BUILDING BLOCKS AND CONSTRUCTION

A Simple T/R Sequencer

Protect sensitive RF circuits from burn out by sequentially switching from transmit to receive.

By Zack Lau, W1VT

(From *QEX*, October 1996)

What, another T/R sequencer? Yes, I decided to make it as simple as possible, while adding +12-V outputs and reverse polarity protection. I think the reverse polarity protection is pretty nifty—who hasn't worried about hooking up batteries backwards and frying something? The 12-V outputs make it real easy to wire up my transverter designs, since they are made up of modules that run off 12 V. Hopefully, this design will spur more designers into adding reverse polarity protection.

The primary reason for using a sequencer is to protect the RF relay and the amplifiers hooked up to it. RF relays can be damaged by hot switching at high power levels. Unfortunately, what constitutes high power is rather fuzzy—I have not found any good references that adequately address this topic. Based on my experience with microwave transverters, I always put in a sequencer when switching more than half a watt. I don't bother with them in inexpensive systems running less than 100 mW.

Like the T/R sequencer designed by Chip Angle, N6CA, this sequencer uses a quad comparator to monitor the voltage on a charging or discharging capacitor.[1] I looked at using an integrator to get more uniform delay intervals, but decided to stick with the simpler circuit. An integrator can generate a nice triangular waveform, as opposed to the exponential curve generated by an R-C network. The 10-Ω resistors are used to provide hysteresis, so the outputs don't switch back and forth unnecessarily near the transition point. They are supposed to provide a little positive feedback, instead of the more common negative feedback used in other applications.

This sequencer first turns off the receiver, then activates the relay, amplifiers and finally the transmit IF outputs when switching from receive to transmit. When going back to receive, it turns off the transmit IF, amplifiers, and relay before reactivating the receiver. The idea is to introduce enough delay between these states to allow everything to settle down, and to reverse the order when switching from transmit to receive. This is especially important with a mechanical relay which may make intermittent contacts for a few ms when switched.

A big advantage to using solid-state amplifiers is that you can switch them off during receive. Not only does this reduce the possibility of hot switching, it reduces the chance of amplified broadband noise getting into the receiver. Using a PIN diode switch to cut off transmit drive also helps to prevent hot switching—I use the transmit IF signal to control this switch.

Since I've never needed to change the polarity of the output of one of my sequencers, I decided some simplification was in order. Instead of the XOR gate Chip used, I decided to use a hard-wired switch based upon the principles of a transfer switch. To switch polarity with this design, you change a pair of resistors from horizontal to vertical, or vice versa. By labeling the blank area of the board between the resistors, identifying the resistors ought to be straightforward. The reduction in parts count ought to enhance reliability.

I've also taken advantage of the improvements in switching transistor technology. The International Rectifier P-channel IRF 9Z34 will easily switch 2 A, enough to power a 5-W GaAs FET power amplifier. Similarly, the Zetex ZTX 789 in a little TO-92 style case will actually switch a small SMA relay that draws a few hundred mA. The bonus to using more expensive PMOS/PNP parts is that the switched supplies are reverse polarity protected. I just needed to protect the comparators with a diode and use bipolar electrolytic capacitors which don't care about voltage polarity. Cheap 2N3906s are used for the RX and TX IF supplies, since they typically draw under 100 mA. You could use ZTX 789s instead of the 2N3906s, with the appropriate bias resistors, for higher current.

You may want to replace Q2 with a VN10LP N-channel FET. This will allow you to hook up the PTT line without pulling the voltage down significantly. Another

[1]Notes appear at the end of this section.

advantage is that you can now hook up the PTT input to a stiff voltage source without frying Q2. But these FETs aren't as easy to get as common NPN switching transistors.

Of course, it makes sense to watch out for even newer FETs, which will undoubtedly offer better performance at lower cost. There is even a trend toward lower gate thresholds, which allow better performance at lower voltages. For instance, the IRF 7104 drops only 154 mV when sourcing 0.57 A ($V_{gs} = -5$ V). But, despite the marketing hype, you can do even better than "full enhancement." It only drops 117 mV with a V_{gs} of -10 V.

Construction

I made the pads big enough to accommodate swaged terminals, which are a really nice way of making dc connections if you have the tooling to rivet them to a fiberglass circuit board. The board is a bit crowded—I wanted the board to fit nicely on the wall of a chassis box only two

