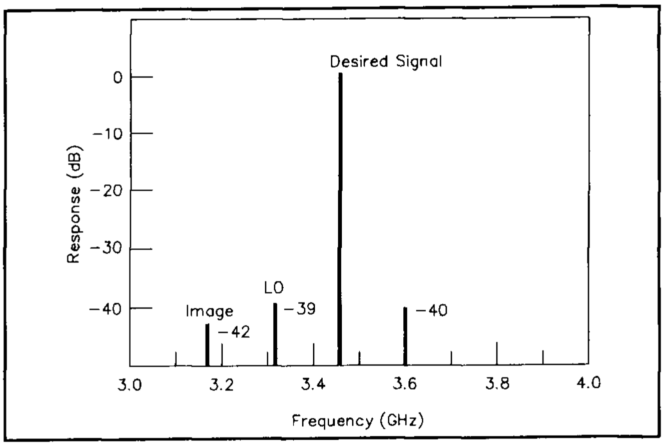


Fig 8—Output spectrum of the 3456-MHz transverter transmitter section.

chokes close to the board so they don't act as antennas. During testing of the prototypes, I noticed that there is some radiation of RF energy from the surface of the board. This is not a problem unless LO energy gets into the transmitter circuits and shows up in the output. Without any shields at all, the LO is more than 35 dB below the carrier.A spectrum analyzer display of the transmitter output is shown in Fig 8.

In the parts list, I have shown bias-resistor values for 13.8-V operation. If only 12-V operation is planned, the bias resistor values should be changed to maintain the correct operating current through each MMIC. Be sure to check resistor dissipation and use ½-W resistors as necessary. The bias current for each MMIC is as follows: MSA-0185, 17 mA; MSA-0285, 25 mA; MSA-0685, 15 mA; MSA-0885, 32 to 35 mA. On one portable expedition, I noticed that the receiver began oscillating after the transverter had been on for a while. I later found that the dry-cell battery had dropped more than 2 V, causing a corresponding drop in the bias current in each MMIC. When operated at lower than rated current, the MMICs are not unconditionally stable. If battery operation is planned, it is a good idea to regulate the power source to the transverter at 10 V or so and adjust the bias resistors accordingly.

For best performance, the mixer diodes should be a matched pair in a microstrip or surface-mount package. The recommended surface-mount pair is the Hewlett Packard HSMS-2822. This part is available from Down East Microwave.[4] If you have some microwave mixer diodes you want to try, go ahead and do it—just make sure the diodes are connected cathode to anode, as shown in Figs 3 and 4. The LO rejection can be measured with a power meter at the transmitter-output port by noting the difference between full carrier output and the residual oscillator level.

Accessories

You will need a few outboard accessories to complete your 3456-MHz station. First, a means of reducing the transmitter output power of your IF rig to 1 mW is required. A recent circuit published in *QEX* not only takes care of attenuating the transmitter, but also provides protected TR switching of the IF line.[7]

On the other end of the transverter, some sort of RF antenna relay is needed. Meeting this requirement may be a little more difficult, because good relays for this band are scarce. I have been fortunate enough to find small SMA-type relays at hamfests in the past, but larger relays with N connectors will probably work. Don't overlook relays equipped with TNC connectors. They are often good to well beyond 3456 MHz. You can take care of any connector mismatch in the jumper cables to the transverter. If you are unsure about the suitability of a relay for this frequency, ask a knowledgeable friend for advice. Remember, just because a relay has good microwave-quality connectors doesn't mean it will provide low loss and operate at an acceptable SWR at 3456 MHz.

If you can find one used, a circulator makes a neat TR switch. Isolation between the transmitter and receiver ports won't be as good with a circulator as with most relays, but at least you won't need the dc source necessary to operate a relay. If you can't find a circulator, you can use an isolator. I have converted many isolators to circulators by removing the isolator load resistor and installing a coaxial connector in its place.

Several types of antennas are popular at 3456 MHz. The most common 3456-MHz antenna is a small dish in the 2- to 4-foot-diameter range. A recent article in *QEX* shows how to build efficient feed systems for the 3.4, 5.7 and 10.3-GHz bands.[8] For respectable performance without the wind load of a dish, you can now buy loop Yagis for 3456 MHz.[9]

Feed-line losses are severe at 3456 MHz. For home-station installations, and even for portable operations, a good Hardline, such as Andrew Heliax, is mandatory. Also, don't overlook the G-line.[10] I have used a G-line for a couple of years now, with very good results.

Station performance can be greatly improved with the addition of an outboard receive preamplifier and transmit power amplifier. A state-of-the-art receiving preamp that makes an excellent front end for this no-tune transverter was described in *QST*.[11] Preamps of this design have been duplicated by many amateurs; they deliver good performance without requiring tweaking on a noise-figure meter.

Several options are available for the transmitter. Some amateurs have found surplus traveling-wave tubes (TWTs) to be a great way of generating lots of power on this band. TWTs typically require only a milliwatt or so of drive for full output, so you will have to attenuate the output of the transverter by about 10 dB. I prefer solid-state, however. A receiving-type GaAsFET like the Avantek ATF-10135 can yield up to +17 dBm quite easily at low cost. Don Hilliard, WØPW, published several good ideas on how to bridge the gap between 10 mW and 1 W.[12] Don shows how, for about $30, you can break the 100-mW level with an Avantek ATF 21170 FET. I have built his circuit using a similar device (an Avantek ATF 25170) with good results.

Summary

What can you expect to work on the 3456-MHz band? A recent article in *QST* discussed the various modes of propagation at 2304 MHz.[13] Everything said about propagation there applies just as well to 3456 MHz from my experience. Stations using dishes have the advantage of increased antenna gains for the same physical size. For example, a 4-foot dish has 3-dB more gain at 3456 MHz than at 2304 MHz. The bottom line is that workable distances on 3456 MHz are on par with those on 2304 MHz. All it takes to prove this is a little more activity!

Notes

[1] A. Ward, "Monolithic Microwave Integrated Circuits," Part 1: *QST*, Feb 1987, pp 23-29, 32; Part 2: *QST*, Mar 1987, pp 22-28, 33.

[2] J. Davey, "Microstrip Bandpass Filters," *Proceedings of Microwave Update '87* (ARRL, 1987), pp 42-53.

[3] J. Davey, "Microwave Filter Update," *Proceedings of Microwave Update '88*, pp 1-8. This book is available from ARRL for $12 (plus $2.50 postage and handling, or $3.50 for insured parcel post or UPS) or from your local dealer.

[4] Complete parts kits for a suitable local oscillator, as well as assembled units, are available from Down East Microwave, Box 2310, RR 1, Troy, ME 04987, tel 207-948-3741.

[5] H. Shuch, "Rat-Race Balanced Mixer for 1296 MHz," *Ham Radio*, Jul 1977, pp 33-39.

[6] Etched boards, complete parts kits and assembled units are available from Down East Microwave (see Note 4).

[7] Z. Lau, "A VHF/UHF/Microwave Transverter IF Switch," *QEX*, Aug 1988, pp 3-4.

[8] D. Hilliard, "Antenna Ideas For 3.5, 5.8, and 10.4 GHz," *QEX*, Jan 1988, pp 3-5.

[9] Loop Yagis are available from Down East Microwave (see Note 4).

[10] R. Dryden, "G Lines for 1296, 2304 and Above," *Proceedings of Microwave Update '87*, pp 54-62. See Note 2.

[11] A. Ward, "Simple Low-Noise Microwave Preamplifiers," *QST*, May 1989, pp 31-36, 75.

[12] *Proceedings of Microwave Update '87*, pp 8-92. See Note 2.

[13] E. Pocock, "Getting From Here to There on 2304 MHz," *QST*, Nov 1988, pp 15-16.

A Single-Board, No-Tune Transverter for 5760 MHz

By Rick Campbell, KK7B

(From *QST*, October 1990.)

Single-board, no-tune transverter designs for the 902, 1296, 2304 and 3456-MHz bands have been published in recent years.[1-4] These boards follow a common theme: They use printed-circuit filters and inexpensive plastic monolithic microwave integrated circuit (MMIC) gain blocks to achieve good performance at low cost. The use of printed-circuit filters and broadband MMICs also eliminates the need for RF alignment or microwave test equipment for proper operation. Low cost and ease of assembly and operation have tempted many amateurs to experiment with the microwave bands.

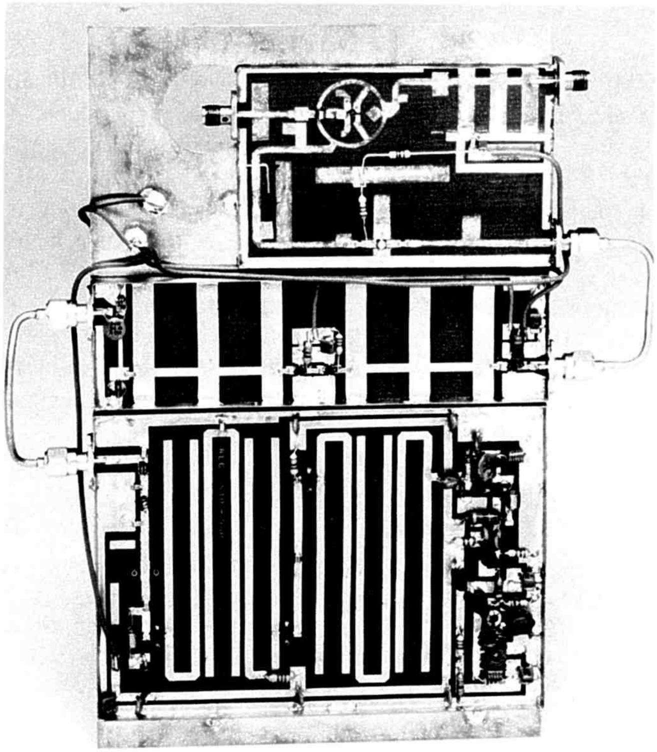
This article describes a single-board transverter for 5760 MHz. Design and construction are similar to that of the 3.4-GHz transverter described by Jim Davey, WA8NLC, in June 1989 *QST*.[5] I won't repeat many of the design and construction techniques described in Jim's article, so a review of that information will be helpful if you're unfamiliar with the single-board transverter concept.[6] In addition to the transverter board described here, you'll need a local oscillator (LO), 1296-MHz IF radio and antenna. Information on completing your 5760-MHz station is presented later.

Design Considerations

In attempting to push the single-board design concept to the 5760-MHz band, two difficulties were encountered:

- The performance of currently available silicon MMICs used in the transmitter and receiver gain stages rapidly deteriorates at this frequency.
- No-tune printed-circuit filters provide insufficient image rejection and LO rejection for use with a 144-MHz IF transceiver.

The problem of image and LO rejection was solved by designing the 5760-MHz board for an IF of 1296 MHz. A single three-section printed-circuit filter with 10% bandwidth provides image and LO rejection of more than 30 dB with a passband insertion loss of less than 1 dB. The problem of obtaining suitable devices for the gain stages was solved by integrating everything except the gain stages into a single circuit board, with each port matched to 50 ohms. Then you are free to use any external gain blocks that may become available.



The KK7B 5760-MHz transverter (top board), along with its companion LO (bottom two boards) offers a no-tune approach to microwave operation.

The basic transverter board—without additional gain stages at 5760 MHz—provides a transmit signal of –6 dBm (250 mW) at the 1-dB compression point (–4 dBm saturated) and a receiver noise figure of 9 dB (assuming that the 1296-MHz IF rig has a 2-dB receiver noise figure). This is acceptable performance for line-of-sight contacts over distances of many miles when used with a small dish antenna. Leaving the gain stages off the transverter board solves another problem: You don't need to worry about finding a suitable 5760-MHz TR relay; the antenna connects to a common filter for transmit and receive.

There are several external gain blocks from which to choose. An excellent one is the 5760-MHz preamplifier described by Al Ward, WB5LUA, in May 1989 *QST*.[7] This preamplifier can also be biased for maximum power and used as a transmit amplifier. Another choice for a transmitter gain block is the Avantek MGA-64135 GaAs MMIC.[8]

Design Notes

Fig 1 shows the block diagram of the 5760-MHz transverter board, and the schematic is shown in Fig 2. There are three basic sections; LO doubler, mixer and bandpass filter.

A 4464-MHz LO signal is mixed with a 1296-MHz IF signal for operation at 5760 MHz. The transverter requires an external 2232-MHz LO; an on-board MMIC doubles this signal to 4464 MHz at +8.5 dBm for mixer injection.

Although you can use any 2232-MHz source, I highly recommend the no-tune LO system described in July 1989 *QST*.[9] The no-tune LO with a 93-MHz crystal delivers a 2232-MHz, +7-dBm signal.

An on-board doubler using an Avantek MSA-0835 MMIC provides an output of +8.5 dBm at 4464 MHz, with all other outputs more than 30-dB down. The second harmonic of the drive signal is obtained by overdriving the MMIC amplifier. The harmonic output is increased by reducing the MMIC bias below the value recommended for linear operation. In the reduced-bias condition, the MSA-0835 will oscillate if the drive signal is removed. A 2232-MHz drive signal of −8 dBm or more is sufficient to stabilize the MSA-0835, and the 4464-MHz output varies little for drive levels between +3 and +10 dBm. The MSA-0835 doubler should be driven by a broadband, flat 50-ohm source. The MMIC in the output of the no-tune 2232-MHz LO is ideal.

The output of the doubler passes through a 2232-MHz notch filter (L4 on the schematic) and a single-tuned circuit at 4464 MHz (L5 and L6). L4 is an open-circuit ¼-λ transmission line at 2232 MHz, so it appears as a short circuit at the drive frequency. L4 is an open-circuit ½-λ transmission line at 4464 MHz, so it appears as an open circuit at the desired second harmonic. The open circuited end of L4 is one-half wavelength away from the output of the MSA-0835 at 2232 MHz and one wavelength away at 4464 MHz, so the MSA-0835 does not see a short circuit at either frequency.

The MSA-0835 is a broadband amplifier, with a noise output from DC to more than 6 GHz. If the MSA-0835 noise output at 1296 MHz is passed to the mixer, some of it will appear at the IF port. This noise will directly add to the receiver noise figure. The broadband 4464-MHz single-tuned circuit (L5 and L6) was added to the artwork to stop the 1296-MHz noise component from getting to the mixer. Careful measurements with the single-tuned circuit indicate that the noise figure is approximately equal to the mixer conversion loss. Without L5 and L6, the noise figure is about 10-dB worse.

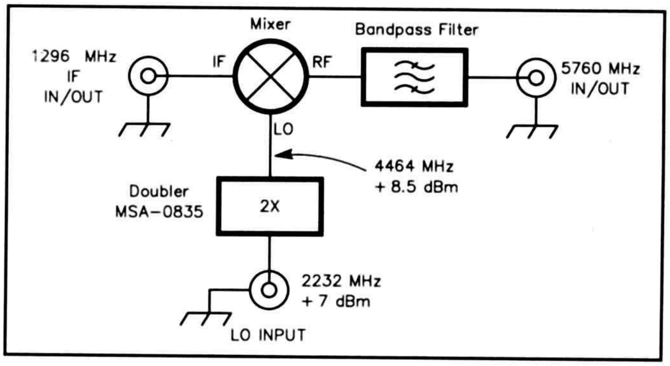
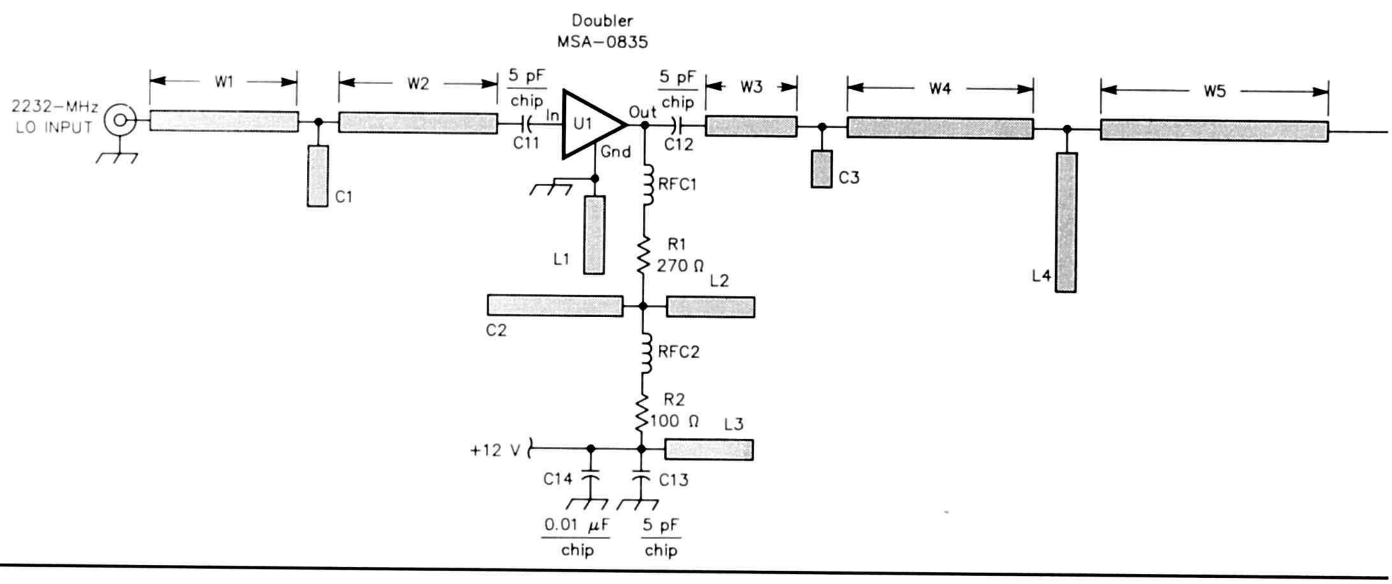


Fig 1—Block diagram of the 5760-MHz transverter.

Fig 2—Schematic of the 5760-MHz transverter. R1 and R2 are ¼-W carbon-film types.

C1-C10—Capacitive stub printed on circuit board.
C11-C13—5-pF, 50-mil-square microwave chip capacitor.
C14—0.01-mF chip capacitor.
C15—18-pF microwave chip capacitor.
D1—HP HSMS-2822 diode assembly.
L1-L3—¼-λ open-circuit stub etched on PC board.
L4—¼-λ stub at 2232 MHz (and ½-λ stub at 4464 MHz) etched on PC board (see text).



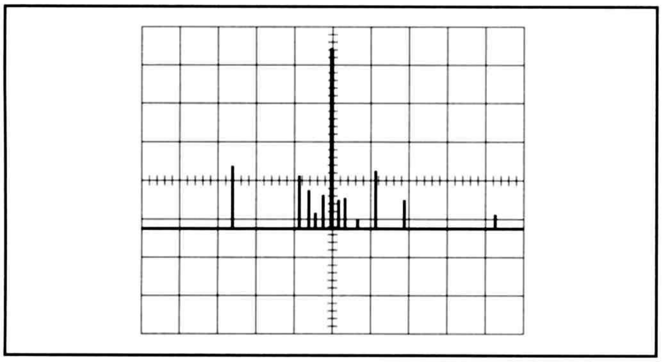


Fig 3—Output spectrum of the 5760-MHz transverter. Each vertical division is 10 dB; each horizontal division is 500 MHz. The pip at the center of the display is the 5760-MHz signal. All other outputs are more than 30-dB down.

When properly driven, this frequency doubler is well behaved, clean and stable. The six versions constructed to date have shown less than 0.5-dB variation in output level when driven from the same 2232-MHz no-tune LO.

The mixer is a standard design, except that the HP-2822 diode pair is used well above its specified frequency range. On the assumption that the imperfections of the diodes were reactive and relatively uniform from part to part, I built a standard 6/4-λ mixer, then added tuning "confetti" empirically to improve the conversion loss. The confetti was then added to the mixer artwork, and subsequent mixers show good uniformity.