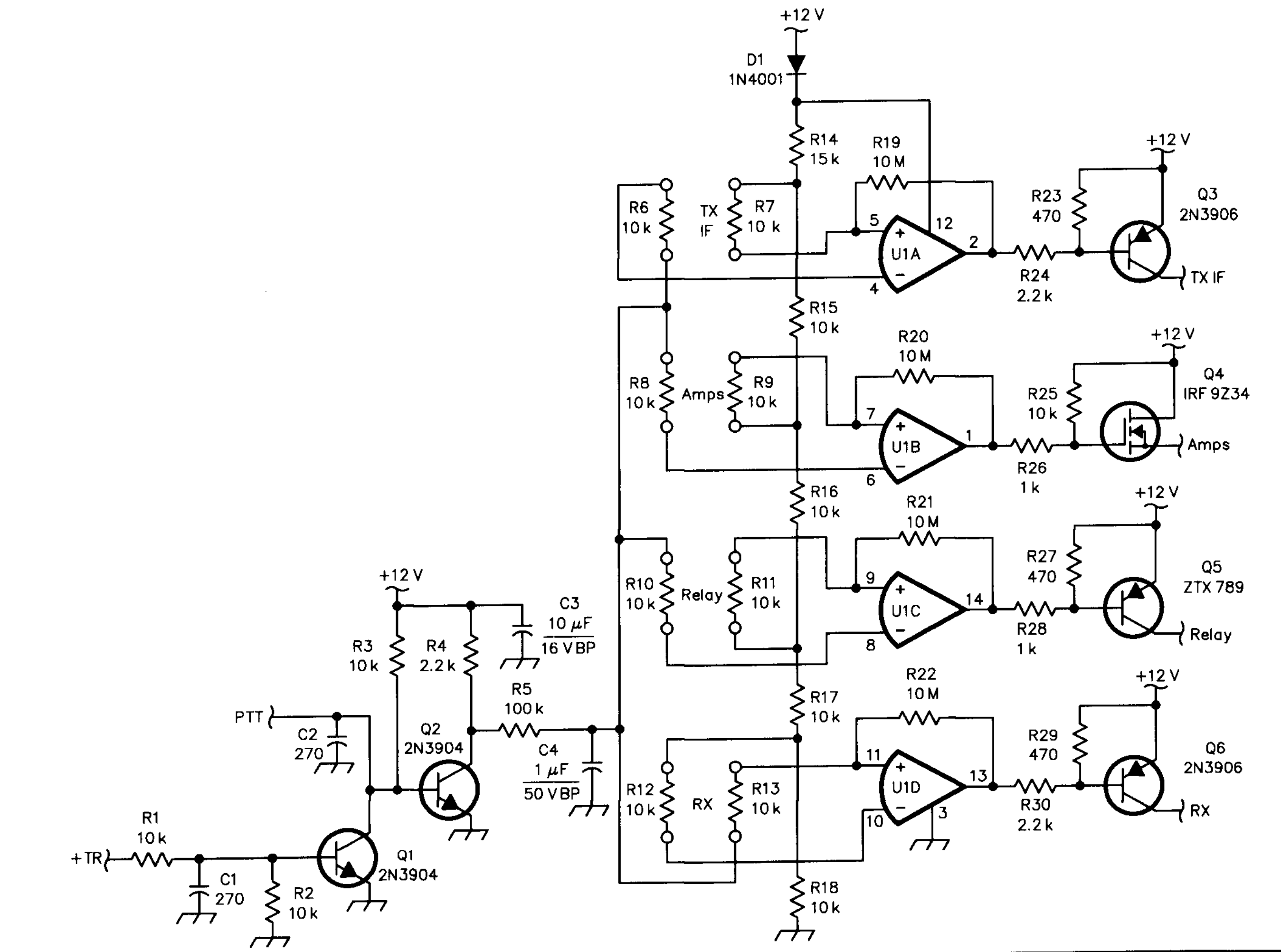


Fig 1—Schematic diagram of the transmit-receive sequencer.

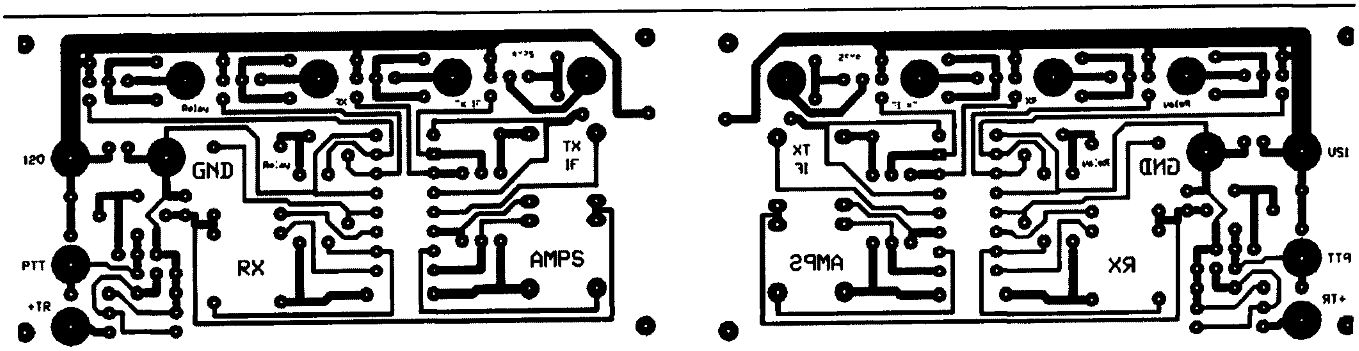


Fig 2—Etching pattern for the transmit-receive sequencer circuit board.