The IF port is also empirically matched to 50 ohms at 1296 MHz. The bare mixer displays a conversion loss of only 6.5 dB at 5760 MHz, which is much better than many commercial mixers designed for broadband performance in this range.

The filter is an off-center-tapped 0.16-dB-ripple Chebyshev type designed using the procedure described by Beebe.[10] It is exceptionally flat, has steep skirts and a loss of well under 1 dB in the passband.

After I integrated the doubler, mixer and filter onto a single board, the interconnecting line lengths and tuning confetti were varied to obtain best performance with 50-ohm pads on all ports. The output spectrum at the 1-dB compression point is shown in Fig 3. The image, which is 32-dB lower than the 5760-MHz output, is off the lower end of the spectrum-analyzer range. Table 1 lists the transverter specifications.

One cautionary note: Any metal box big enough to hold the transverter board will be a resonant cavity at a series of frequencies within the gain bandwidth of the MMIC. Imagine

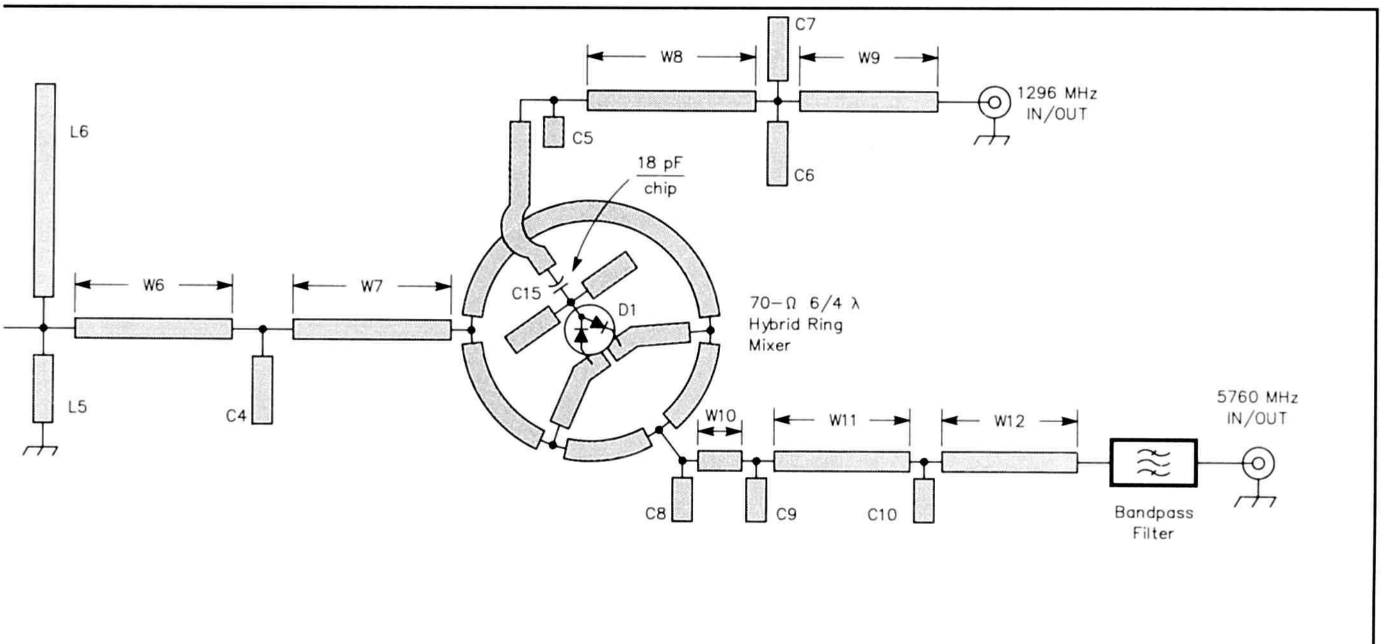
Table 1

Transverter Specifications

General
- Frequency range: 5650-5925 MHz
- IF range: 1240-1300 MHz
- LO required: 2.18-2.32 GHz at +7 dBm nominal; +3 to +10 dBm acceptable
- Power requirements: 12 V dc at 10 mA; 8-15 V acceptable.

Transmitter
- Output power at 1-dB compression: −6 dBm.
- Saturated RF output: −4 dBm.
- LO signal at RF port: −36 dBm, max.
- IF drive level: approx 0 dBm.

Receiver
- Noise figure: 9 dB.
- Conversion loss: 7 dB.
- Image rejection: 30 dB.



L5, L6—Broad tuned circuit at 4464 MHz (and short circuit at 1296 MHz) etched on PC board (see text).
RFC1, RFC2—Inductance of R1 and R2 leads (not critical).
U1—Avantek MSA-0835 MMIC.
W1-W12—Tuned lengths of 50-Ω microstrip transmission line etched on PC board.

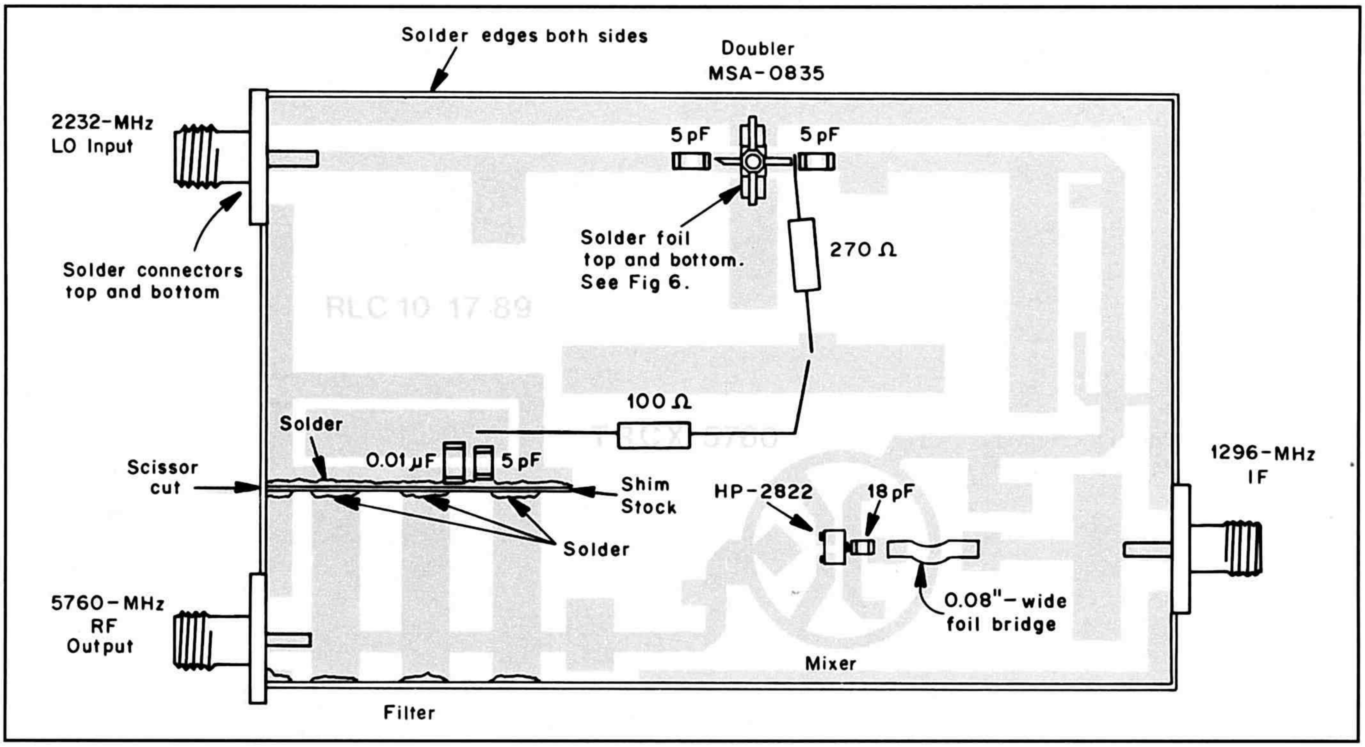


Fig 4—Part-placement diagram for the 5760-MHz transverter board (not shown actual size). All components mount on the etched side of the board. See Fig 6 for mounting details for the MSA-0835 MMIC.

building a 40-meter linear amplifier in a box the size of a football field! No-tune transverters and LOs that work perfectly on the bench often exhibit spurious oscillations when enclosed in a metal box. The transverter board will work fine if it is simply wrapped with plastic-bubble packaging material or nonconductive foam and mounted near the antenna feed. For permanent installations, a plastic food container works well.

If a shielded metal box is necessary, I recommend shielding individual stages with thin brass or copper shim stock and then enclosing the shielded individual stages in a larger box. Spurious oscillations in the no-tune LO or lower-frequency transverters may often be cured by replacing "hot" MMICs like the MSA-0835, MSA-0685 and MAR-6 with well-behaved parts like the MSA-0235, MSA-0285 and MAR-2.

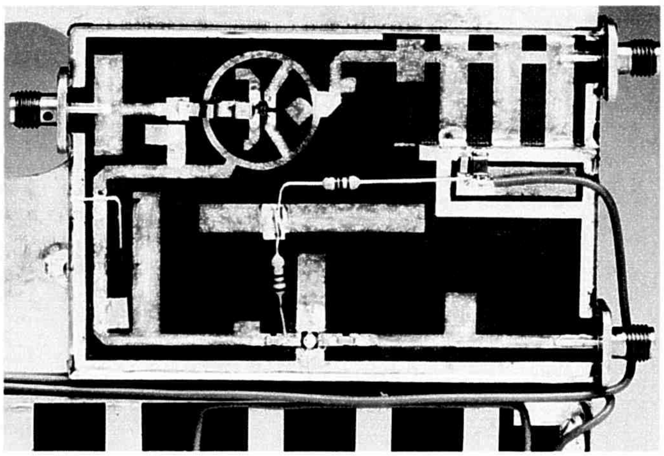


Fig 5—A finished 5760-MHz transverter.

Construction

The transverter is constructed on 0.031-inch-thick Teflon-glass substrate with a dielectric constant of 2.5. The board is clad with ½-ounce copper on both sides. One side is etched; the other is unetched and acts as a ground plane. The material I used is made by Taconics Plastics, Ltd, Petersburg, NY 12138, and the part number is TLX-9-0310-R5/R5. The filter requires that dimensional tolerances of ±0.001 inch or better be maintained in the fabrication of the board. (Close construction tolerances are essential for microwave filters that require no adjustments.) Because of the critical tolerances necessary and the many variables involved in the *QST* printing process, an etching pattern is not included in this article. If you are interested in making your own board, send an SASE to the ARRL Technical Department for a dimensioned copy of the artwork.[11] Or, if you wish, you can purchase an etched board from Down East Microwave.[12]

Part placement is shown in Fig 4. All components mount on the etched side of the board, and construction is conventional for microwave circuits. Additional details may be seen in Fig 5.

Brass sides enclose the board to support the SMA connectors and provide continuity between the grounds on the top and bottom of the board. The sides are made from strips of 0.032-inch-thick, ¼-inch-wide copper or brass shim stock available at hobby stores. The perimeter of the inside walls is soldered to the top and bottom of the board. This provides a ground connection to the component side in several places, as well as a ground for the connectors.

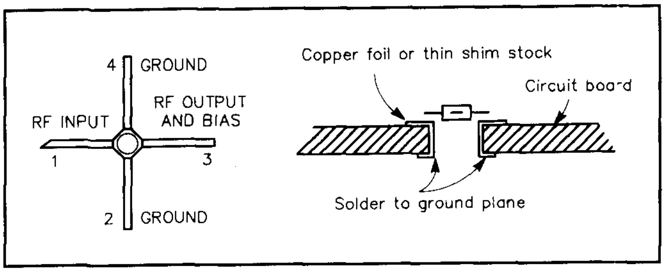


Fig 6—Mounting details for the MSA-0835.

Additional grounding is needed for the MSA-0835. Cut a hole in the board as shown in Fig 6, and wrap copper foil through the hole to connect grounds on the top and bottom of the board. Then solder the foil on both sides.

Both ends of the filter elements must be grounded. The ends of the elements along the edge of the board are grounded simply by soldering them to the vertical board edge strips, which must also be soldered on the bottom. The ends of the elements toward the center of the board are grounded by making a straight scissor cut as indicated by the arrows, and inserting a strip of thin copper shim stock the full length of the slit. Then solder the shim stock to the PC-board copper on both sides of the slit, on the top and bottom of the board.

Additional copper foil is used to make a bridge between the 1296-MHz IF connector and the mixer. Cut the foil to 0.08-inch wide and solder one end to the printed trace from the IF connector. Then bend the foil into an arch over the printed mixer ring and solder it to the pad leading to the 18-pF coupling capacitor.

Completing Your 5760-MHz Station

In addition to the transverter and LO boards, you'll need an IF radio and antenna. There are two convenient approaches to the IF rig:

- Use a 1296-MHz multimode transceiver (ICOM, Kenwood and Yaesu make suitable radios).
- Build a 1296-MHz no-tune transverter board[13] to use as a first IF, and use a 2-meter multimode radio as the IF transceiver.

In either case, it's worth a look at Zack (KH6CP) Lau's transverter control circuit described in August 1988 *QEX*.[14]

Selecting an antenna for 5760 MHz is easy: Nothing matches the performance of a small dish. I've had good success with a 19-inch dish that mounts easily on a small camera tripod. It shows a 10-degree beamwidth and about 25 dB gain with a simple feed. Don Hilliard, WØPW,[15] and Tom Hill, WA3RMX,[16] have presented some excellent antenna ideas for 5760 MHz.

Conclusion

This transverter board, used with the no-tune 2232-MHz LO, provides a straightforward approach to a transceive system on the 5760-MHz band. Although it's not as attractive as the previously described transverters for 902 through 3456 MHz in terms of output level, noise figure and the use of a 144-MHz IF, it is by far the simplest board in the family. Construction of a high-performance transceive system should be relatively easy for an experimenter ready to move up to the 5760-MHz band.

Acknowledgments

I thank Tony Bickel, K5PJR, Merle Cox, W7YOZ, and Jim Davey, WA8NLC, for inspiration and encouragement, and Mark Schreiner, NK8Q, Dave Erickson, N9JBI, and Jim Berner, K8OSF, for help in evaluating the prototype 5760-MHz transverters.

Notes

1 R. Campbell and D. Hilliard, "A Single Board 900 MHz Transverter with Printed Bandpass Filters," *Proceedings of Microwave Update '89*, pp 1-8. (Reprinted in this book.)

2 R. Campbell, "A Single-Board No-Tuning 23-cm Transverter," *Proceedings of the 23rd Conference of the Central States VHF Society*, pp 44-52. (Reprinted in this book.)

3 J. Davey, "A No-Tune Transverter for 3456 MHz," *QST*, Jun 1989, pp 21-26. (Reprinted in this book.)

4 J. Davey, "No-Tune Transverter for 2304 MHz," *Proceedings of Microwave Update '89* (ARRL, 1989), pp 30-34. (Reprinted in this book.)

5 See Note 3.

6 Copies of this article are available from the ARRL Technical Department Secretary for $3 and an SASE.

7 A. Ward, "Simple Low-Noise Microwave Preamplifiers," *QST*, May 1989, pp 31-36, 75.

8 Contact Avantek, 3175 Bowers Ave, Santa Clara, CA 95054, tel 408-727-0700, for the name of your local dealer.

9 R. Campbell, "A Clean, Low-Cost Microwave Local Oscillator," *QST*, Jul 1989, pp 15-21.

10 G. Beebe, "Analysis of a Class of Microstrip Bandpass Filters," MSEE thesis, Michigan Technological University, Feb 1988.

11 Send a no. 10 SASE to the ARRL Technical Department Secretary; request the Oct 1990 *QST* 5760-MHz transverter template.

12 Etched circuit boards, parts kits and assembled and tested circuit boards for this project are available from SHF Systems through Down East Microwave, Box 2310, RR 1, Troy, ME 04987, tel 207-948-3741. Circuit boards for the other transverters and the no-tune 2232-MHz LO referred to in the text are available from the same source.

13 See Note 12. Also see Z. Lau, "Product Review: SHF Systems 1240K 1296-MHz Transverter Kit," Feb 1990 *QST*, pp 33-34.

14 *QEX*, Aug 1988, pp 3-4.

15 D. Hilliard, "Antenna Ideas for 3.5, 5.8, and 10.4 GHz," *QEX*, Jan 1988, pp 3-5.

16 T. Hill, "A Triband Microwave Dish Feed," *QST*, Aug 1990, pp 23-27.

10-GHz Gunnplexer Communications

By ARRL Staff

The M/A-COM Gunnplexer transceiver shown in Fig 1 will fit conveniently in your hand and is operated from a single 12-V power supply, either ac line operated or batteries! This makes the Gunnplexer ideal for fixed or portable operation.

The Gunnplexer is most often used in wide-band FM systems and is ideal for audio, full-color video and data transfer. With suitable peripheral equipment, systems for full-duplex audio, color video and up to several megabit data transfer are possible. Such high data rates should allow direct computer-to-computer memory transfer that would not be practical (or legal) on other lower-frequency amateur bands.

The Gunnplexer

The heart of the Gunnplexer is a Gunn diode oscillator, named after its inventor, John Gunn of IBM. Refer to the cutaway drawing of the Gunnplexer, Fig 2, for this discussion.

The Gunn diode is mounted with a varactor diode in a resonant cavity. When a regulated voltage is applied to the Gunn diode it oscillates; the frequency of oscillation is determined by the capacitance of the varactor diode and two mechanical tuning screws. The mechanical tuning screws can be likened to coarse tuning controls and are factory set for the appropriate tuning range. The voltage applied to the varactor diode (1 to 20 V dc) tunes the frequency electronically a minimum of 60 MHz. Power is coupled out of the cavity through a small iris that has been designed as somewhat of a compromise between maximum power output and isolation from changes in diode impedance and load.

The Gunn oscillator is also used to provide the local-oscillator signal for the detector diode. A ferrite circulator couples an appropriate amount of energy into the low-noise Schottky mixer diode and isolates the transmitter and receiver. Because the Gunn oscillator functions as both the transmitter and the receiver local oscillator, the IF at each end must be at the same frequency. Furthermore, the frequencies of the Gunnplexers must be separated by the IF. This is illustrated in Fig 3. Intermediate frequencies of 30 MHz are more or less standard for audio work in the US. Both 45 and 70 MHz are used for video and high-speed data work.