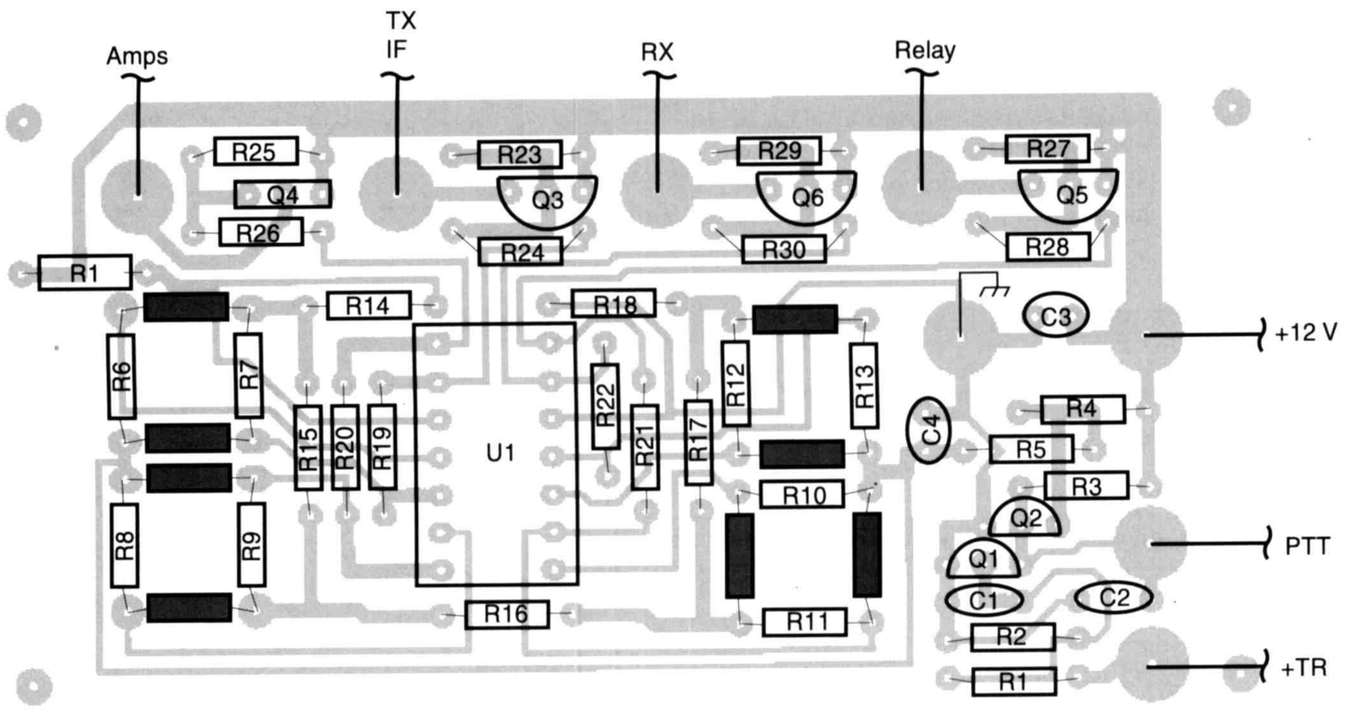


Fig 3—Parts placement diagram for the transmit-receive sequencer circuit. The shaded resistors indicate alternate positions for R6 to R13 to invert the signal sense. See text.

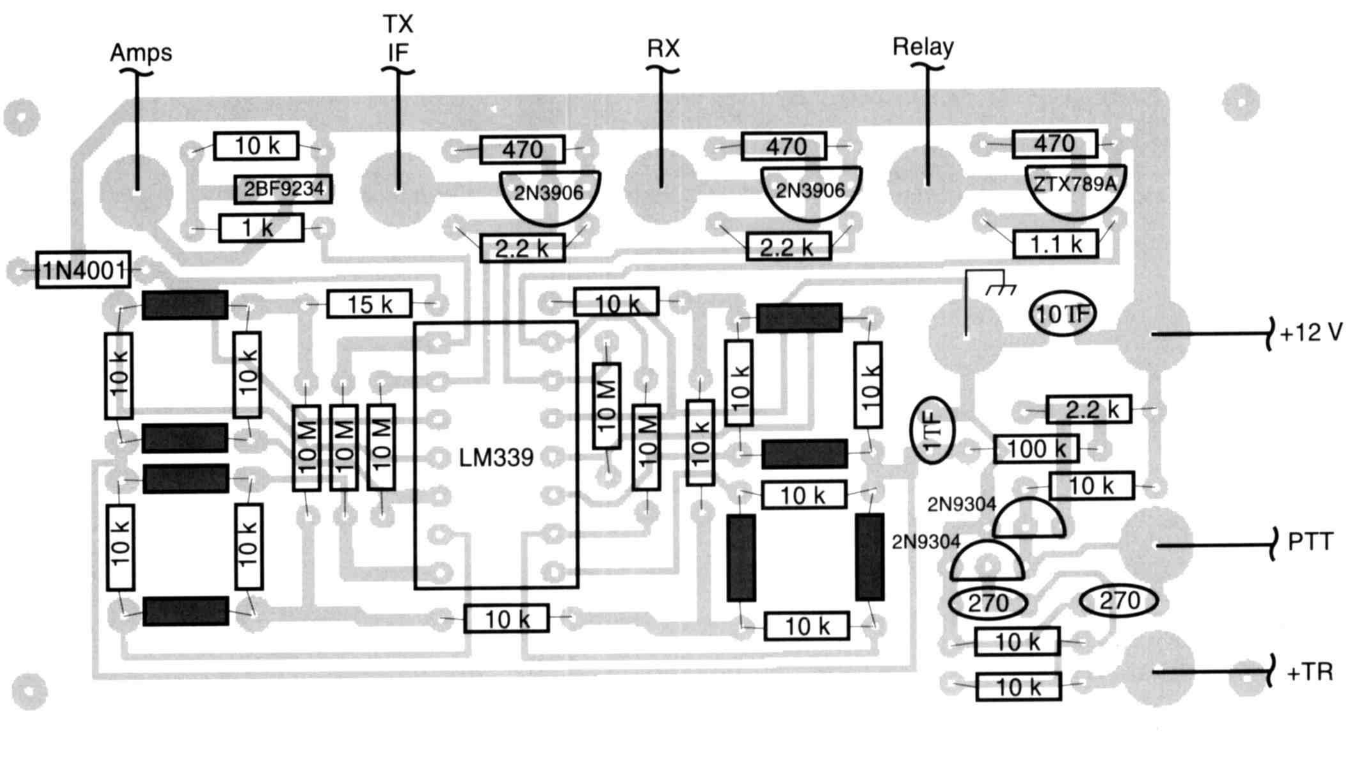


Fig 4—Parts placement diagram using component values and part numbers.