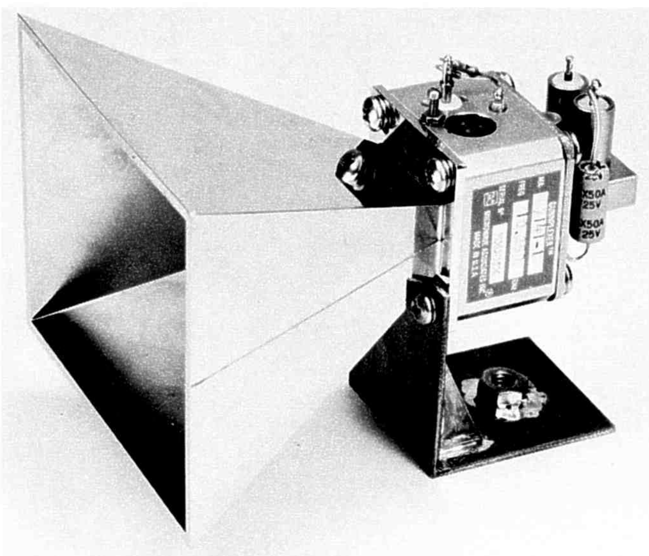


Fig 1

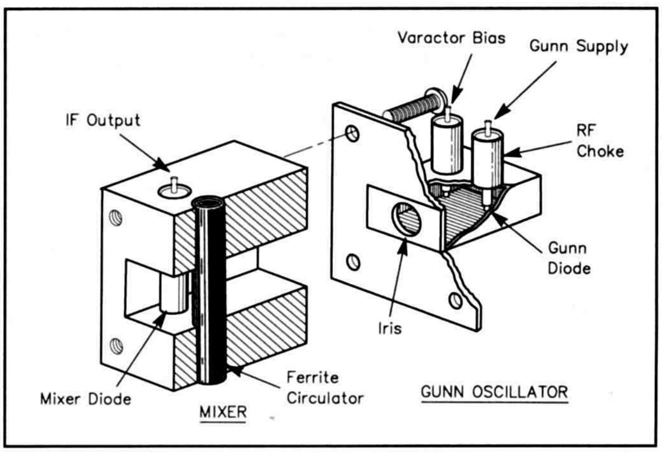


Fig 2

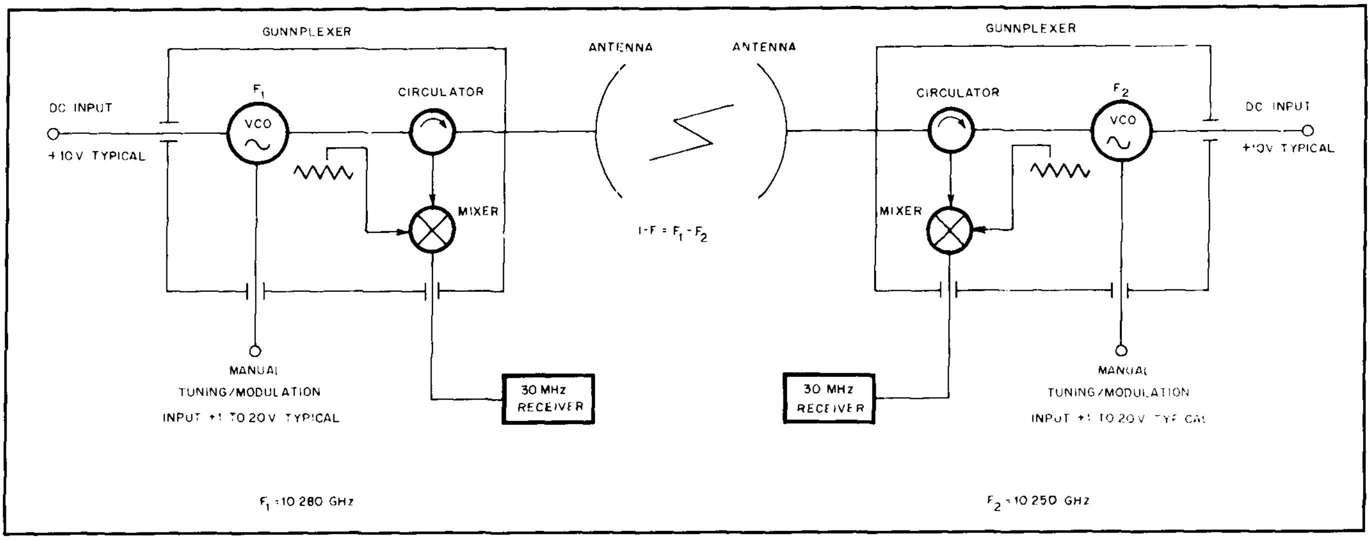


Fig 3

As can be seen from Fig 3, the Gunnplexer communications system is full duplex. In other words, both parties can talk and listen at the same time, without throwing switches. This is something that may take a while to get used to as most amateurs are programmed for VOX or PTT operation. In short, it is the ultimate break-in system!

One detail of the Gunnplexer that does require some specific attention is frequency control. The Gunnplexer has a frequency stability specification of –350 kHz frequency change per degree Celsius increase. This does not pose much of a problem with wide-bandwidth applications such as video or data transfer. For relatively narrowband audio work (200 kHz and less), however, some form of AFC, a phase-lock or other frequency-control scheme is required. In most cases simple AFC circuitry is sufficient and quite simple to implement. The electrical characteristics of the Gunnplexer are given in Table 1.

Bare Minimum Audio Communications System

The simplest of communications systems using the Gunnplexer transceivers can be formed with two 88- to 108-MHz FM receivers, two Gunnplexers, two microphones and associated amplifiers, and two sources of 12 V dc. A diagram of such a system is shown in Fig 4. Some of the low-cost FM converters and receivers for automotive use make good IF strips for this communications system. The AFC signal developed in the converter or receiver can be routed to the Gunnplexer varactor diode to lock the two units together. The microphone amplifier can be a single 741 op amp as shown in the diagram.

There are two shortcomings with this system. The first involves the use of the FM broadcast band as the IF. If mountaintop DXing is planned, it is likely that strong FM broadcast stations will be received no matter how short the lead between the detector and the FM converter or receiver is made. It will be necessary to select that part of the band where there are no strong signals present.

Table 1
Gunnplexer Specifications @ T_A = 25° C

Electrical Characteristics	
RF Center Frequency	10.250 GHz[1]
Tuning	
Mechanical	± 50 MHz
Electronic	60 MHz min.
Linearity	1 to 40%
Frequency Stability	–350 kHz/°C max.
RF Power vs Temperature and Tuning Voltage	6 dB max.
Frequency Pushing	15 MHz/V max.
Input Requirements	
Dc Gunn Voltage Range	+ 8.0 to 10.0-V dc[2]
Maximum Operating Current	500 mA
Tuning Voltage	+1 to +20 volts
Noise Figure[3]	<12 dB
RF Output Power[1]	
Model	*P out (mW)*
MA87141-1	10 min. 15 typ.
MA87141-2	20 min. 25 typ.
MA87141-3	35 min. 40 typ.

Notes
[1]Tuning voltage set at 4.0 volts.
[2]Operating voltage specified on each unit.
[3]1.5 dB NF at 30 MHz.

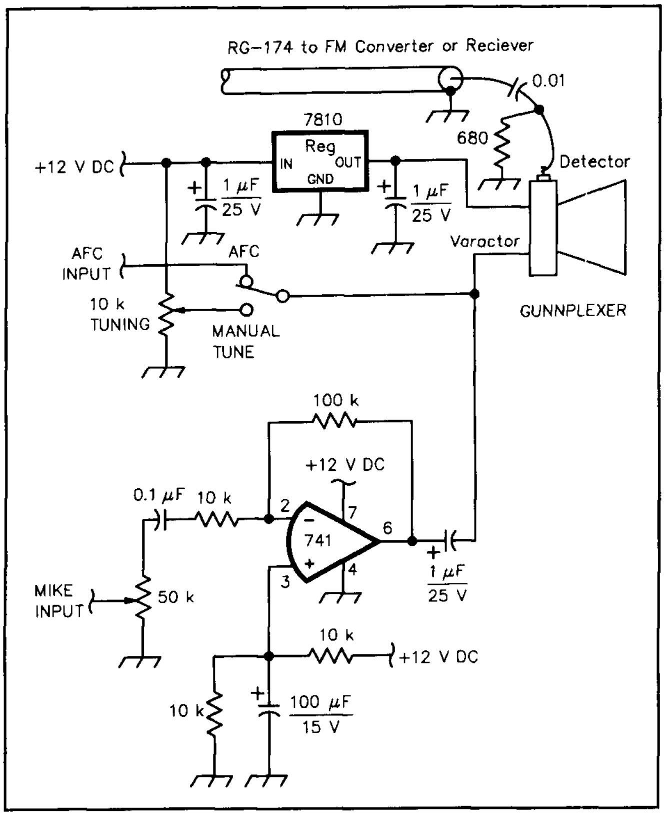


Fig 4

The second item involves AFC. Since it is possible to tune the Gunnplexers over a 60-MHz range with the varactor diode tuning, it is possible to tune them on either side of each other. This means that the single-polarity AFC system in the receiver or convertor will work for only one combination. If the units are operated so that the AFC polarity is incorrect, the AFC will push the received signal out of the receiver passband. If this happens, simply tune the two Gunnplexers to produce the other IF signal. AFC lock should then be obtained.

Modifications for the SSB Electronics 10 GHz Transverter

By Kent Britain, WA5VJB

(From *Microwave Update '91*)

The SSB Electronics 10-GHz Transverter is a truly state-of-the-art piece of equipment. They were the first commercial 10-GHz transverter. There are, however, several areas where these units can be improved.

Power Leads

All three boxes use insulated pins for the 12-V power connections. Replacing the pin with a feedthrough capacitor makes the box RF tight. This really helps in the high RF environment found on many hill tops. I also suggest adding a diode from the power lead to ground in each box. Use a 1N4001 or a 15-V, 1-W Zener. This adds reverse polarity protection for those little accidents that occur in the field (Fig 1).

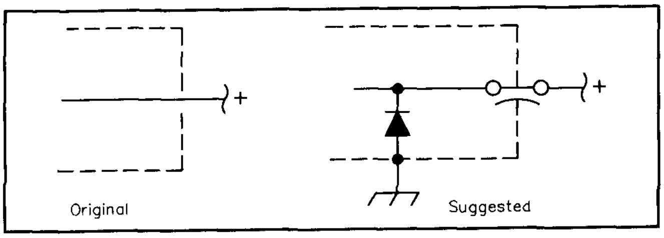


Fig 1—Reverse-polarity protection.

Image Filtering

The basic unit has very little rejection of the 10.080-GHz image. While there are several filters in the transmitter, there is only one filter in the receiver. With very careful tuning, I was only able to get 8 dB of image rejection. Sweeping with a signal generator, I was also able to get responses on six frequencies between 10 and 11 GHz.

These extra images have resulted in the publication of some very optimistic noise figure measurements. Typically, the noise figure runs about 3-½ dB for the basic unit. An external preamp and 10-GHz filter will really clean up the receiver. After adding a pair of ½-in. plumbing end cap "cheap" filters, all spurious responses were more than 40-dB down. WA5TNY and W7CNK are using waveguide filters ahead of their transverters (Fig 2).

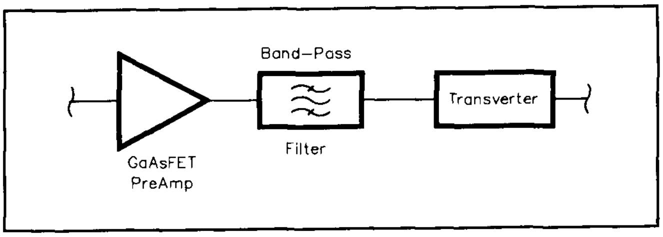


Fig 2—An external preamp and filter will remove receiver spurs.

The Local Oscillator

The LO module puts out 6 to 7 mW at 2.556 GHz. If the LO output power ever drops below 5 mW, the 2.556 to 10.224 GHz multiplier stops multiplying. The LO tuning can be quite critical in the 2.556-GHz filter stage. The filter depends on self resonance in the three trimmer caps to form the filter.

If you simply tune the LO for max output, you'll see a pretty dirty output on a spectrum analyzer. Tune for cleanest response, and you may not get the 6 mW the next stage needs. Several stations have solved this problem by completely replacing the LO module. Using a Frequency West Brick crystaled for 2.556 GHz, or even 10.224 GHz (bypassing the quadrupler stage) has been popular.

Another solution is to add a filter-amp stage between the LO and receiver modules. Replacing the jumper cable with an MSA-0404, MAR-4, or MSA-1105 MMIC, followed by a plumbing cap or interdigital filter can really clean up the output and produce plenty of drive for the next stage.

Buffer Amp

The output of the 10-GHz GaAsFET receiver mixer goes directly to the IF output. The slightest accidental transmission back into the receiver is *fatal* to this GaAsFET. Many of the new solid-state switching 2-meter rigs can actually start transmitting faster than a relay can close. Installing a buffer amp and resistive pad on the IF output can protect the receiver while providing enough gain to drive secondary receivers, moon/sun-noise meters, or long IF cable runs. On my transverter an 18-dB pad is permanently mounted on the IF

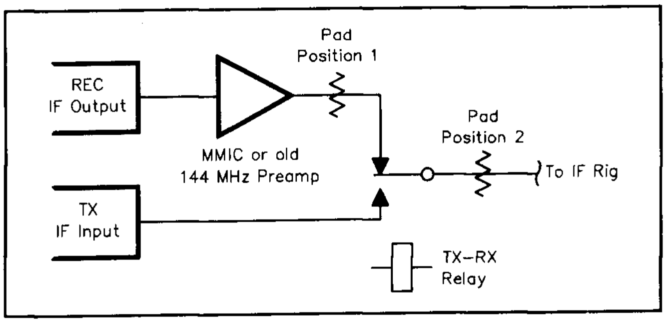


Fig 3—A buffer amp and attenuator protect the receiver from the IF rig transmitter.

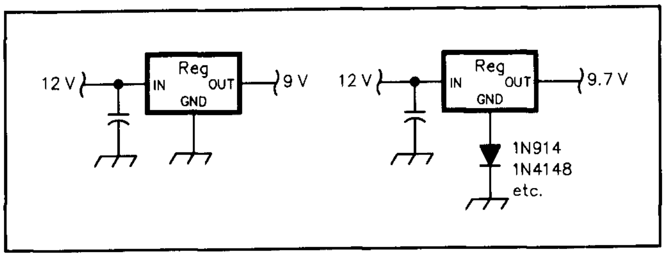


Fig 4—Adding a diode in the GND leg of the regulator raises LO supply voltage, improving stability and increasing drive.

connection and is used during both transmitting and receiving (Fig 3).

Transmitter Module

The output of the transmitter usually exceeds SSB's specs, and is quite clean. The only problem was driving a TWT to saturation. When a TWT is driven to saturation to get that last watt out for EME, the TWT is often driven to 15 dB of gain compression. This makes any spurs or images 15 dB stronger at the TWT output. An extra 10.368-GHz filter between the transverter and the TWT keeps all those precious watts on 10.368 MHz.

Increasing LO Output and Stability

Allen Katz, K2UYH, was also having problems with LO drive and poor stability. His solution was to add a diode in the ground leg of the 9-V regulator in the LO module. This increased the oscillator and multipliers operating voltage from 9 to 9.7 V. Allen says this really helped (Fig 4).

Conclusion

The 10-GHz SSB Transverter is an excellent piece of gear. Most of these modifications attach between units. The mods replace the normal jumper cables, so modifications to the SSB boxes themselves are not necessary.

SSB/CW Equipment Concepts for 24 and 47 GHz

By Tom Hill, WA3RMX

(From *Microwave Update '89*)

INTRODUCTION

As interest and activity increases on the microwave bands, the relentless push for ever-higher frequencies leads us up into the so-called Millimeter Wave bands. In the past, most work at these frequencies has been done using wideband FM, as the stability of the available components did not allow for narrowband work. Today the technology for better frequency stability and phase noise is becoming more available within the amateur community, and we are beginning to take some advantage of it. This article shows the techniques and equipment I have used at 24 and 47 GHz. It is not a construction article, as it does not contain complete construction details, but it does present the concepts I used in my implementation of Millimeter Wave SSB and CW. These ideas may be useful as you venture higher in frequency.

MODULATION MODE

If we exclude, for now, the more exotic synchronous detection and DSP signal recovery methods as being beyond the scope of this project, then we are left with CW as the

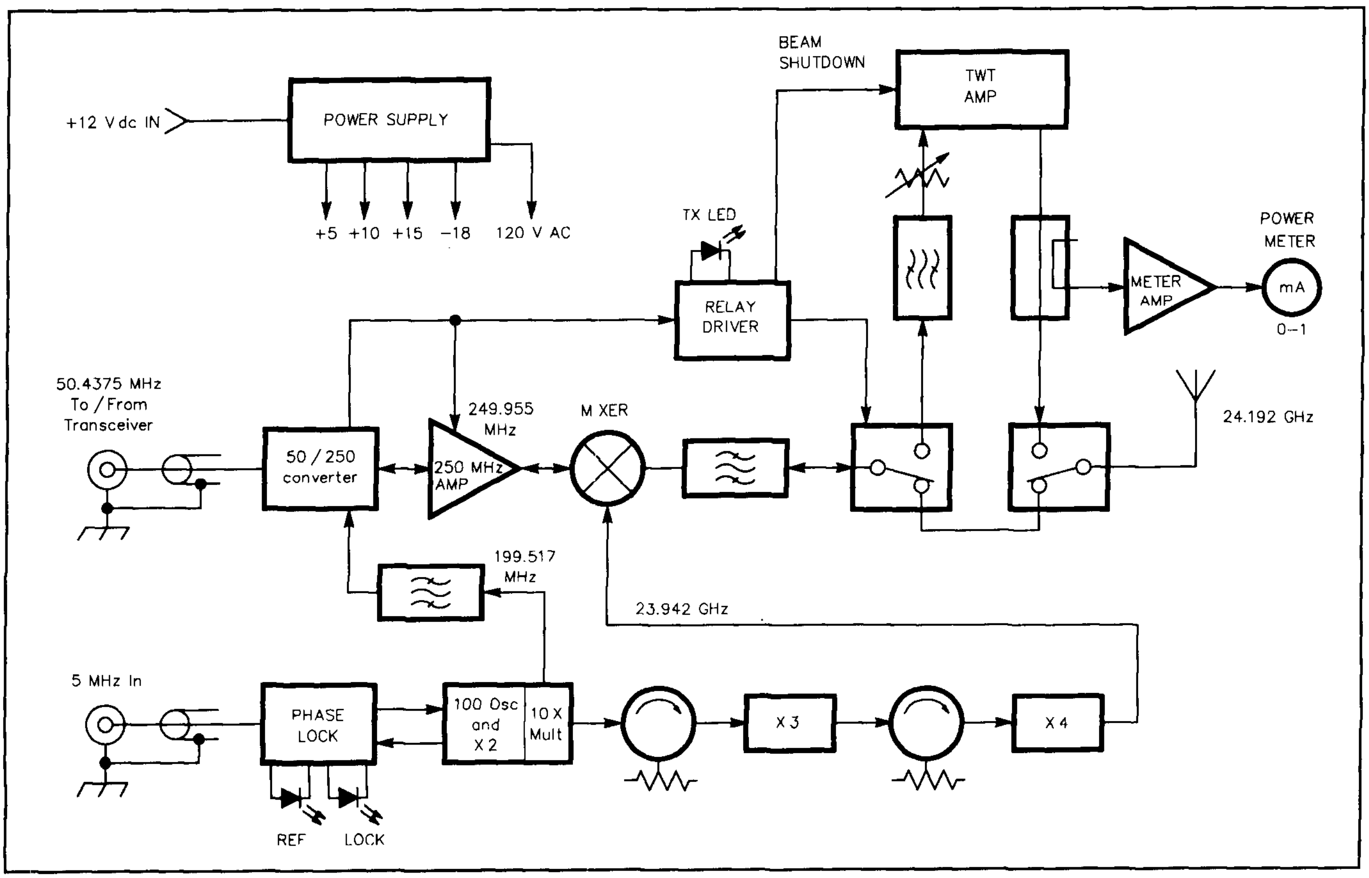


Fig 1—Block diagram of 24-GHz transverter with 6-meter IF.

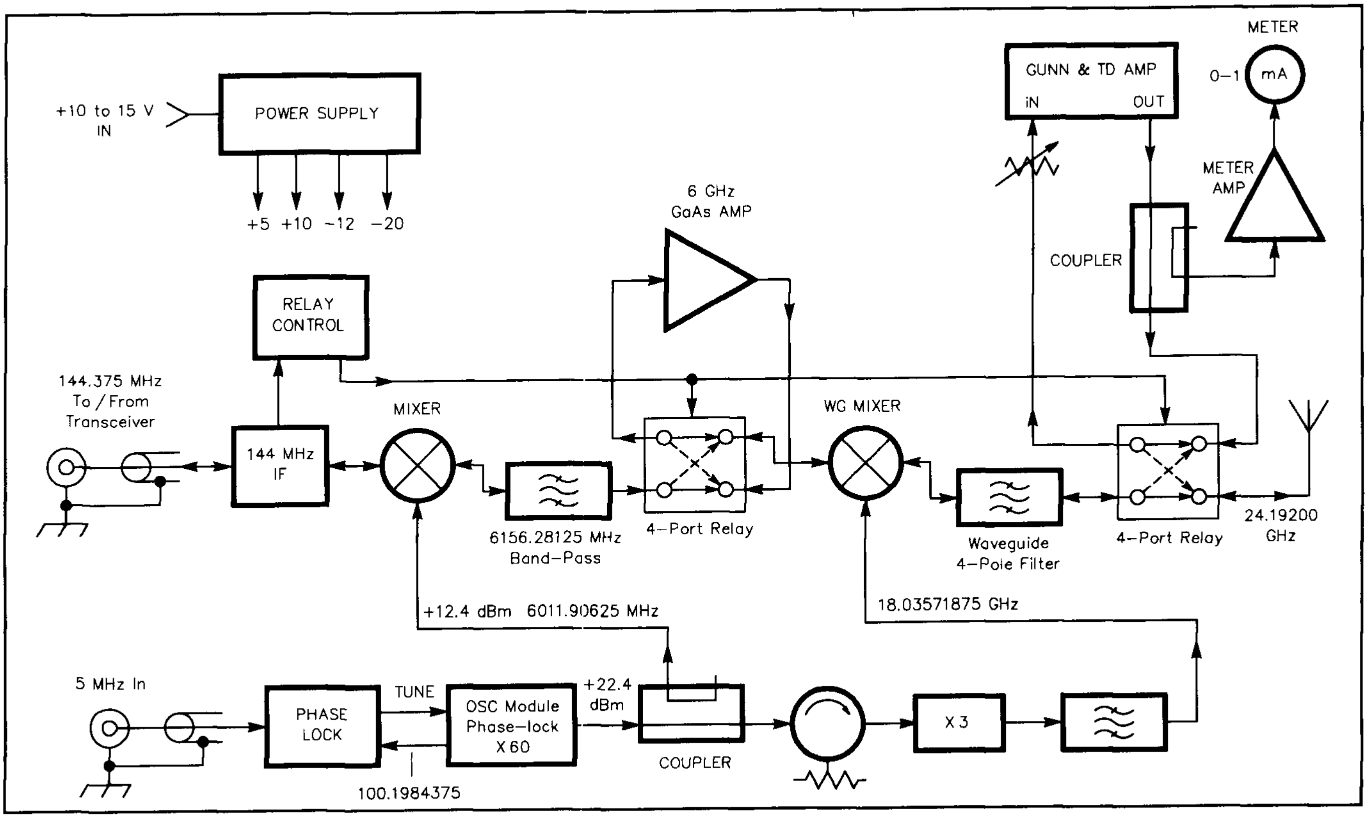


Fig 2—Block diagram of 24-GHz transverter with 2-meter IF.

modulation/detection scheme most likely to recover the weakest signal. It is important to consider that weak-signal aspect, as the atmospheric losses at these frequencies are fierce and the attainable transmitter powers are miniscule at best. The receiver for CW will do SSB for free, so this is a good choice when the signals are strong.

A linear transverter allows the most modulation flexibility, and has only a small penalty in output below using the LO chain itself as a CW transmitter. Even this limitation can be overcome today, since surplus TWTs (Traveling Wave Tubes) are available for 24 and 47 GHz.

THE TRANSVERTERS

24 GHz

My first millimeter transverter was for 24 GHz. The block diagram is shown in Fig 1. It was built as an extension of a previously done 10-GHz rig. I used what I had available at the time, which was an NEC digital telephone-link radio. The original local oscillator chain was the useful part, and I modified the last multiplier to work at 4X instead of the original 2X. With a 6-meter last IF, the existing LO chain provided for a mid IF at 250 MHz, which was barely high enough for the waveguide filter to lower the LO feedthrough at the RF output to an acceptable level.

The TWT amplifier is contained within the same cabinet with the rest of the system. The various blocks are essentially the same as the later 47-GHz rig, and so for a more complete description, refer to the circuit descriptions which are for the 47-GHz rig.

With the growing availability of the 6-GHz phase-locked oscillator modules (CMI, CTI, and others), it seems desirable to use these as they provide a simpler, smaller, lighter, and more cost-effective solution. I have designed such a rig, and the block diagram for this is shown in Fig 2.

47 GHz

There are two separate rigs for 47 GHz. This is so that I will have someone to talk to. The two differences between these two are the different IF—6 meters for one, and 2 meters for the other—and the fact that the one has the TWT built in, while the second has the TWT amplifier built as an external accessory. Since they are essentially the same, I will present only the one with the separate TWT. The block diagram for the transverter is Fig 3, and the amplifier is shown in Fig 4. In this version, the last IF is 2 meters, and the high IF is near 2 GHz.

DUAL CONVERSION

The choice of IF is often influenced by what SSB radio you may have available. On the lower bands this will likely be the only IF used. But with 24 GHz and up it is usually desirable to have a high IF as well, with a dual-conversion rig the result. This is necessary because the filter used at the operating frequency must remove both the image frequency and the LO. Since it is often difficult to obtain fully balanced mixers for these frequencies, the LO power coming out of the mixer along with the desired signal may be 10-dB stronger than the desired signal. This places quite a strain on the design of the output filter if the IF is low. The original 24-GHz rig used a 250-MHz high IF, which is quite marginal. Although the image conver-

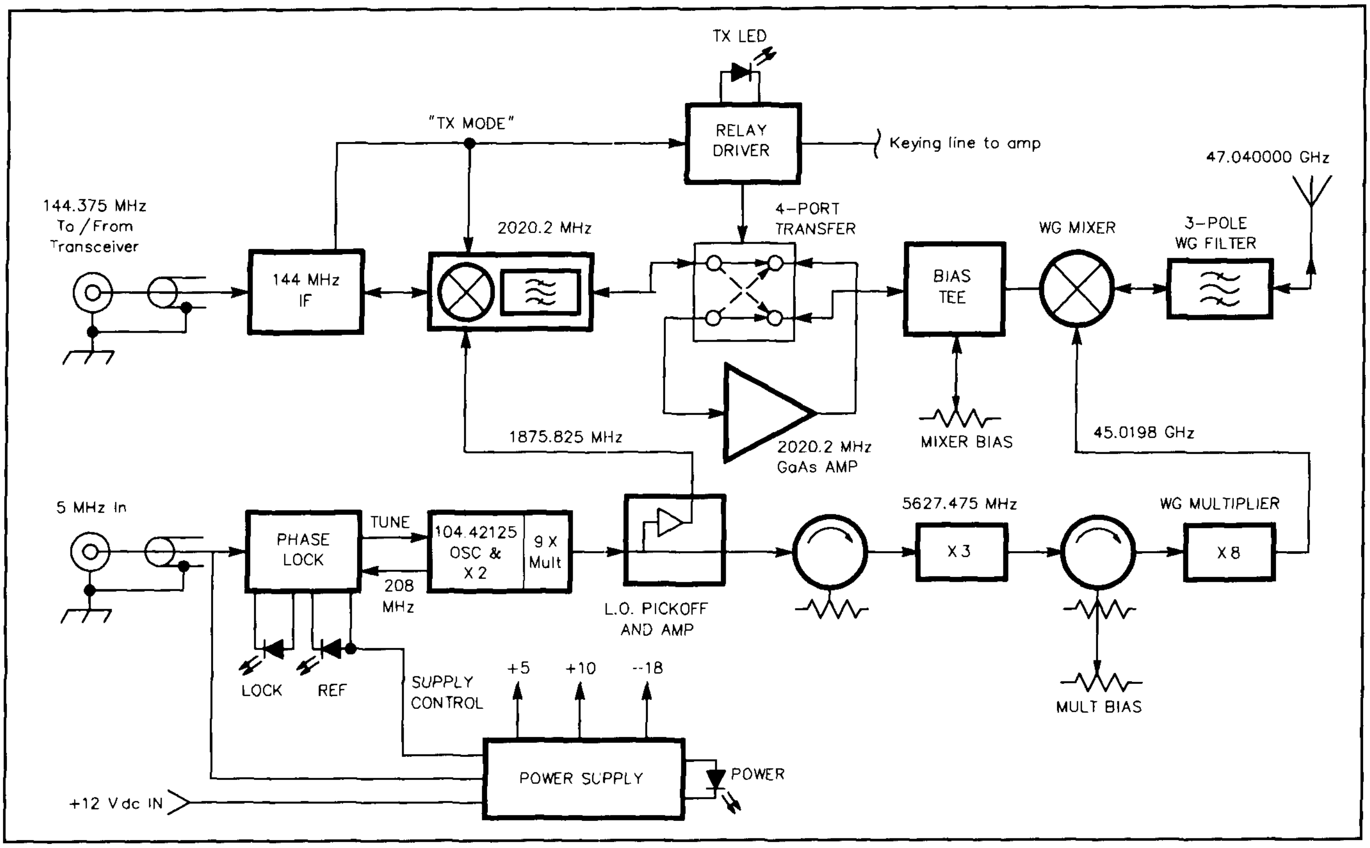


Fig 3—Block diagram of 47-GHz transverter with 2-meter IF.

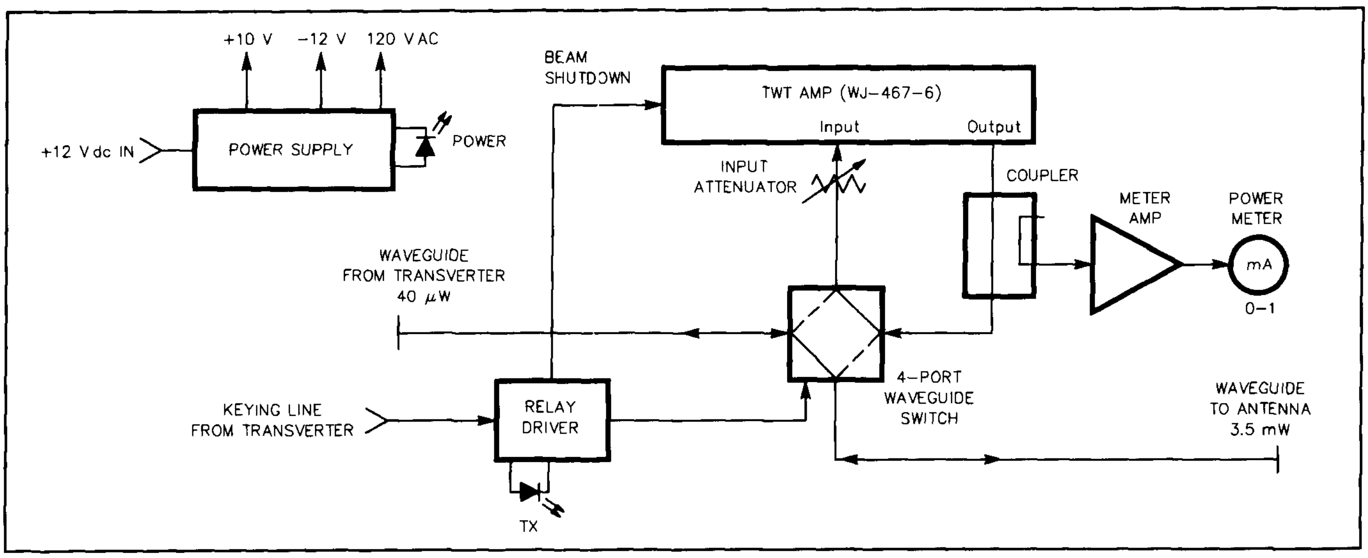


Fig 4—Block diagram of 47-GHz TWT Amplifier.

sion is about 70 dB below the carrier, the LO is only 20 dB down! I therefore put a second filter in the input to the TWT so that when I go to high power (4 mW) the LO is about 35 or 40 dB down. I didn't want the extra 2 dB loss of the second filter in the receive path if I didn't need it. The new 24-GHz rig uses a 6-GHz high IF which nicely gets around these problems. The 47-GHz rigs use 2-GHz high IFs which, while not as good as 6 GHz, they made use of the oscillator/multiplier assemblies that I had available at the time.

The LO scheme used in all these transverters has another feature. I always make the frequencies come out so that all conversions are done with multiples of the same crystal. This means that if I can control the frequency and residual FM of this one oscillator, there will be no uncertainty added by any other oscillators. The importance of this will be further explained later in the section on Frequency Stability. This also means that I have fewer modules to build if I can just tap off the various multipliers as I go.

CLEAN LOCAL OSCILLATOR

Residual FM

If the LO is varying quickly (or FM-ing) it will add this FM to received signal before it passes it to your IF rig. Free-running oscillators, such as Gunn oscillators, tend to have poor residual FM. Note that even 100-kHz-wide FM communications are difficult with Gunn-diode LOs, and narrowband operation is impossible. The stability of a crystal is necessary.

You can indeed phase-lock such a Gunn to a crystal, but the loop calculations need to be done carefully to ensure that the loop response is wide enough to get all that noise and FM clear out to several hundred kHz. Either a locked oscillator or a multiplied crystal can be made to work, but I have found the multiplied approach to be most convenient with the parts I have available.

When you are multiplying the signal from an oscillator by a factor of 470 or more, then any FM-ing or drift of that oscillator will be multiplied by the same factor. This means that if a copiable CW or SSB signal is to result, the crystal oscillator must be very good to start with.

Phase Noise

Any oscillator has some slight amount of fast "jitter," which is called phase noise. This usually manifests itself as a rise in the broadband noise as you measure closer and closer to the carrier itself. This causes two basic problems.

One is that if it is strong enough and wide enough to actually place noise into the mixer at the intended receive frequency, the apparent noise figure of the receiver will be degraded. This is not usually a problem with reasonably high IF frequencies, but it used to sometimes show up in klystron rigs with 10-MHz or lower IFs.

The other problem is that even the relatively low phase noise of a crystal oscillator will begin to show up when it is multiplied as many times as you must to get to 24 GHz or higher. Even if the multiplier adds no noise of its own, the noise will increase in amplitude by a factor of $20 \log(N)$, where N is the multiplication factor, and the answer given is in decibels. This computes to an increase of 53.4 dB in the phase-noise level of 100-MHz oscillator that is multiplied up to 47 GHz. Since the phase noise is much stronger as you get closer to the carrier, the noise that is in the area of 300 Hz to 3000 Hz (the SSB voice area) must be particularly low to start with, or the incoming signal will be drowned out by the receiver's own audio noise!

One of the easiest ways to get better phase noise is to run the crystal oscillator with a high level of power in the crystal itself, in order to get the crystal to help filter out the noise of the active device. This is exactly the worst thing that you can do to a crystal oscillator if you want good frequency stability (particularly long term).

Frequency Stability

The frequency of the transverter must stay put or you will have to chase the signals all over the band. For CW to work in a narrow bandwidth, the signals must be inside this bandwidth. The intelligibility of SSB signals is severely degraded if the frequency is wandering around. If you want to reduce the variables that are all working to prevent you from making your mmWave contact, a good place to start is with a rock-solid signal.

A point of comparison is that a typical "high accuracy" crystal oscillator in a two-meter FM rig has about 5 parts per million accuracy, for about 700 Hz possible error. This transcribes into about 235 kHz of accuracy at 47 GHz. Two rigs using this accuracy of oscillators trying to make contact could be almost *half a megahertz apart!* On the other hand, a high-accuracy, ovenized oscillator can be sometimes found surplus with one part in a billion (1×10^{-9}) frequency drift per day. This gives my transverter an accuracy of 47 Hz at 47 GHz. The 2- meter SSB radio I use as the IF is probably not that good.

This accuracy is only achievable if all of the oscillators in the transverter are locked to the same reference. This is why the idea of using a LO chain that has taps for the different conversions rather than separate oscillators is so important. This saves building separate phase-locks for each one. The one does it all.

REFERENCE LOCK

The answer that I use is not new here (it is in several commercially available pieces of test equipment).[1] Run the main crystal oscillator for best phase noise, and lock it to an accurate external reference. I use a frequency near 100 MHz as it is fairly easy to get acceptable phase noise here. The exact frequency used for each rig will be determined by the aforementioned process of getting all the IFs to come out right. The crystals are reasonably affordable, but are not accurate enough for my purposes.

The accuracy comes from phase-locking the main crystal oscillator to the ovenized reference oscillator also previously mentioned. See Fig 5 for the schematic of the phase lock for the 47-GHz rig. The lock is accomplished with a purposely narrow loop bandwidth. Usually a phase-lock loop has a wide enough bandwidth to reduce the phase noise of the locked oscillator to that of the reference oscillator. However in this case the locked oscillator is considerably cleaner than the reference oscillator. The reference has fantastic accuracy, but not very good phase noise. The loop is cut for about a 1-Hz bandwidth, which allows the main oscillator to determine the phase noise further out than 1 Hz from the carrier.

This scheme also allows me to have only one expensive and hard-to-get reference and put it with whichever rig I am using at the time. This reference oscillator is in its own box, with batteries, and is left on 24 hours a day all year long. This ensures maximum frequency accuracy, as the drift of these references decreases the longer you leave them on.

CIRCUIT DESCRIPTIONS

Refer to Fig 3, the 47-GHz block diagram, for the overall signal flow as each of the individual modules is described.

144-MHz IF Module

The 144-MHz IF module is shown in Fig 6. The SSB rig I used is a Mizuho hand-held, which puts out some voltage through the antenna connector when in the transmit mode. This is sensed here to key the parts of the transverter which are

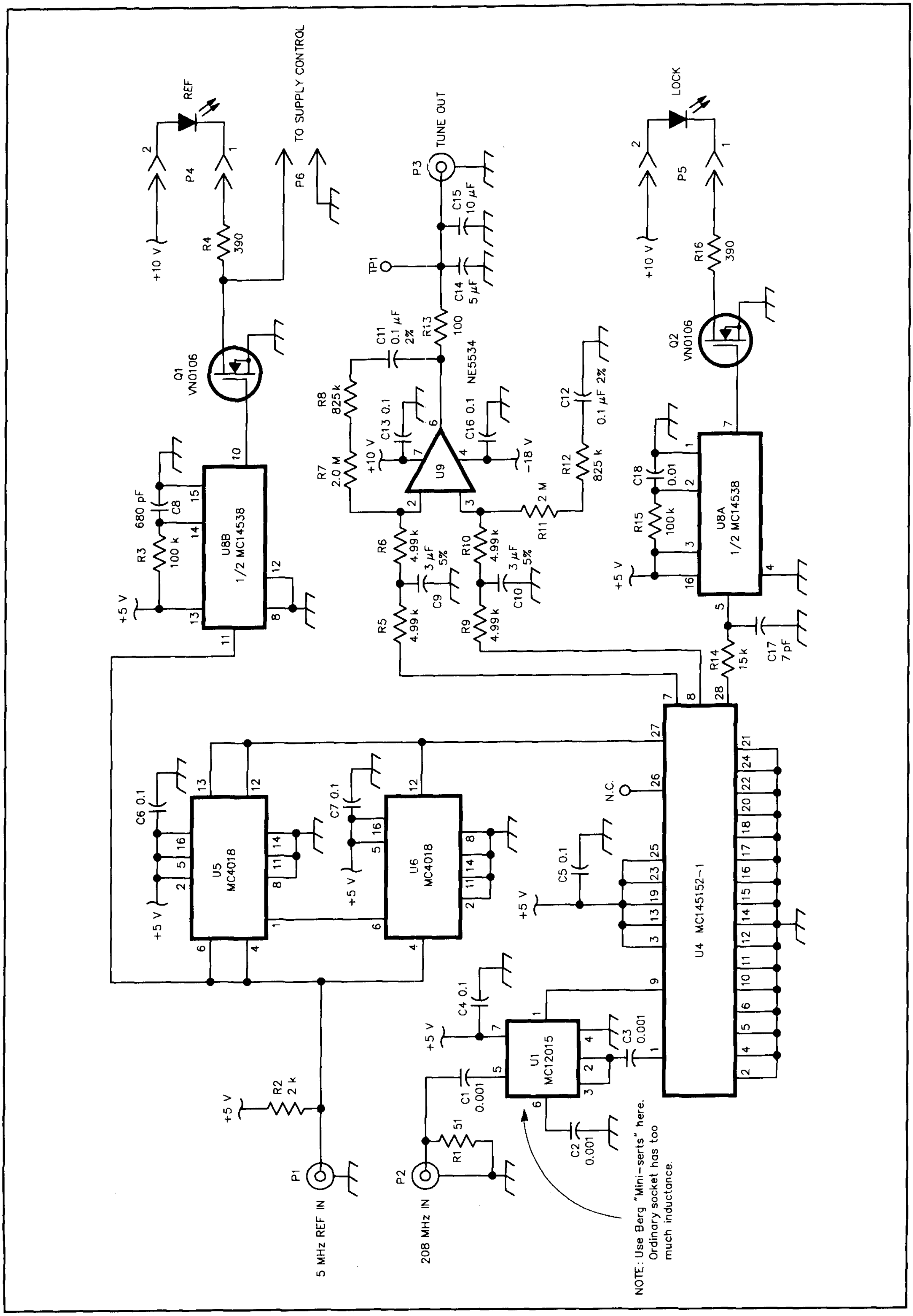


Fig 5—Schematic of the 47-GHz Phase Lock.