inches high. A mirror image of the etching pattern is provided—it simplifies the toner transfer process that some people use to make circuit boards. Similarly, a parts placement diagram using part values and component designations is also provided in Fig 4.

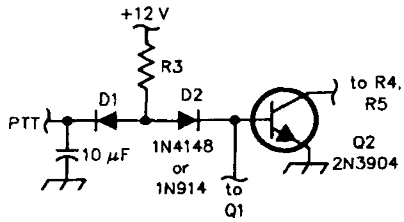
To test the board, I made a little fixture out of LEDs and dropping resistors. A separate fixture makes it easy to line up the LEDs in the proper sequence. It may be useful to slow down the sequencing by bridging the 1-µF timing capacitor with a 10-µF capacitor. This makes it easier to see the LEDs turn on and off.

Notes

[1]Angle, Chip, N6CA, "TR Time-Delay Generator," *The ARRL Handbook for Radio Amateurs*, 1997, pp 22.53-22.56.

Feedback

Unfortunately, the T/R sequencer doesn't actually protect its outputs from reversed polarity, though the sequencer itself survives reverse polarity. Incandescent test lamps instead of LEDs are recommended for anyone experimenting with



such circuits. The problem isn't easy to solve–power FETs contain a parasitic diode across the junction. Normally, protective circuits get around this problem by reversing the source and drain connections, but this doesn't work here. That fix would disable the use of the FET as an on/off switch–it would be effectively on all the time.

Mario Miletic, S56A, suggests the improved PTT input circuit above. He also suggests that the decoupling capacitors C1 and C2 be increased to 10 nF, though the author thinks that 270 pF is more appropriate for a VHF/UHF/SHF station.

A "Fool-Resistant" Sequenced Controller and IF Switch for Microwave Transverters

Feeling foolish since you blew up that nice new transverter because the T/R switching wasn't sequenced right? Here's a way to avoid that problem.

By Paul Wade, N1BWT

(From *QEX*, May 1996)

Last summer, I suffered the failures of two 10-GHz preamps and one coax relay. Fortunately, none occurred at critical times, and I finally rigged up an inconvenient but safer two-switch scheme to prevent further problems. But I did resolve to come up with a better control system this winter. Ideally, it would be fool-proof, but fools are too resourceful for that, so I've tried to make it as fool-resistant as possible.

Discussion

For several years, I've been using variations of a transverter IF switch by KH6CP.[1] This has worked well in several of my transverters, and I've made various improvements, but it does not adequately sequence various switching functions.

Three sequencing techniques are commonly used. The first is to intercept the PTT line so the transceiver is controlled by the switch box. Often, this requires modification of the transceiver, particularly those that do break-in CW only, transmitting when the key is touched. I want to be able to interchange various transceivers without modification so I can lend spare equipment to willing rovers.

The second approach uses a fixed-sequence switch, usually a series of time delays, which, once started, go through the sequential operations without further safeguards.

The last, and least successful, method is to switch an external relay directly from a transceiver's PTT line. Often, the current available from the PTT line is inadequate for driving a relay, and I know of several cases where the transceiver has been damaged by this technique.

My preference would be a switch that goes through sequential operations but checks that appropriate conditions are met before proceeding to the next step—in logic design, this is called a *state machine*. As I started to sketch out the sequence of operations I wanted, I realized that ordinary T/R switches, such as relays, don't have an appropriate state to deal with break-in transceivers, which are delivering RF before the switch is ready for it. To deal with this, I use a PIN-diode IF switch and have designed the controller to have a third state, in addition to transmit and receive, in which all applied RF power is absorbed. I call this third state the *safe* state. Since one of the functions of the IF switch is to attenuate the transmitted power from the IF transceiver, the safe state is implemented by adding two PIN diodes to absorb the power.

Design

The first step in the design process is to sketch out the desired timing for the switching sequence. This evolved to the timing diagram shown in Fig 1, which goes through one cycle from receive to transmit and back to receive. The second step is to synthesize a logical state machine that generates the desired timing. The final step is to actually design a circuit that implements the logical state machine. Following this progression helps to ensure that the final circuit will operate as intended since there is a clear target to work toward.

The desired operation of the state machine sequence is shown in the *state diagram*, Fig 2. The system starts in the inactive receive state. When an activation signal is received, the system moves to the *safe* state, absorbing all RF power. The switching sequence can then continue at whatever speed is required, not releasing the RF power to the transmit circuitry until the system is ready to go to the transmit state. Normally, this would mean removing power from the receive section, then driving the microwave T/R relay, waiting long enough for it to switch (or to sense that the fail-safe contacts closed, if you are fortunate enough to have a relay with this feature) and finally, applying power to the transmit section. When the activation signal is removed, we go from the transmit state to the safe state, reverse the switching procedure, then return to the receive state.