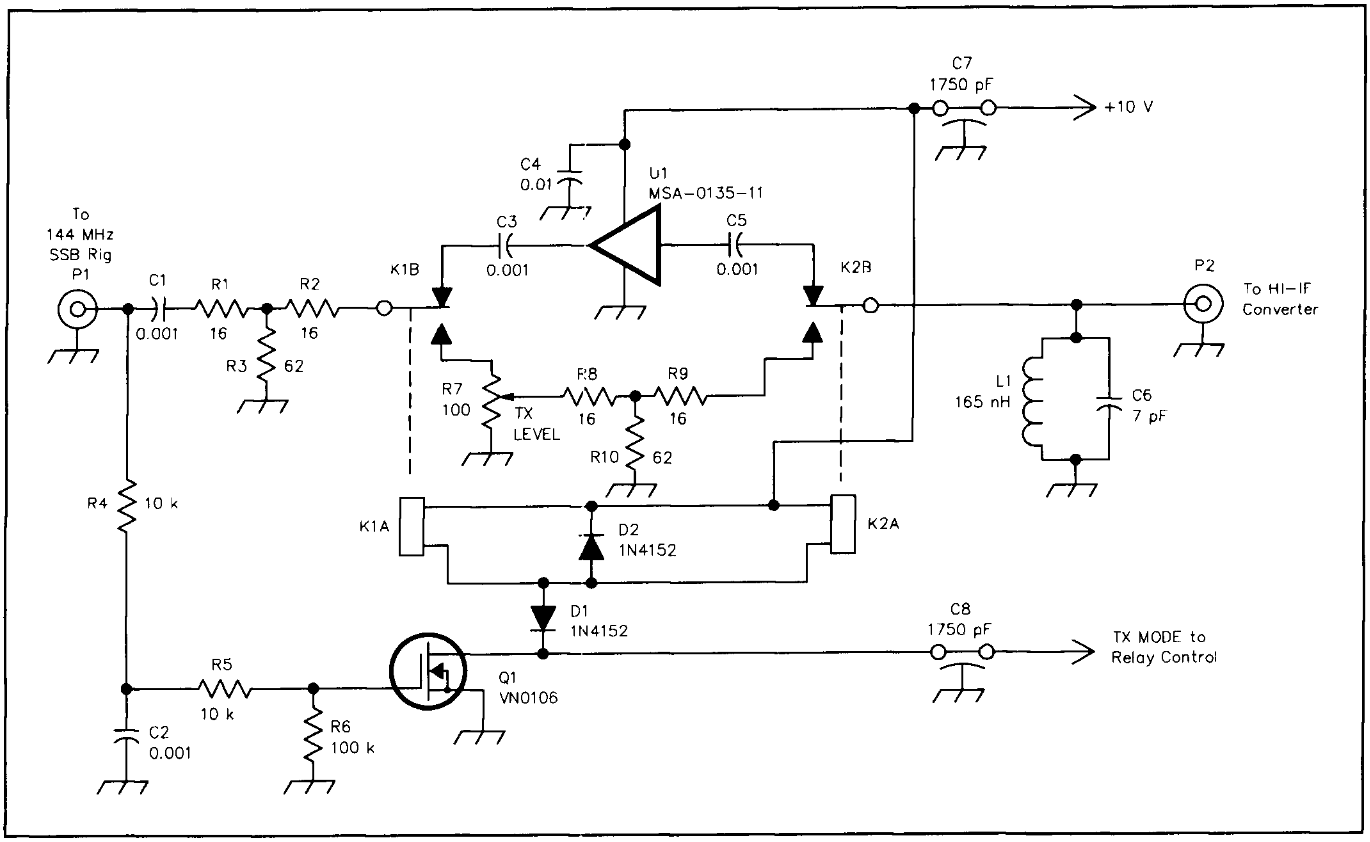


Fig 6—Schematic of the 144-MHz IF Module.

not inherently bi-directional. This module attenuates the 200 mW 2-meter transmit signal to the level which the rest of the rig can tolerate. The gains are set up so that the millimeter wave mixer saturates first, so the "TX LEVEL" pot can adjust the power out for the maximum possible without too much distortion. On receive, the signal from the high IF is amplified by U1 to overcome some of the 2-GHz converter loss and the pad in the 2-meter line.

2-GHz Converter

Fig 7 shows the 2-GHz converter. This is a surplus assembly originally intended for 2072 MHz, which I retuned to 2020 MHz. It is balanced only for the LO, but that is not a problem with the frequency spacings involved here. The 4-pole TEM (coaxial) mode filter is more than adequate to get rid of the image of the 144-MHz IF.

2-GHz GaAs Amplifier

The 2-GHz amp, Fig 8, is a single GaAsFET, which is tuned for lowest noise figure by adjusting the positions of the taps on the input inductor. The noise figure is indeed important here, since this is the first active stage on receive, and therefore the noise figure here will directly add to the system noise figure. This unit achieved 1.7-dB noise figure with 11.5-dB gain. Another dB could be achieved here with a bit more work on the input section, but this was simple and effective. The output is a simple broadband transformer that gives excellent stability.

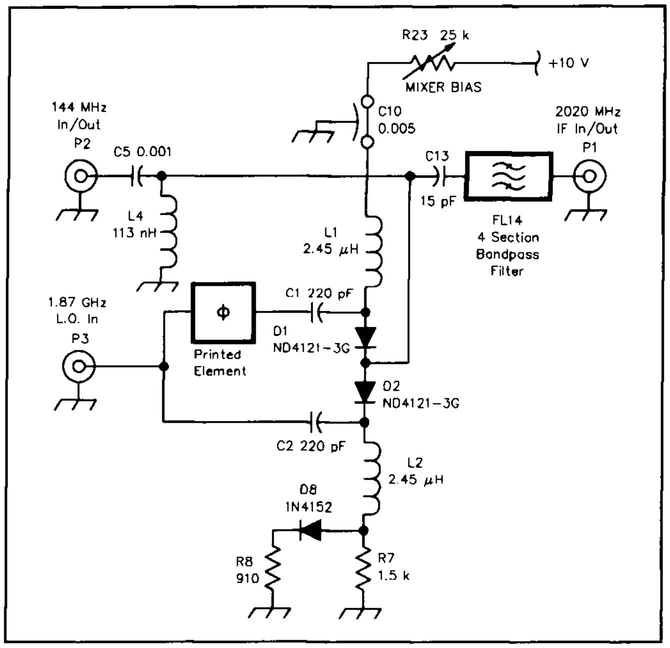


Fig 7—Schematic of the 2-GHz Converter Module.

Waveguide Mixer

The waveguide mixer is a converted surplus unit. There is a basic sketch of it in Fig 9. It originally had the back end filled with a terminating resistor for flatness, and the LO came in the coax. This was used for harmonic mixing with the LO

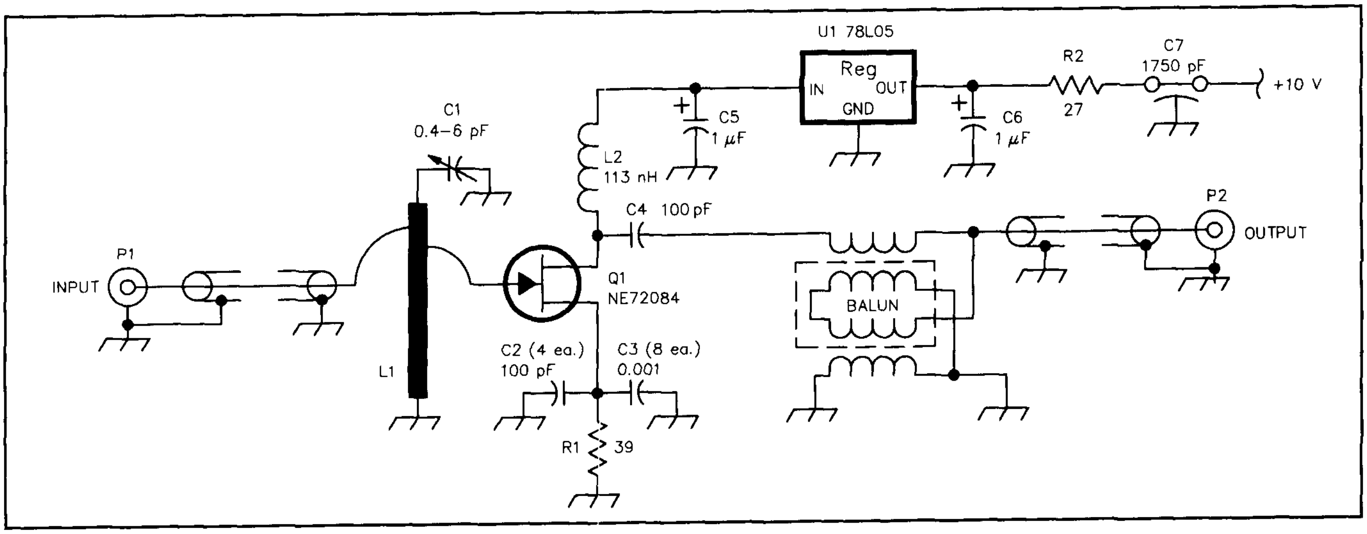


Fig 8—Schematic of the 2-GHz IF GaAsFET Amplifier.

at 2 to 6 GHz. I am using fundamental mixing for much better conversion loss, so I opened up the back and inject the LO here. I installed three tuning screws on each side of the mixer diode, and with a *whole lot* of tuning I got not only a reasonable match, but also there is some reflection of the LO back towards the diode from the RF side and vice-versa, which improves conversion loss over the plain diode. Fig 10 shows the bias tee that gets the dc bias into the mixer which is required by the single diode to optimize conversion loss.

LO Chain

The LO chain is mostly a surplus digital telephone transmitter. The original oscillator and multiplier is simply re-tuned to the new operating frequency. The crystal oscillator had to be very carefully tuned for best residual FM. There is a double and a power amp that feeds a snap diode to get the 1875 MHz, at about +20 dBm. The oscillator also needed a varactor added to it for the tuning voltage that allows the phase lock to work. This has a 4-pole R-C filter added to its tune line to filter out any op-amp noise, 60-Hz noise, etc, so as to improve the locked residual FM. After the snap diode output, I added an LO pickoff to tap a small amount of power off and amplify it to drive the 2-GHz converter. This is a simple resistor and Avantek MSA amplifier. The varactor tripler and the two isolators around it are stock from the surplus telephone gear.

The waveguide multiplier is a bit more difficult. It is another of the surplus mixers (same as the waveguide previously discussed), with a snap diode mounted in it, except that after removing the resistive terminator I closed off the end to reflect the power back and help increase the output power. This also has the same tuning screws and is *even more difficult* to tune. Some positions of the tuning screws return some of the other unused harmonics to the diode to re-convert for more output. They perform a similar function to the idler circuits in a varactor multiplier. In fact, this is almost acting as much like a varactor as it is a snap, since the operating frequency is above the operating range of this diode. The diode is a Metellics mesa snap diode, part number MMD-840-C11, with a 35 pS snap time. They have faster ones now, so I might experiment more

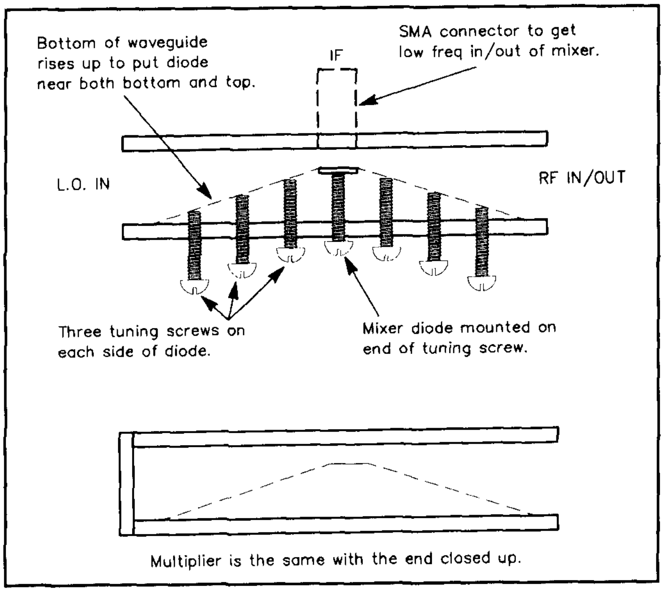


Fig 9—47-GHz waveguide mixer.

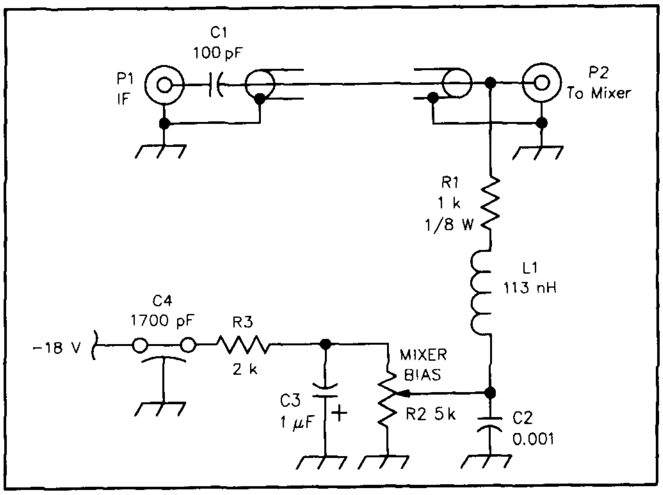


Fig 10—Mixer bias tee.

in the future. The multiplier requires bias too.

Phase Lock

Fig 5 is the phase lock that locks the 100-MHz oscillator to the external reference. This is a Motorola PLL chip, U4, that can control a dual-modulus prescaler chip, U1. The 200 MHz from the oscillator and multiplier is fed to the pre-scaler, and the 5-MHz reference is divided by the two MC4018 dividers, U5 and U6, before going to the PLL chip. These extra multiply and divide operations were required to get the correct lock ratio for the 104.421296296 MHz required frequency, while maintaining a high enough frequency at the phase detector to keep sidebands from being inserted on the carrier. U9 is the loop amplifier, and I had to use the "pole splitting" technique of adding C9 and C10 to reduce the phase-lock sidebands. U8A detects the lock condition so as to let the operator know if the frequency is right, and U8B senses the presence or absence of the reference.

Power Supplies

The whole unit is required to operate from batteries for mountaintopping, so it must operate from 10.5 volts to about 15 volts dc. This is accomplished with two supplies. Fig 11 shows the linear low-overhead regulator that supplies +10 volts to most of the RF circuits. The surplus oscillator/multiplier on the other hand, required −18 volts at about 300 mA (of all things)! This is provided by a small switcher that is shown in Fig 12. This was designed to not need any transformer or inductor, so it is easily reproducible.

It has its own free-running oscillator, but the *very* fast transitions of the switcher FETs do cause some VHF birdies. There is a fair amount of filtering to reduce this, but there is still a very tiny amount left. If the free-running oscillator were to drift to just the right frequency, then one of its harmonics might be barely audible in the 2-meter IF. It would drift around a bit, causing the operator to maybe mistake it for a weak signal. In order to prevent this, U2 in the supply divides the reference frequency down to the switcher frequency, and if the phase-lock module tells the supply that a reference is present, the divider output is used for the switcher drive. The switcher is a simple "rail-to-rail" topology, with a linear post-regulator for overall simplicity, if not the ultimate in efficiency.

The relay control module runs all the transmit/receive switching functions and gets its input from the detector in the 144 IF module.

POWER AMPLIFIER

The power amplifier is a traveling wave tube. For one of the 47-GHz rigs it is contained in its own separate housing, so that I can leave it behind if I don't need it. In this case the definition of "power" is a bit different than you may be used to. The most powerful amplifier I could find was a surplus receiving type of TWT. This amplifier was made by Watkins Johnson, model number WJ-467-6. It is a self-contained amp, with tube and power supply in one compact unit. Fig 4 is the block diagram of the amplifier I built.

TWT

The TWT that I managed to find is a unit that is intended to operate from 26.5 GHz to 40 GHz, with 35 dB of gain and 4-mW minimum output (most of them do close to 10 mW out). 47 GHz is well outside of its operating range, by 54% of its original bandwidth. The tube as received was not useful at this frequency. I opened it up and readjusted the helix voltage

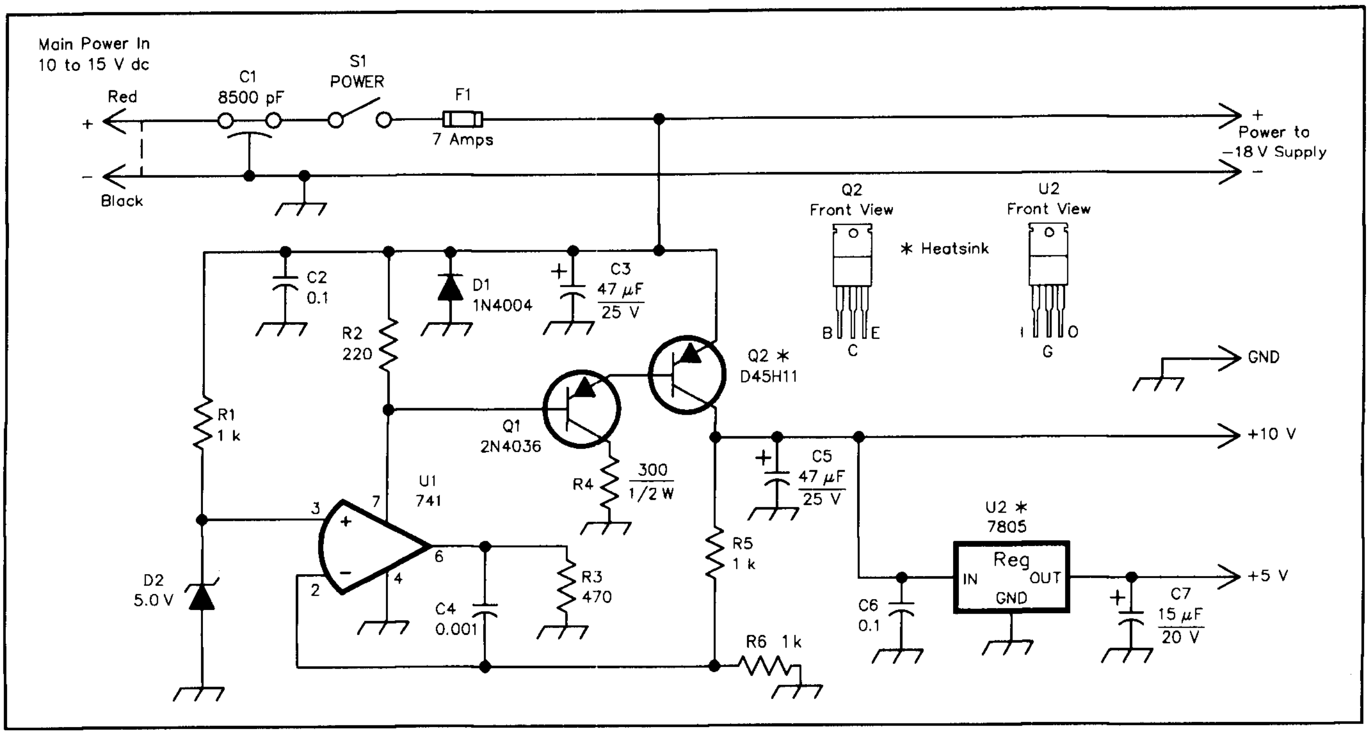


Fig 11—Ten- and 5-volt power supplies.

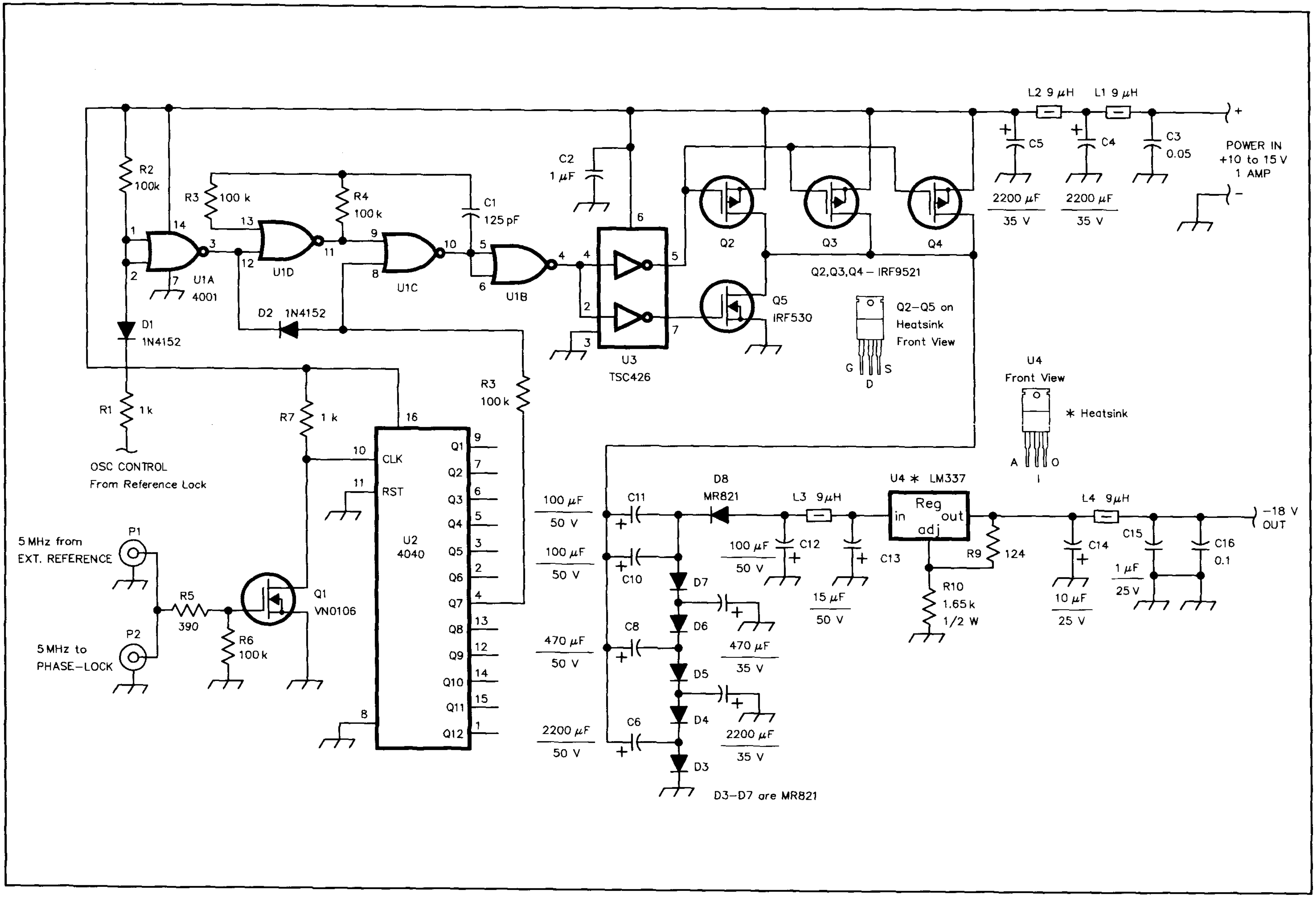


Fig 12—Negative 18-volt switching supply.

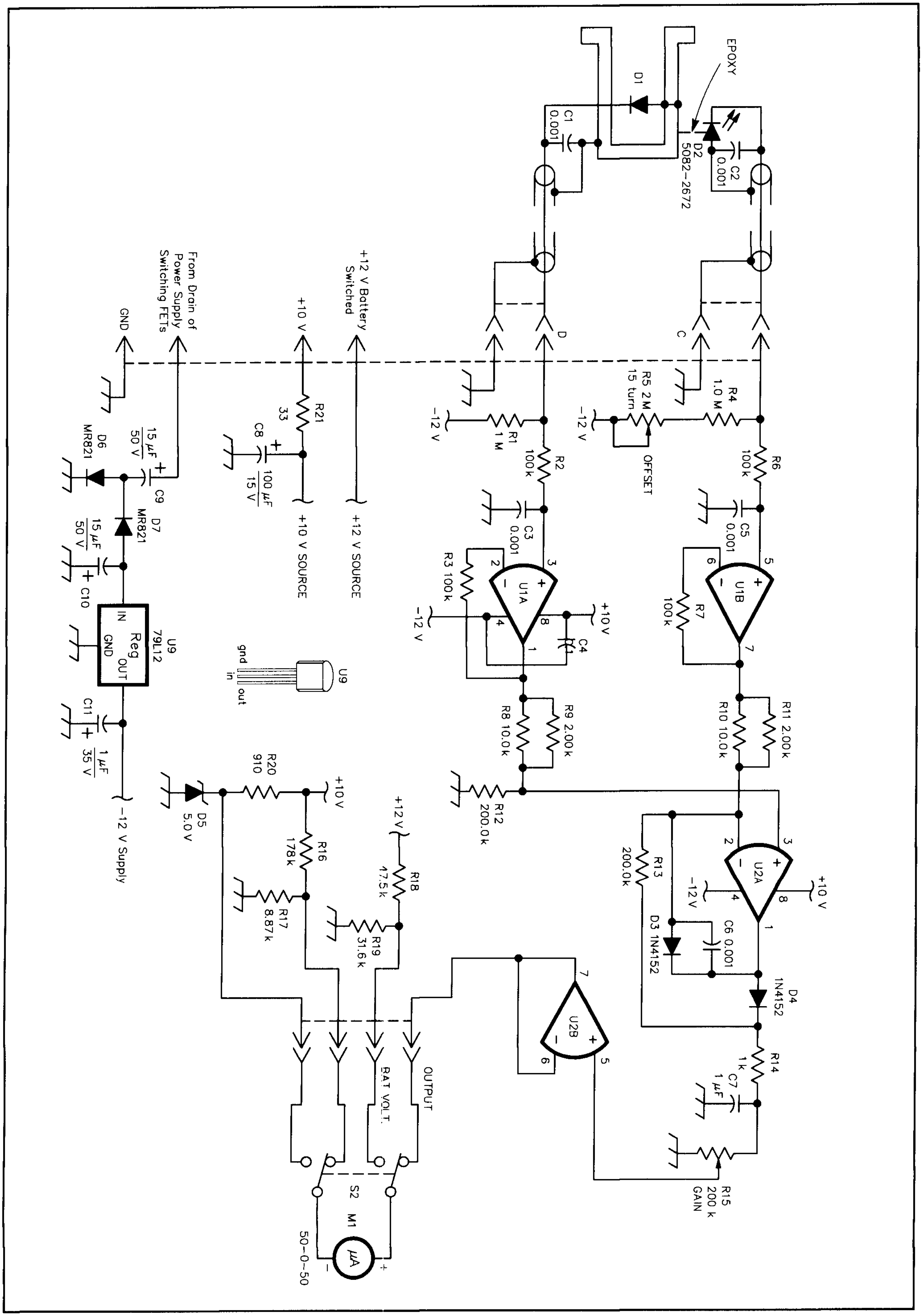


Fig 13—Power detector and meter driver.

lower until the tube attained useful gain at 47 GHz. This, however, lowered its power output to well below one milliwatt. To fix this, I lowered the grid bias so as to raise the collector current until I got the output power back up. The tube now shows 24-dB gain with 4.5-mW power out.

I also added an external connection to the TWT module that controls the grid voltage in order to shut down the beam. This is necessary so that the tube is inactive during receive. The tube's noise figure is slightly worse than the barefoot transverter, so it would only hurt if I were to try to use it during receive. The TWT is therefore bypassed in receive.

The TWT requires 120 VAC for its primary power, so the switching power supply that runs the amplifier was designed to run at 380 Hz. It drives a transformer to get the necessary voltage for the TWT.

Attenuator

There is also a waveguide attenuator that I built by boring a small hole in the wide side of a piece of waveguide, and pushing in a piece of microwave absorber. It is adjusted by pushing the absorber to different depths in the guide. This attenuator is needed to prevent overdriving the TWT. If it is overdriven it will not only distort, but it will even reduce its power out as it gets driven harder.

Power Meter

In order to be able to set the attenuator correctly (a one-time operation), a power meter is helpful. It is also quite useful to keep an eye on the general health of the transmitter. I built a coupler out of two pieces of waveguide soldered next to each other with a hole bored between them. This taps off a small bit of signal to a diode which then feeds a peak-reading meter. The peak detector circuit allows reading of SSB signals. The schematic of the waveguide diode mount and peak detector is Fig 13.

OBTAINING PARTS

The parts for this project are proportionately harder to find than "ordinary" microwave parts, as the frequency is that much higher. I have found many of them (including one of the TWTs) at the Dayton HamVention. There are an assortment of places in California with *real* expensive parts (the waveguide switch was $100, an 18" broken piece of flex guide was $50). Occasionally I find millimeter parts in ordinary local hamfests. It took me over two years to collect enough parts to begin serious construction on the 47-GHz rig.

Another note worth mentioning here is that I never located much waveguide for any band that includes 47 GHz. I wound up using all parts designed for the 26.5-40 GHz band. This means that I must keep all discontinuities in the guide to a minimum, to prevent mode-hopping. This would result in excessive losses.

OPERATION

The 47-GHz rig described here was used with an 18-inch dish at one end of a world record DX contact that stood for 4 years. The other rig had a 29-inch dish with it. Both dishes are equipped with rifle 'scopes to aid with aiming. This contact was between WA3RMX at Crater Lake Oregon, and K7AUO on top of Mt. Ashland, Oregon, for a distance of 65.37 miles.[2]

CONCLUSION

I hope these ideas will spur on more hams to venture into the Millimeter Wave arena. Modify whatever you can get a hold of, and get on the air! With the current push to reassign our bands away, how long before someone finds these useful?

Above 10 GHz the calling frequencies are (in GHz):

- 24.192
- 47.040
- 80.640[3]
- 120.000
- 144.000
- 245.760

I expect to hear more people on these bands soon!

Notes

[1]This scheme is used in Tektronix spectrum analyzer models 494AP and 2756P, as well as in other places.

[2]For a complete description of this contact see *QST*, Dec 88, p 87 "The New Frontier" column by Bob Atkins, KA1GT.

[3]Although this frequency does not strictly follow the rule of the lowest reasonable multiple of 96 MHz, I am hoping we will use this multiple instead, since in this particular band the atmospheric losses decrease with increasing frequency.

A Solid-State Laser Transceiver

By Roger Wagner, K6LMN

(From *Proceedings of the 37th Annual West Coast VHF/UHF Conference*)

INTRODUCTION

Welcome to the world of Amateur Radio lightwave communications! Here, we are communicating voice/CW/data/video, line-of-sight over unguided coherent light beams at frequencies of 350 to 680 THz (terahertz or million megahertz) or about 3 million times the frequency of the 2-meter band. This translates in wavelength to 700 nM (deep visible red) to 400 nM (visible violet). Of course, infrared and ultraviolet can be used as well. The FCC has given hams and non-hams any of the frequencies above 300 GHz, where extremely deep infrared begins.

In this article, I have developed the concept of an all solid-state CW/phone transceiver, such as is used on the UHF bands. The units are slightly larger than a hand-held transceiver, complete with batteries. The power output is only 5 mW, and the receiving "antenna" is a 3-inch telescope. As on the lower bands, range is a function of power output and antenna size. Communicating over laser beams sure is a lot of fun and the band is certainly not crowded! A simple laser diode/LED AM-voice modulator is just a couple of amplifier chips and a transistor or two. A simple receiver is a small solar cell followed by an amplifier chip. Most parts are available from Radio Shack.

Unfortunately, there has been a shortage of good and practical articles in the ham magazines on the use of semiconductor lasers and associated solid-state photodetectors. Most of the authors in this field got their start with gas lasers. I hope this article will shed some light on laser diodes, and encourage your investigation in this new world of communication.

BACKGROUND

After some 35 years of HF through UHF operation I wanted to try something new—lightwave communications. I started out using automobile quartz-iodine headlights. Believe me, you can throw a light beam a long way with one of these. My goal was voice/CW communications and I quickly found out the heated filament didn't modulate well at frequencies above a few hundred hertz. Next, I began experimenting with Radio Shack's high-powered red LED. It offers an output of some 2000 millicandelas for a cost of less than $2. I built up a couple of voice-modulated AM transceivers, using solar panels as photodetectors. Range was several blocks and fidelity was excellent. Then, I experimented with various lenses to collimate the LED's wide-angle beam. Range improved considerably since the beamwidth was now down to 2°. I do a lot of VHF/UHF contesting and I was now ready to generate some extra points for ARRL VHF/UHF contests. To my surprise though, only coherent light sources, ie, lasers, are permitted.

I purchased several laser diodes, complete with collimators, and added the laser assemblies to my transceivers. Now I'm legal for VHF/UHF contesting. My two transceivers were used for the 1990 ARRL June VHF contest from Mt. Pinos by the N6CA group. I also used them during the 1992 ARRL January VHF Sweepstakes. My range so far is only 5 miles. I think longer distances are possible, but I have great difficulty pointing the extremely sharp beam (0.02° beamwidth).

SAFETY

Lasers are not toys! Never point a laser at eye level at close range. The beam emitted by a laser diode is harmful if aimed into the unprotected human eye, especially at night, when the iris is wide open. Never point the laser into known aircraft flight paths.

LASER BASICS

Laser is an acronym for Light Amplification by Stimulated Emission of Radiation. A laser generally consists of an active medium that can amplify light, and reflectors that return a portion of the light back to the medium. A small portion of the amplified light that bounces back and forth escapes into a beam with these characteristics:

- Highly directional
- Single or narrow spectrum of wavelengths
- Phase coherent
- Continuous for relatively long periods

The medium can be any of the following:

- Solid-state (ruby, YAG)
- Liquid (dye laser)
- Gas (Helium-Neon, Argon)
- Semiconductor (laser diode)

The wavelength of the emitted light is a function of the medium and mirror spacing. All of these lasers have their applications, but only two are presently used for free-space communications by hams.

HELIUM-NEON LASERS

What about using commonly available surplus HeNe and HeCd laser tubes? Since the HeNe (gas) lasers have been around for some time, they are cheap and plentiful on the surplus market. They cost between $40 and $100, including high-voltage power supply. Gas lasers do the job well if you are interested in just pointing a narrow beam of light at something. To me, they aren't very useful for communication, as they are difficult to fully modulate. In addition, they require high-voltage power supplies (over 1000 Vdc). You can achieve 10-15% AM-voice modulation by modulating the power supply. For CW, a chopper wheel (such as a compact fan), in the beam path will produce square-wave modulation at about 1000 Hz. Now, just key the power supply on and off to transmit 100% modulated MCW.

ARRL contest rules state that lightwave-frequency transmitters must generate coherent light, and be capable of communicating at least 1 km. The receiver must have at least one stage of amplification. As LEDs and headlights don't generate coherent light, they can't be used for contest QSOs. Serious ham DX is still being worked with gas lasers and mechanical chopper wheels.

SEMICONDUCTOR LASERS

Semiconductor lasers (laser diodes) are now coming to the surplus market at reasonable prices. They are used in lecture pointers, bar-code readers, CD players and laser printers. Why use a laser diode for communication? The primary advantage of laser diodes over HeNe lasers are:

- Greater efficiency
- Physically rugged
- Compact size
- Light weight
- Long life (over 50,000 hours)
- Easy to modulate up to 1 GHz
- Simple power supply (3-12 Vdc)

The typical laser diode looks like a TO-5 case transistor

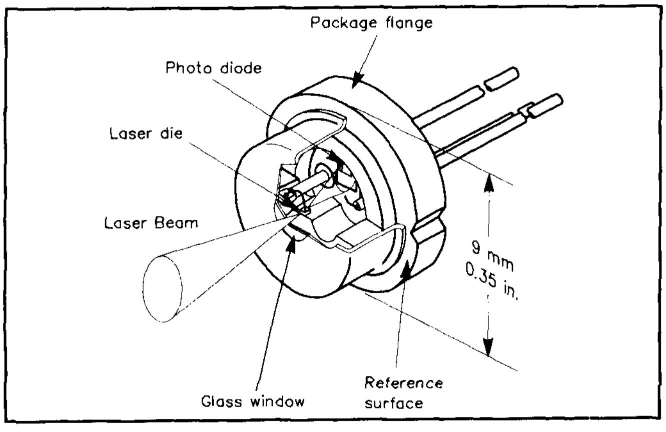


Fig 1—A typical laser diode.

with a glass window (Fig 1). Unfortunately, low-cost laser diodes are at near-visible infrared wavelengths of around 750-820 nM. Visible red (670 nM) laser diodes are now available, but at much higher cost. A small collimator lens is required to narrow the beam from about 10 × 30 degrees to 1-2 milliradians (mrad). With an adjustable collimator the beam can be focused to any width required. A HeNe laser has a typical divergence of 1 mrad, without external optics.

For the same optical power output, the HeNe laser appears almost 10 times brighter to the eye than the typical laser diode. The HeNe wavelength of 633 nM is closer to the eye's most sensitive wavelength (green, 540 nM). For communications, the red and near infrared (670-850 nM) semiconductor lasers are a better match to commonly available silicon photodetectors, which have a sensitivity peak at near-infrared wavelengths. Laser-diode efficiency is typically 20%, compared to 2% for HeNe lasers. Above the lasing threshold, laser diode efficiency approaches 80%. If you are contemplating portable, battery operation, this feature is significant. Table 1 is a comparison of several available laser diodes.