Since this state machine is intended to be used in several transverters and with various IF transceivers, I added some options to increase flexibility:

1. RF sensing ensures that any RF power applied to the IF port will cause switching to the safe state—even if no control signal is applied—to protect the transverter from

[1] Notes appear at the end of this section.

damage. Full RF switching may be enabled by setting the J17 jumper, allowing the use of any transceiver, even a hand-held, for the IF.

2. PTT polarity selection is provided since some transceivers ground the PTT output on transmit while others provide a positive voltage. There are separate inputs for these two PTT polarities; each requires low current and has a switching threshold of about 5 V.

3. Single-cable switching supports transceivers that put the dc PTT voltage on the RF output cable. Jumper J3 sets the PTT polarity for the IF cable.

4. A transmit-ready signal can be sensed. Some amplifiers require a warm-up period, so this input must be grounded to indicate that everything is ready to transmit; otherwise, the switching sequence will remain in the safe state and not continue. This input could be automatically or manually switched.

5. Fail-safe sensing detects when the fail-safe contacts on a coax or waveguide relay have closed—and prevents transmitting until they do. I've not yet found a good coax relay with this feature, but my 10-GHz waveguide relay does have it. Jumper J15 selects between fail-safe operation and time-delay-only sequencing.

6. FET output drivers for the safe state activate a coax relay, activate dc power switching and drive LED indicators for the operator. I like to have three LEDs: TRANSMIT READY, SAFE, and TRANSMIT.

7. A DPDT relay may be jumpered to be switched by any of the FET drivers to operate at the desired point in the desired sequence.

The components for each of these options are indicated on the schematic diagram, Fig 3, and may be populated as desired.

Circuit Description

Some have suggested using a small microprocessor to implement switch sequencing. The flexibility and programmability of this approach would be great, but I am very cautious about putting microprocessors in high-intensity RF fields. Since my intent is to include both the PIN-diode switch and the controlling state machine in a small metal box, there may be a significant amount of RF in the box. Therefore, I chose to design using components that are cheap, proven and readily available, and are also slow enough not to respond to RF. And wherever possible, I use these components in circuits I have used before and know to work well.

Let's take a quick tour of the schematic diagram, Fig 3. The IF transceiver connects to J1, and its transmit power is reduced by the attenuator, R1, R2 and R3. The values shown provide about 14 dB of attenuation. Since low-inductance power resistors are becoming hard to locate, it may be necessary to adjust the attenuator design to fit the available component values. I described how to do this using a computer program, PAD.EXE, in *QEX*.[2] The program, which calculates resistor values and power ratings for attenuators, is available from the *QEX* Web site, **http://www.arrl.org/files/qex/** in file qexpad.zip. The input attenuator used here is designed for the 2 to 3-W output available from small portable transceivers.

The attenuator is followed by the PIN-diode switch. A PIN diode acts as an RF conductor when dc is flowing through it, but acts as an RF open circuit when reverse-biased. Each PIN diode in this circuit is supplied with +6 V at one end, so the other end may be switched between +12 V and ground to reverse the bias. D1 and D2 select the transmit or receive path on the IF side, while D5 and D6 select the path on the transverter side. The transmit path goes through an adjustable attenuator that can be adjusted for 20 to 38 dB of total transmit attenuation. This is needed because most mixers require around 1 mW or less of power. The receive side uses an MMIC amplifier stage, A1, to overcome the loss of the input attenuator—the MAR6 provides enough gain to end up with 6 dB of net gain ahead of the transceiver, with a noise figure better than that of most transceivers.

The safe state is provided by PIN diode D4, which shorts the output end of the attenuator. Turning off FET Q4 causes current to flow through D4, making it an RF conductor, and causes D5 to be reverse-biased, making it an open circuit for RF. Thus, RF flowing into the transmit path has no output path and must be dissipated in the attenuators. The reflected power must pass through the attenuator twice, for a total loss of 60 dB, so essentially no reflected power is seen by the IF transceiver.

An additional safety feature is provided by D3, which is turned on by FET Q2. D3 shorts out any transmit energy that leaks through D2 (when it's off) and also disables MMIC amplifier A1 by reducing the dc voltage supplied to it.

The switching states for the PIN-diode switch are straightforward: in the receive state, FET Q3 is turned on, which

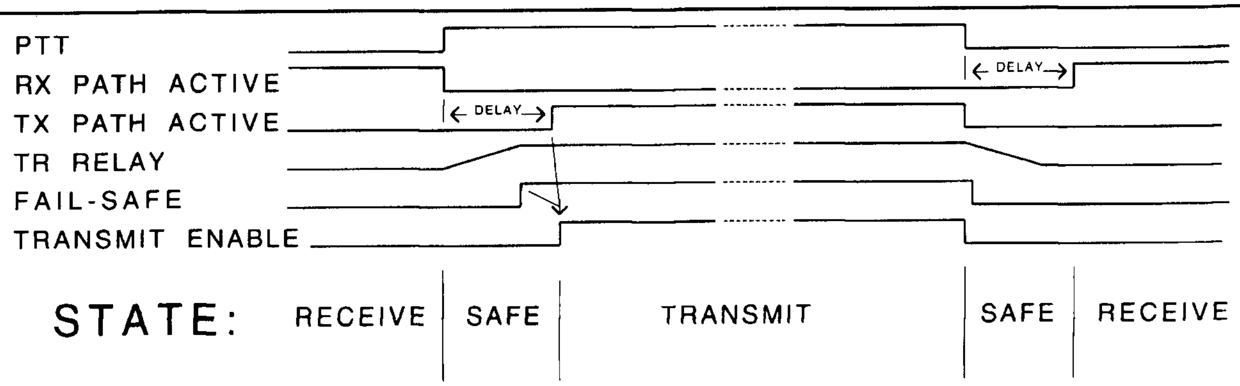


Fig 1—Sequencer timing diagram.