OPTICAL RANGE

The maximum line-of-sight (LOS) range of a free-space, all solid-state laser communications system can be calculated by:

$$R = \sqrt{\frac{(Po \times Ar \times Tor \times Tot \times Ta)}{Pt \times D}} \quad \text{(Eq 1)}$$

where

R = LOS range in meters between transmitter and receiver

Po = Peak power output from laser (watts)

Ar = Area of the receiver lens or mirror in square meters

Tor = Transmissivity of receiver optics, including filters

Tot = Transmissivity of transmitter collimator and beam-shaping optics

Ta = Atmospheric transmittance

Pt = Threshold power sensitivity (0-dB signal + noise : noise ratio) (watts)

D = Collimated laser-beam divergence (radians)

This equation assumes that the apertures of the transmitter optics have been matched to the beamwidth of the laser diode for minimum loss. Otherwise, another term must be included in the numerator.[1]

The receiver's minimum detectable power term (Pt) needs some explanation. This discussion pertains only to solid-state photodetectors. The minimum incident power on a photodiode required to generate a photocurrent equal to the total photodiode noise current is defined as the noise equivalent power, or NEP. The NEP is dependent on the square root of the receiver bandwidth, which is usually limited by the audio passband. The NEP is given in watts/ $\sqrt{Hz}$. The noise generated by the diode is a combination of shot noise and Johnson noise. The shot-noise current is a function of the dark leakage current; Johnson noise is related to the diode's internal resistances. Both are device and temperature dependent. Photodiode NEP can be expressed by:

$$NEP = \frac{In}{Rs} \quad \text{(Eq 2)}$$

where

NEP = noise equivalent power (watts/ $\sqrt{Hz}$)
In = photodiode noise current (A/$\sqrt{Hz}$)
Rs = photodiode sensitivity (A/W)

Typical silicon PIN photodiode figures are In = 2.6^{-14} A/ $\sqrt{Hz}$, and Rs = 0.5 A/W at 670 nM. NEP then = 5.2^{-14} W/ $\sqrt{Hz}$ A low-noise op amp (OPA27) following the photodiode has a much greater noise current of 4^{-13}A/ $\sqrt{Hz}$. The geometric sum of the diode and op amp noise currents is then 4.01^{-13}A/ $\sqrt{Hz}$, yielding NEP = 8.02^{-13} W/ $\sqrt{Hz}$.

Let's predict the range for my laser transceivers:

Po = 0.003 watt max.
Ap = 0.0046 meter^2 (3-inch lens diam)
Tor = 0.95 (estimate for plastic Fresnel lens, no filters)
Tot = 0.9 (est)
Ta = 0.9 (extremely clear night at mountaintop altitudes)
Pt (2.7-kHz receiver bandwidth) = 4.18^{-11} W
D = 0.001-radian circular beam

Range = 500 km, or 311 miles for a signal barely detectable in the noise. This is the theoretical maximum; no consideration is given for modulation type or atmospheric effects, such as air turbulence or transmitter-to-receiver beam misalignment. The theoretical VHF radio LOS range between a transmitter and receiver, both at 9000 ft (unobstructed path) is about 268 miles. The optical range should be a bit less. The beam diameter at long range is approximately:

Beam diam = Full-width Beam Divergence (radians) × Range

Thus, beam diameter at 1 mile = 0.001 radian × 5280 ft = 5 ft.

TRANSMITTER BEAM SHAPING AND OPTICS

Unlike gas lasers, laser diodes alone emit a divergent asymmetric beam, due to diffraction effects in the diode's asymmetrical laser cavity. In addition, unlike the long cavity of the gas laser, the semiconductor laser cavity is extremely short (sub-millimeter) and the short cavity does not allow for a collimated beam. A typical near-visible laser diode without external optics emits an asymmetric (elliptic), divergent cone of light of approximately 11 × 37 degrees, 3-dB beamwidth. Referring to Fig 1, the ratio of the orthogonal beamwidths is called the aspect ratio. For these diodes it is approximately 3.5:1. The aspect ratio for the typical visible laser diode is nearly 5:1. Laser diodes also have a small amount of astigmatism. External optics are thus required to transform the output of a laser diode, to give a corrected, circularized, collimated light beam. The ideal collimated beam has no divergence. In other words, it is a parallel beam of light out to infinity distance.

When a simple lens is used to collimate this asymmetric output, a narrow beam with an elliptic cross-section results. A slightly elliptical beam is acceptable for amateur communications. It is, however, desirable to correct this distortion. An anamorphic prism pair or cylindrical-lens telescope can be used. Prisms are easier to align and can be adjusted to correct different amounts of ellipticity. An elliptically shaped fixed iris can also be used to circularize the beam, but at a considerable loss in beam power level.

There are two common ways of achieving beam collimation for laser diodes in free-space operation. A simple plano-convex singlet lens is a very economical solution to laser-diode focusing and collimation. A well-designed doublet lens is usually superior to a corresponding singlet, because of the much higher degree of aberration correction which is possible. There are also available aspheric lenses, which have shorter focal lengths than equivalent plano or bi-convex lenses. The corrected aspheric lens has a much higher numerical aperture (to capture maximum emitted light) than the corresponding spherical singlet, while having less aberration.

Beam astigmatism is caused by the light sources for the orthogonal axes being at slightly different distances. In the simpler, gain-guided devices like visible laser diodes, this distance can be 40-50 microns. In the case of index-guided lasers (typical infrared), the astigmatism is typically less than 5 microns. Astigmatism often needs correction if a small spot size or ultra-precise collimation are required. This is particularly true in the case of visible lasers. Cylinder-ground lenses are available to correct astigmatism. By slight rotation around the optical axis, the lens can correct astigmatism without introducing additional aberration. The astigmatic correction lens, if required, should be placed after the collimator lens. Sony laser diodes use an integral slanted window over the diode chip to achieve very low astigmatism, without external corrective lenses.

For these corrective lenses, we are not talking about large-diameter optics like those used in telescopes. Typical collimating lenses have diameters of about ¼ inch. With good optics, a circular beam of less than 0.2 mrad divergence can be achieved. A simple plano-convex lens can easily converge a near-visible infrared laser diode to a 1 × 3 mrad beam. Antireflection coatings should be used to prevent light scattering, and to reduce path losses. A simple coated plano-convex collimator lens has a transmissivity over 95%.

LASER TRANSMITTERS

Now we have the transmitted beam shaped, what about modulating it? There are several types of modulation suitable for communication with laser diodes. As on the lower bands, the type of modulation used depends on whether you wish to transmit voice, data, Morse code or video. Fig 2A shows a diagram of a simple laser transmitter and its analogy to an HF AM-voice transmitter. At lightwave frequencies, the FCC does not dictate the mode of modulation or maximum bandwidth. Here, simplicity and low cost may dictate the modulation method. A continuous (CW) or unmodulated light beam carries no information and the receiver detector can only detect the presence or absence of the light beam.

Three basic methods of modulating the laser diode are:

- Modulated CW (MCW), using an external light chopper
- Amplitude modulation for voice/CW/data/video
- Pulse modulation

MCW

MCW is used by the HeNe laser gang. It is simple and cheap, but is suitable only for Morse code. To generate a

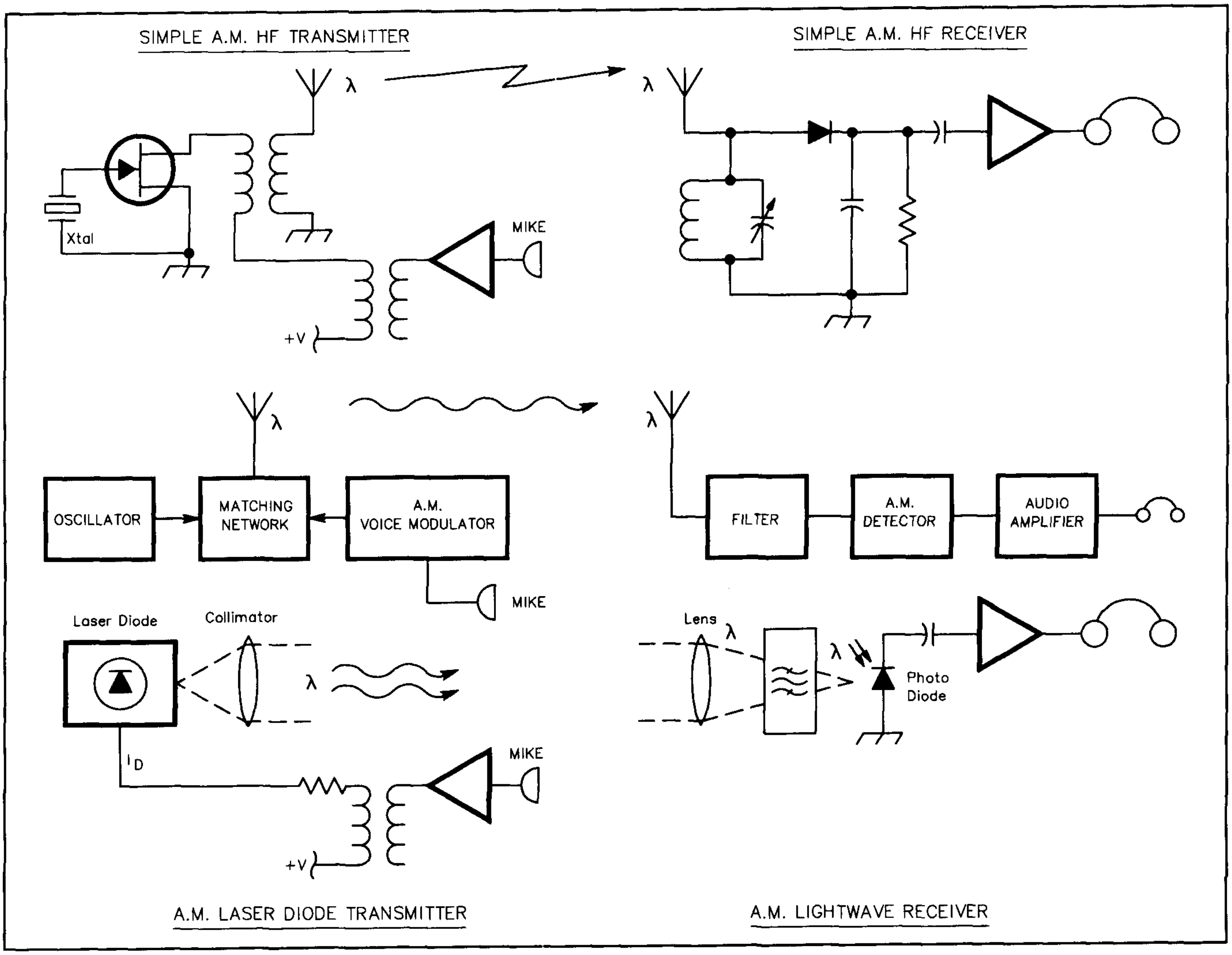


Fig 2—Comparison of HF and lightwave transmitters and receivers.

modulated light beam, all that is needed is a motor-driven wheel (fan) spinning in front of the light beam, to interrupt it at an audio rate. A solenoid-operated shutter, liquid-crystal shutter, or even your hand, can be used to key the light beam. Liquid-crystal shutters are sluggish, especially at the low temperatures often found at high elevations. They have a transmissivity of only 70% when clear. Of course, you can also key the laser power supply, or modulate a laser diode supply.

Amplitude Modulation

Amplitude or intensity modulation is rather simple to implement with laser diodes, and is suitable for voice, data, Morse code and video. Another plus is that AM is compatible with the receivers used for gas lasers. This feature is handy during contests, as it allows contacts with stations using either laser type. For AM, the laser diode output is established at the 50% output level with no modulation present. The modulating waveform then modulates the laser from the bias point up to 100% and down to 0% output. For Morse code, the sinewave tone oscillator can be keyed. The average light power output in all cases is 50% of the peak laser output. A 30-100 kHz subcarrier can be amplitude modulated, and itself used to modulate the laser. This method complicates the receiver slightly, but offers a major advantage: this receiver will be nearly immune to stray light modulated by ac power like street lights. For television transmission, couple the camera baseband output through a wideband stepdown transformer to the laser diode. For best picture quality, drive level and bias may have to be more carefully adjusted than for low-fidelity voice transmission.

Another modulation scheme involves coupling a few milliwatts of HF SSB into the laser diode. For a simple receiver, just hook up a wideband photodetector, such as a PIN diode (biased for maximum bandwidth) to the HF receiver antenna input. A low-noise, high-frequency preamp following the photodetector will improve receiver performance. Speaking of exotic modulation techniques, for coherent CW, the laser diode transmitter tone oscillator can be phase-locked to a master frequency such as WWV. At the receiving end, the detector can be referenced to the same master frequency. With an extremely narrow receiver baseband bandwidth of 2 to 10 Hz, a tremendous system gain over voice modulation can be achieved. The sloppy tone generated by mechanical chopper wheels typically used with gas lasers will not permit the narrowband DX work possible with laser diodes in such a system.

Pulse Modulation

Pulse modulation is quite suitable for Morse code operation, but can be used for voice and data modulation as well. All that is necessary is to pulse the laser on to full power output and then pulse it off, with a squarewave. The light output is a pulse or squarewave. Using pulse modulation, you don't have to worry about laser diode thresholds and linearity of light output. For Morse code, drive the laser diode with a squarewave current of 0-75 mA peak at audio frequencies and key the oscillator to form characters. Use a simple transistor switch driven by a square-wave oscillator like the 555 IC. The simple receiver can be any HeNe-system photodetector and receiver.

For voice and data modulation, either pulse-width modulation (PWM) or pulse-position modulation (PPM) can be used at a subcarrier frequency. For PWM, the subcarrier frequency is normally 30-100 kHz. The audio modulation input varies the pulse width or duty cycle. At the receiver, the recovered pulse train passes through a low-pass filter before the audio is recovered. Because the carrier frequency is so far above the audio range, this system isn't compatible with most HeNe laser receivers.

For pulse-position modulation, the pulse train average frequency is again set at 30-100 kHz. This time though, the pulse phase or frequency is varied by the audio input signal. The receiver detector is a little more complicated: a phase or frequency detector is required. The benefit? A static-free, hi-fi communications system.

A word of caution: All laser-diode manufacturers strongly recommend slow turn-on to avoid transients that may destroy the device. If you plan to use high-speed pulses, verify the drive waveform with a high-frequency oscilloscope. You should also take precautions to avoid spurious oscillations in the driver. A 100-pF capacitor may be connected in parallel with the diode. Keep the leads short. Test to be sure the additional capacitor doesn't cause your driver stage to oscillate. Another good idea is to add a small silicon diode in parallel with but in reverse polarity to the laser diode. A 1N4001 is suitable for frequencies up to 100 kHz. Do not exceed the manufacturer's specified maximum power ratings.

Laser Diode Operation

Before getting into the actual driver circuits, we must understand the laser diode's operation. The shape of the laser diode voltage-versus-current curve is quite similar to that of a silicon diode.

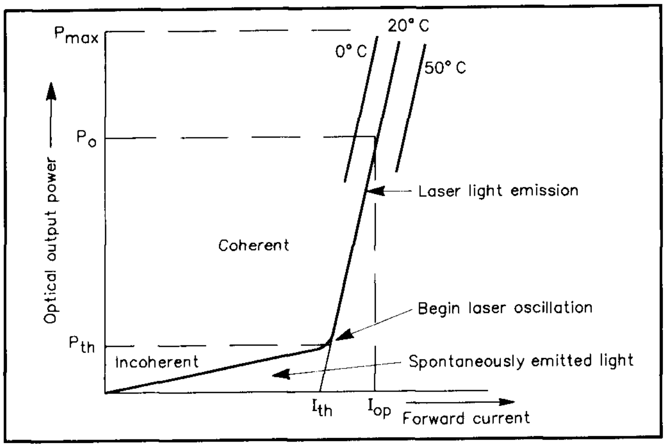


Fig 3—Laser diode forward-current optical output characteristics.

The major difference is the threshold voltage: 0.6 V for silicon diodes, but 1.8-2.2 V for laser diodes. The laser diode should always be mounted in a heatsink. Laser diodes have a Zener diode-like curve. The voltage drop remains fairly constant with increases in current. The laser-diode curve also resembles that of an LED, but its light output is nearly proportional to drive current above the lasing threshold. The lasing threshold of a laser diode is constant for a given temperature; as temperature increases, so does threshold current.

Drive current should be adjustable. The transmitter design should regulate drive current according to temperature, unless the diode is always going to be operated at room temperature. Laser-diode drive is a variable current, not voltage. The bias circuit must accommodate a 1.5-2.5 V swing, but it is diode current that modulates the light output. A calibrated laser light meter is a very helpful tool for adjusting the transmitter.

Laser Drive Circuits

A simple wideband laser-diode bias circuit can be a simple resistor and negative temperature coefficient (NTC) thermistor

Table 1
Comparison of Laser Diodes

Model	*Mfr*	*CW Pwr MW Max*	*Wavelength Peak #*	*Beamwidth*	*Ith*	*Iop mA*	*Lo mA*	*Notes Cost $*
LT022MC	Sharp	5	780	11	33	50	65	Surplus
LT024MD	Sharp	30	780	10	29	55	85	
SL151U	Sony	5	670	11	30	75	85	Low astig
SL151V	Sony	5	670	11	30	75	85	
SLD201U3	Sony	50	780	14	28	80	120	
SLD304V	Sony	1000	810	13	28	450	1400	Cost >$2500!
TOLD9200	Toshiba	3	670	7	34	76	85	
TOLD9201	Toshiba	5	670	10	35	80	90	
TOLD9211	Toshiba	5	670	8	33	50	60	Low astig
TOLD9410	Toshiba	3	650	7	35	70	80	Low wavelength
NDL3210	NEC	6	675	9	34	50	60	
LN9705	Panasonic	5	788	10	35	40	50	
SDL5311	Spectra Diode Labs	100	830	10	30	35	170	Cost $800

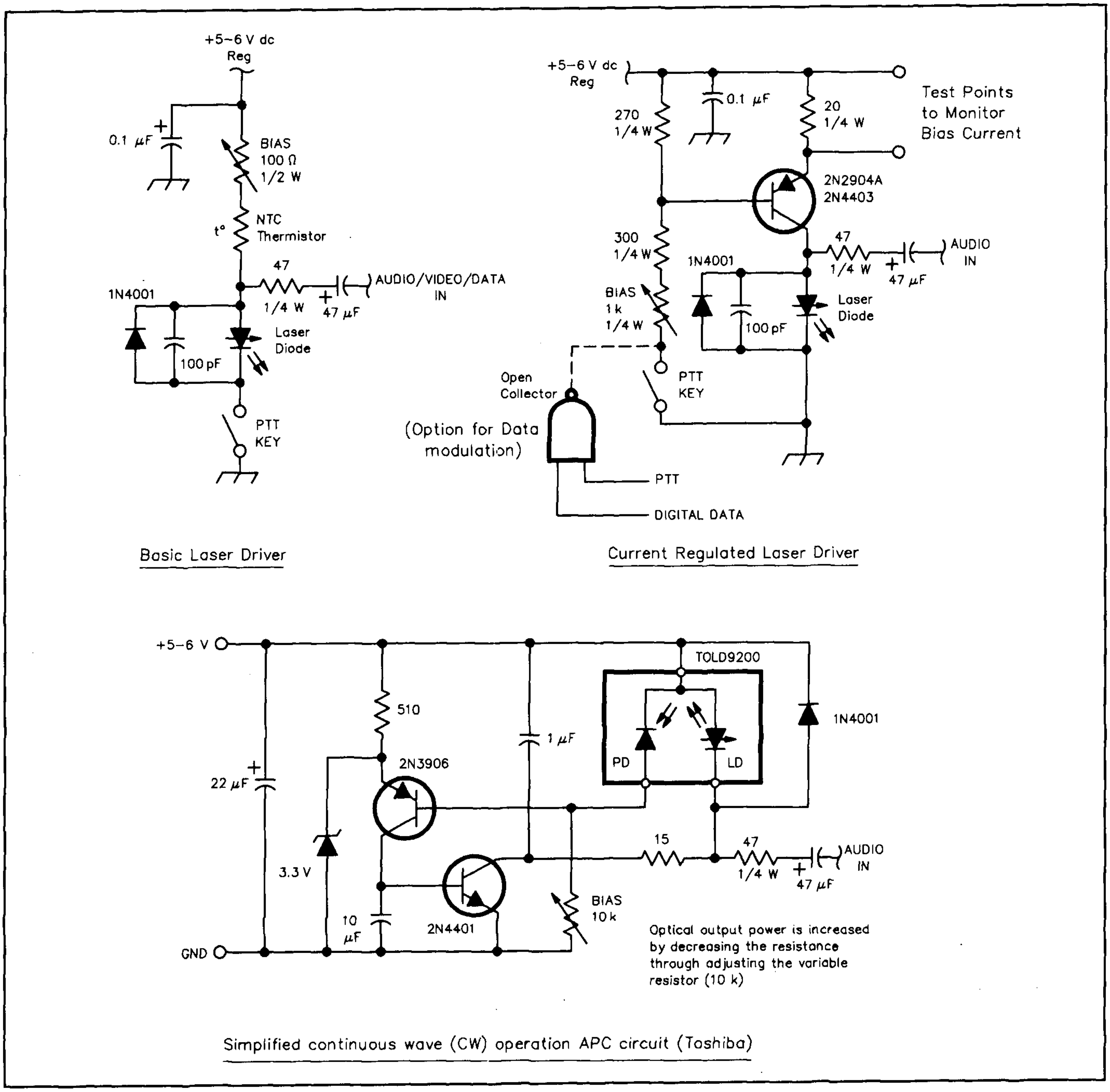


Fig 4—Laser diode driver circuits.

from the laser diode anode or cathode to a regulated voltage source of 5-12 Vdc. The modulating signal can then be capacitor or transformer coupled to the laser diode. A better method is to create a constant-current generator with a transistor or op amp (Fig 4). Temperature compensation can be easily included in this circuit. I do not use the monitor PIN photodiode built into the laser diode. It is commonly used in a feedback circuit (APC) to automatically stabilize the laser diode output over the temperature range. If you don't use the monitor diode, conduct initial transmitter testing with an inexpensive high-current red LED. Electrostatic discharge can destroy a laser diode. Ground yourself and your soldering iron when handling them.