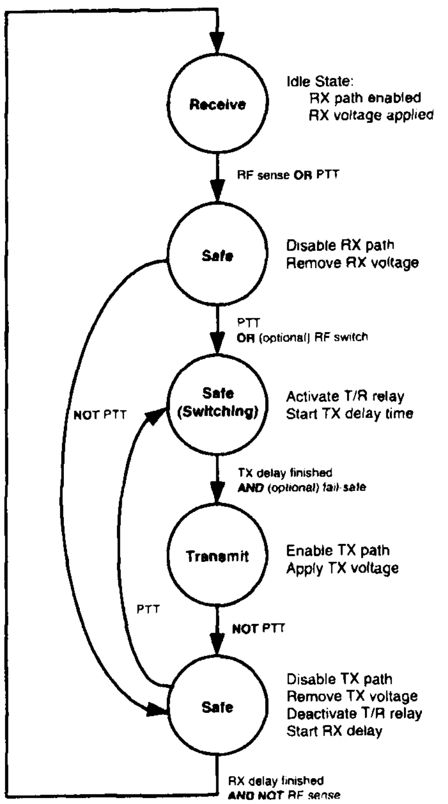


Fig 2—Sequencer state diagram. The circles show the individual states, while the text to the right of each circle shows the actions performed upon entering the state. The text next to each connecting line shows the conditions necessary to advance to the state the line goes to.

causes D6 to turn on. FETs Q1, Q2 and Q4 are turned off, so D2 and D4 are on while D1, D3 and D5 are off—only the receive path is active. The next state is the safe state, reached by turning on FETs Q1 and Q2 and turning off Q3; this turns on D1 and D3 while turning off D2 and D6, so the IF input is switched to the transmit side but the output side is not connected and D4 shorts the attenuator output. Finally, the transmit state is reached by turning on Q4, which turns off D4 and turns on D5, completing the transmit path.

With the resistor values shown, the PIN-diode currents are adequate for an input power level of about $^1/_8$ W, so the input attenuator should reduce the IF power to this level or lower. For higher powers, it would be necessary to increase the on current through the diodes, particularly D1. However, it shouldn't be necessary to switch much power, since the RF output to the mixer input should be milliwatts or less.

All the FETs used in this circuit are N-channel enhancement-mode MOSFETs used as switches. The sources are all grounded and the gate is the control element. When the gate voltage is close to the source voltage, or ground, the FET is turned off, and no current flows from drain to source. To turn the FET on, the gate voltage must be several volts more positive than the source voltage, allowing current to flow from drain to source with only a few ohms of resistance. For practical purposes, we may consider the drain to be shorted to ground when the FET is on and open circuited when the FET is off. Since these are insulated-gate FETs, no gate current is possible and no dc power is required for switching.

The gate voltage required to switch the smaller FETs is roughly 2.5 to 3 volts, but larger power FETs such as Q14 require a somewhat higher voltage, so the whole switching circuit operates at 8 V, provided by the three-terminal regulator, IC3.

The rest of the schematic describes the control logic. The RF-detect section, from C13 to Q5, drives IC1A to force the system to the safe state as soon as RF input is detected. The PTT section, from Q6 and D9 to Q7, is a DTL (diode-transistor logic) gate with a switching threshold set by Zener diode D10. The PTT output is inverted by IC2A to also drive IC1A and force the safe state. Note that IC1A is drawn as an OR gate, with inversion bubbles on the inputs to show that they are asserted low; thus the output of IC1A is asserted when either input is in the low, or asserted, state. The output of IC1A is inverted by IC2B which activates OR gate IC1B, thus driving the *RX Disable* signal to turn FETs Q1 and Q2 on and put the PIN-diode switch in the safe state. The output of IC1B is also inverted by IC2C to drive the *RX Enable* signal. This turns FET Q3 off when Q1 and Q2 are turned on, and vice-versa. Finally, IC1B also drives FET Q10, which is turned on in the safe and transmit states so it may be used as a signal to control the voltage supplied to receive stages and preamps.

The PTT section has two inputs, PTT-L on J4 and PTT-H on J5. PTT-L must be grounded, or asserted low, to activate, while PTT-H requires a positive voltage, or high assertion. Both inputs have an operating threshold in the 2 to 5-V range, so any input voltage below the threshold is considered low and any input above the threshold is considered high. The high threshold provides considerable tolerance for different rigs, dirty contacts, etc. The PTT section can also be activated through the IF cable input on J1—any dc voltage on J1 is delivered to the logic circuit through RFC2. Jumper J3 selects the polarity for the IF input; the right-hand position selects PTT-L and the left-hand position selects PTT-H.

The transmit-ready section, from J6 to Q8, is another DTL gate. Its output drives IC1C, which is drawn as a NAND gate; both inputs must be asserted high for the output to be asserted low. The other input to IC1C is selected by jumper J17; in the lower position, it is the output from the PTT circuit. Thus the IC1C logic function requires both transmit ready and PTT to be asserted. The upper jumper position takes the output from IC1A, which also includes the RF detection, making the logic require both transmit ready and either PTT or RF detect. This allows switching using only RF detection. Capacitor C15 sets the hang time for RF switching. With the values shown, switching time seems fast for SSB or for slow CW, so a bit of experimentation might be needed to find a time that feels right.

The output of IC1C is inverted by IC2E (note the inversion bubble on the input, to match the output of IC1C which is asserted low) to drive FETs Q13 and Q14, one of which should be used to enable the T/R relay. The IC1C output also drives FET Q11, which is an inverter with a time delay set by R28 and C18. Q11 drives the *TX Enable* signal, so completion of the time delay turns on FET Q4 to allow the transmit power to flow through J2. When jumper J15 is in the upper position, the completion of the time delay will also allow IC1D to switch, driving Q12 and enabling the transmit state. The lower position of the jumper forces IC1D to wait until J14 is grounded by the fail-safe contacts on the T/R relay.

When PTT is released and no RF is detected, the output of IC1A is deasserted. This voltage transition passes through FET Q9, an inverter with a time delay set by R24 and C16. Until the time delay completes, pin 13 of OR gate IC1B remains asserted, keeping the PIN-diode switch in the safe state while all the other switches are released. Since the safe state prevents any RF from getting through, sequencing of the switches isn't critical in this direction.

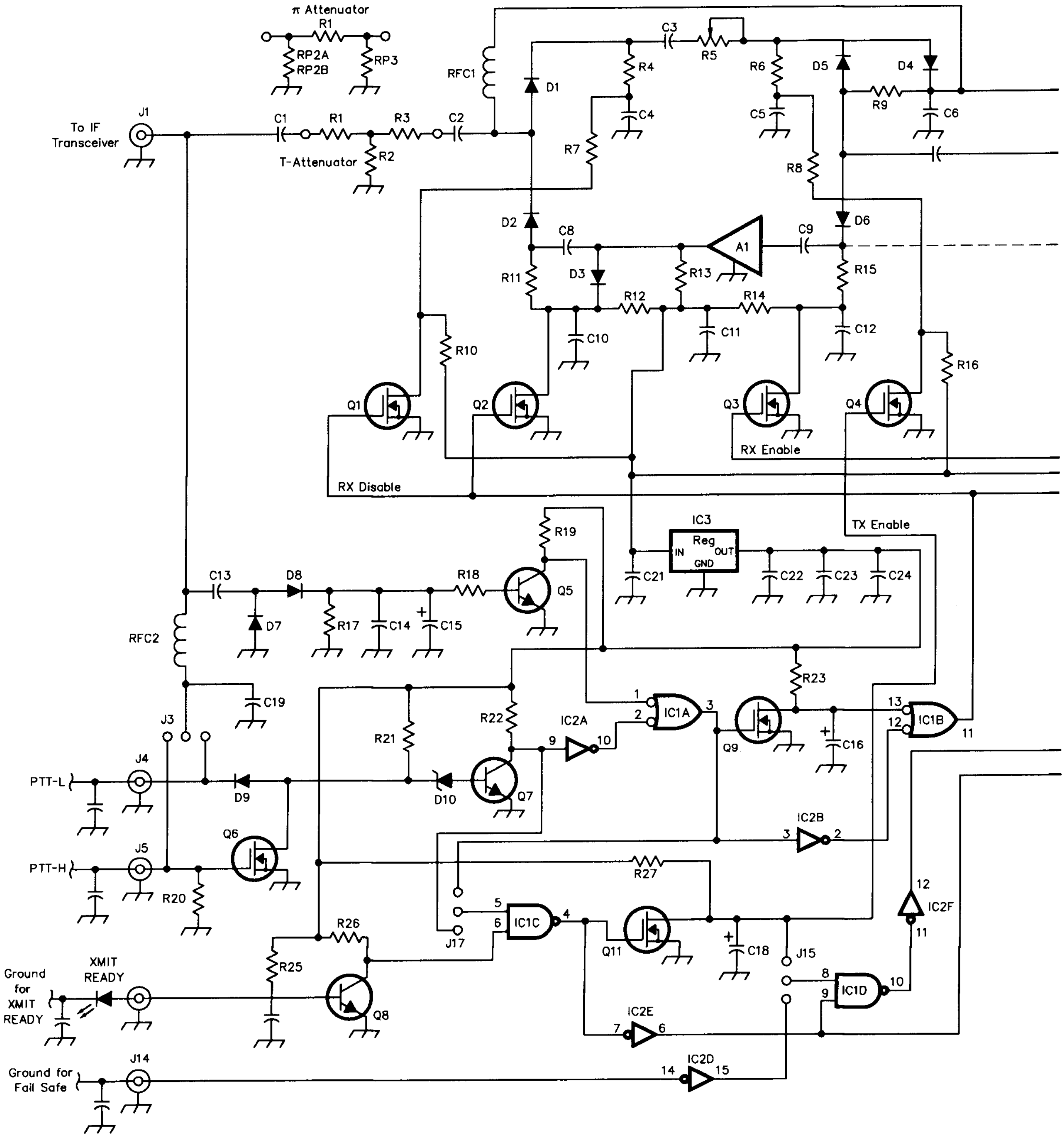
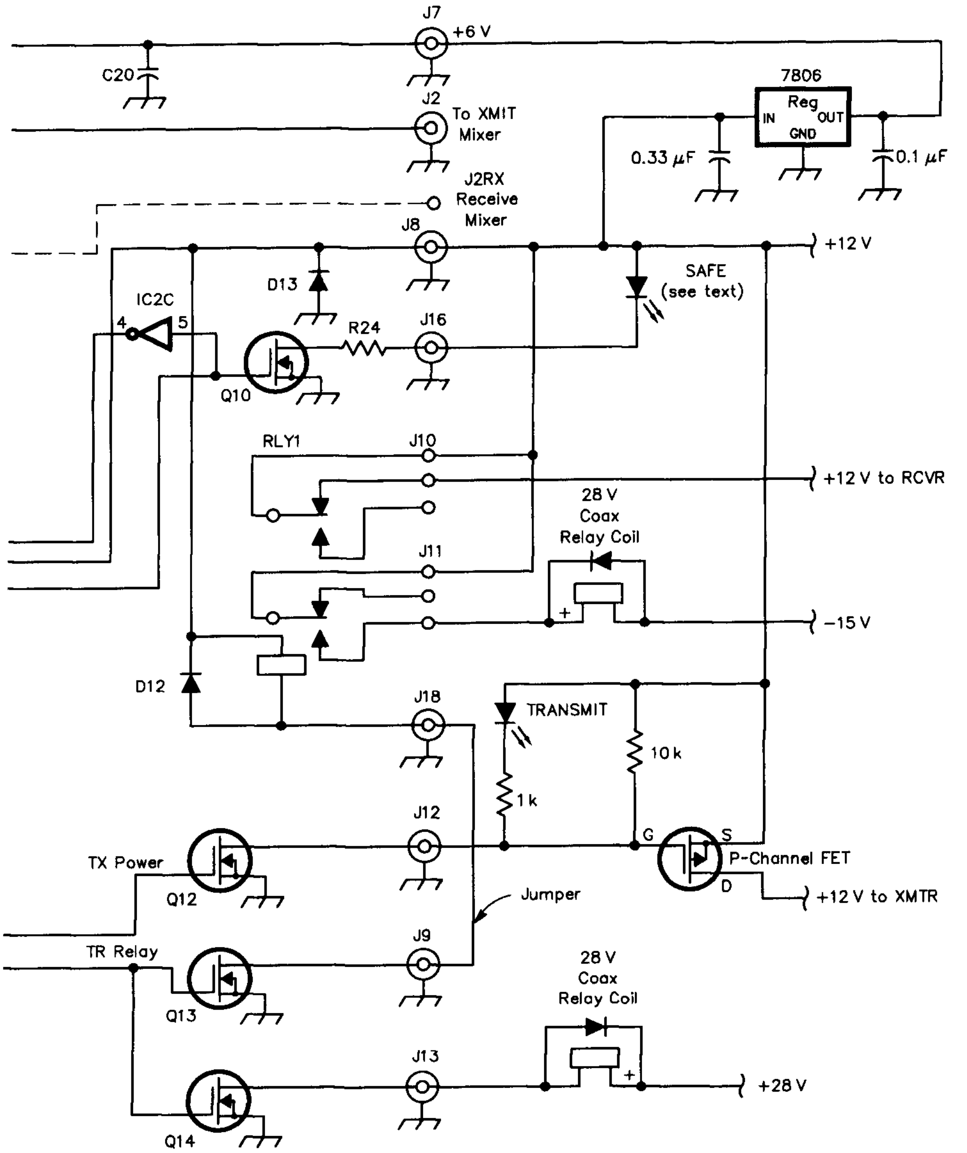