LASER RECEIVERS AND OPTICS

The schematic of a direct-detection (non-superhet) free-space, optical receiver and its analog in a basic HF receiver appears in Fig 2. The receiver "antenna" is a telescope used to gather as much of the lightwave signal as possible. It can be a reflecting (mirror) or refracting (lens) telescope. Its main purpose is to focus the incoming light on the photo detector. From there, the light signal is converted or amplified, and converted to an electrical signal. The optical filter is analogous to the front-end tuned circuit, and is used to remove unwanted signals. The best overall choice for a telescope is a refractor using a Fresnel lens. This lens, composed of concentric engraved

Table 2
Comparison of Photodetectors

Device	NEP @ 670 nM $W\sqrt{H_Z}$	Gain	Detect Area cm^2	Power Supply V dc	Relative Size	Comments
PMT	$2\text{-}60^{-16}$	>10000	>4	>1 kV	large	fragile, fast
PIN	$0.2\text{-}9^{-14}$	none	0.01-2	0-100	small	ultra fast
APD	$1\text{-}9^{-14}$	200	0.01-2	0.2-2 kV	small	ultra fast
Photo diode	$0.1\text{-}1.0^{14}$	none	0.01-big	none	small-big	medium speed
Photo transistor	?	>1000	0.01	<50	very small	medium speed
Photo resistor	?	none	1-5	1-100	medium	slow

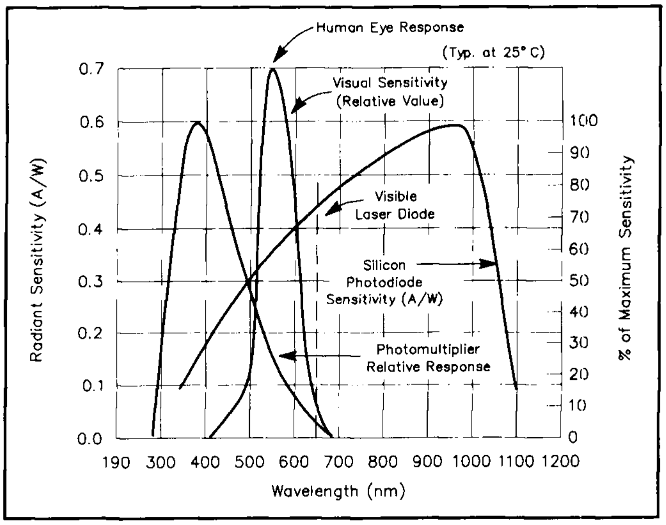


Fig 5—Photodetector responses at various wavelengths.

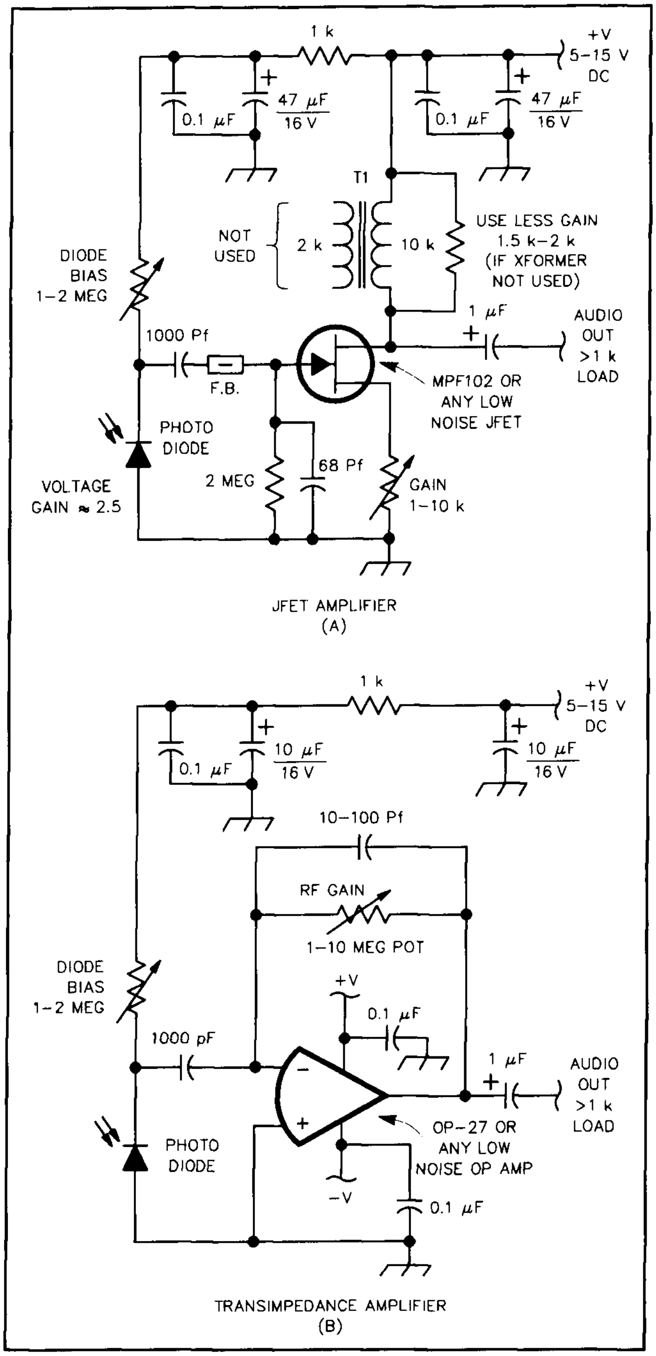


Fig 6—Low-noise photodiode preamplifiers.

rings and usually made of plastic, is light, thin and has high transmissivity. It is also surprisingly cheap for such large diameters. For example, a surplus 12-in. diameter Fresnel lens cost me $4 at a flea market. The focused image is not as good as a conventional glass lens, but the low cost and light weight of the plastic Fresnel lens are definite advantages.

I highly recommend filters selected for the wavelengths to be received. A narrow interference-type filter reduces stray light and allows laser communications in daylight. When working laser-diode stations, remember that, unlike gas lasers, some diode lasers may drift 10 nM up and down in wavelength with temperature and drive level. Variations from device to device may result in additional difference in wavelength of plus or minus 20-nM. The interference-type filter is sensitive to beam angle, and should be placed where the incoming beam is most collimated, usually in front of the lens. Loss is about 3 dB. Colored plastic and glass filters are cheap, but usually broadband, with losses of only 1 dB. Don't forget a lens cap for the telescope, to protect the lens and avoid frying the detector if the telescope is accidentally pointed at the sun!

The receiver detector can be any photodetector:

- Photomultiplier tube (PMT)
- PIN photodiode
- Avalanche photodiode (APD)
- Silicon photodiode and solar cell
- Phototransistor
- Photoresistor

Table 2 is a brief comparison of typical detectors. Because of its superior sensitivity and built-in gain, the PMT is the king of detectors for serious DX work. The typical PMT is blue sensitive. Its sensitivity drops off at the red and near-infrared (Fig 5). As you can see in the figure, the silicon PIN photodiode, followed by a low-noise, high-gain preamp is a

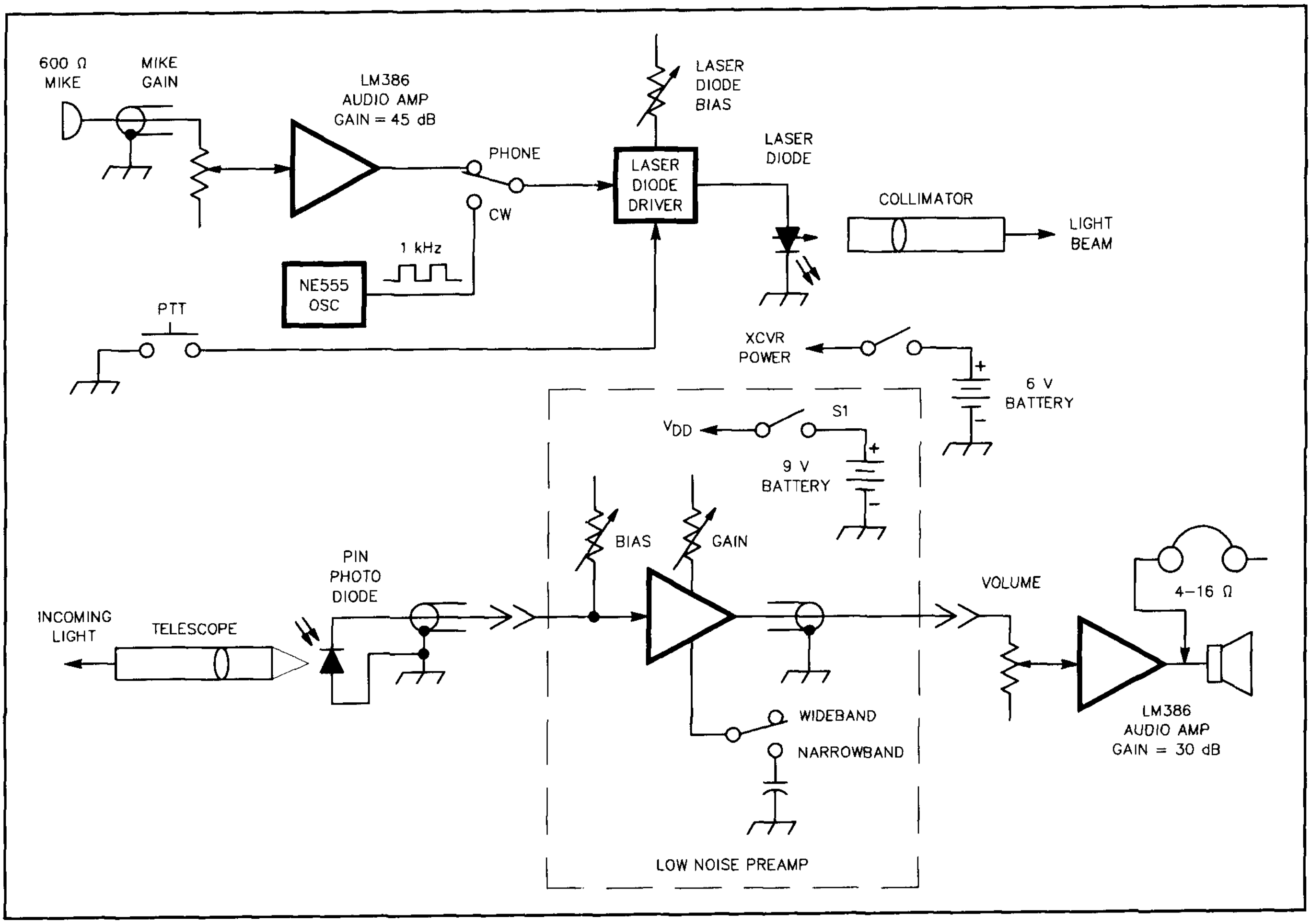


Fig 7—Block diagram of K6LMN laser transceiver.

practical second choice. It is the perfect mate for the red and near-visible infrared lasers. Its sensitivity peaks in the near-visible infrared region, and drops only about 20% at 670 nM. For lowest noise, the diode should not be back biased; for wide bandwidth (out to 1 GHz) though, the PIN diode requires 1-100 Vdc back bias. When avalanche photodiodes drop in price they will be more attractive, as they offer built-in amplification, like the PMT.

Typical silicon solar cells and photodiodes show peak sensitivity at 800 nM. A single, small solar cell chip at the focus of the telescope offers an inexpensive photodetector for the beginner. For a few dollars more, the highly sensitive PIN diode is a better choice for serious work.

Fig. 6 shows schematics of two common photodiode amplifiers. The transimpedance amplifier (B) is most useful for wide-bandwidth applications, and is commonly used with fiberoptic detectors. I like the JFET approach for a simple photodiode audio preamp, because of its low noise, resistance to self oscillation, and simplicity. A variable-frequency, variable-bandwidth filter should follow the preamp. A good choice might be one of the aftermarket audio filters. For optimum S+N/N ratio, limit the bandwidth as much as possible.

LIGHTWAVE TRANSCEIVER DESCRIPTION

Had enough theory? The K6LMN laser-diode transceiver provides push-to-talk, full-duplex voice/data/CW communication capability at visible light wavelengths in free space. The transceiver is a lightweight, compact (6- × 4- × 2-in.) package, with a self-contained power supply (4 AA cells). This unit is the heart of a complete lightwave station, including an "antenna" mounted on a 12- × 12-inch platform equipped for tripod or table mounting. A low-power red LED (665 nM) in parallel with the laser diode provides visual monitoring of the modulation. Fig 7 shows a block diagram of the station.

In addition to AM phone, the transceiver features MCW mode. With the key down, the laser is switched rapidly between 0 and 5-mW (full-power) output, by a 1-kHz square wave. The square wave is generated by a 555 IC. Average power output is thus 2.5 mW. An unmodulated CW mode may be added, to aid in visual sighting of the beam.

The laser transmitter output or lightwave emitter is a 5-mW peak, 670-nM (448 THz) laser-diode, integrated with a lens collimator. The collimator is adjustable, for divergence tweaking. A coherent and extremely narrow divergence (about 0.3-mrad high × 1.0-mrad wide, after the collimator) red light beam is the result. A die-cast aluminum case from a surplus laser printer serves as the diode heat sink. The laser-diode assembly is mounted on the transceiver cabinet. A low-astigmatism Toshiba 9211 laser diode is the emitter. It features a stable, single-transverse mode (clean spectrum) and low lasing-current threshold. A PIN photodiode integrated with the laser diode can be used to monitor and stabilize the power output, although I don't use this feature.

A 386 audio amplifier IC amplifies the 600-Ω mike input. Its high-level audio output amplitude modulates the current to the laser diode. The laser diode is purposely biased at the threshold of lasing, to reduce power consumption and the possibility of destroying the diode on voice peaks. The resulting modulation is slightly non-linear, but adequate for amateur voice/data communication. Only a small amount of temperature compensation is provided, and the bias point is not tightly regulated. An internal variable resistor sets the bias.

With the transmitter unkeyed, the current drawn from the internal batteries is about 10 mA. In transmit mode (no modulation), current consumption rises to 50 mA. On voice peaks, the current may reach 110 mA. The alkaline AA batteries used have a rated capacity of 1500 mAH. I use PTT, instead of operating full duplex, to conserve the batteries. Depending on path length and atmospheric clarity, it may take 15-30 minutes of continuous transmission and searching for the other station to complete a QSO.

For receiving, an "antenna" jack is provided, to accept audio input from any low-level-output photo detector. No dc bias is provided at this jack to power an active photodetector, such as a phototransistor or PMT. This transistor is designed to operate with solar cell detectors or sensitive photodiodes, followed by a preamp. The detector input circuit is protected against levels exceeding 1.5-V peak-to-peak, but less than 100 mA maximum. The input impedance is 10 kΩ. The detector input is amplified by another 386 IC. The output can drive 4-16 Ω external speakers or headphones, and a speaker jack is provided on the control panel.

The receiving antenna, which is mounted on the platform adjacent to the laser, is a super-cheap design. It is a 3-inch refracting telescope made of a free mailing tube, a $4 plastic Fresnel lens and a $5 silicon photodetector purchased at a swapmeet. The telescope mount is a bit flimsy, and needs refinement. The lens focal length is 12 in. The photodetector is a United Detector Technology model PIN 10RP438-1 silicon PIN diode, with a 1-cm^2 detection area. A broadband red bandpass optical filter in front of the detector filters non-red stray light, but can accommodate HeNe and near-visible infrared lasers. The photodetector is followed by a high-gain, low-noise JFET preamp mounted in a separate, shielded box. The photodiode is back biased with a variable voltage of 0-9 Vdc. A pot on the preamp front panel adjusts the bias for day or night operation with weak or strong signals. Another pot controls preamp gain. A front-panel switch sets the preamp bandwidth for voice or Morse code. Preamp output is connected to the antenna or detector input jack on the transceiver. My other transceiver uses a 15-V solar panel for an antenna, and has no preamp.

This transceiver is very simple to operate, with a minimum of controls. Although the transmit and receive channels are independent, crosstalk within the unit provides sidetone. The transceiver can be self tested by bouncing the beam off a nearby reflecting object. For sighting, I look along the transceiver case edge and point at the other station. Instead of using a spotting scope boresighted with the laser, I use a marked, vertical white pole placed several yards in front of the transceiver along the expected path. A heavy-duty, rock-steady tripod is an absolute necessity for serious DX work. Moving the beam a mere 0.06° moves the beam off axis 5 feet in one mile!

For phone operation, plug in a mike, flip the mode switch to Phone and push the PTT switch. Because the transceivers are full duplex, there is no transmit/receive changeover delay. The bandwidth of the voice channel is about 2.7 kHz (300-3000 Hz), which should be usable for data transmission.

THE FUTURE

I can see into the near future, where a lightwave repeater is a possibility. Bandwidth, as indicated by laboratory work being done with fiberoptic cables, is over 40 GHz. While the extremely sharp beam is suitable for point-to-point links, it doesn't lend itself to wide-area repeater use. The answer may be an array of semi-collimated laser diodes for the transmitter, and an array of photodiodes for the receiver.

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