Fig 3—Schematic diagram of the IF switch and sequencer. (See Table 2 for parts list.)

Finally, IC3 regulates the logic voltage to 8 V to maintain constant time delays. The R and C values specified yield time delays of 200 to 300 milliseconds, but the delay can be increased or decreased by changing the values. For instance, increasing C16 from 10 µF to 16 µF would increase the time delay by about 60%. Alternatively, increasing R24 from 33 kΩ to 51 kΩ would have the same effect.

Relay RLY1 may be driven by Q13 to operate at the same time as the T/R relay, during the safe state, or driven by Q12 to operate when entering the transmit state. Notice diode D12 across the relay coil. This serves to protect the FET from the reverse voltage spike caused by removing the current from the relay coil. All relay coils should have a diode to protect the driving circuitry; even a relay driving another relay can suffer contact damage from the switching spike.



Construction

I decided that this circuit is complex enough to justify layout of a printed-circuit board since my intent is to use copies in several transceivers. All the components between the two vertical rows of jacks on the schematic diagram, Fig 3, fit on the PC board. A double-sided board with plated-through holes was needed for full interconnection; the top-layer pattern is shown in Fig 4, and the bottom layer is in Fig 5. Boards are available from Down East Microwave.[3]

All the chip capacitors are mounted on the bottom of the board. I chose to put the PIN diodes on the bottom also to keep lead lengths short in the RF path. All of the other components are on the top side of the board, as shown in Figs 6 and 7. Note that the smaller power FETs, such as Q1, have inconsistent pin-outs that vary with part number and manufacturer. Check the data sheet, and make sure that the source lead connects to ground, which is the wide trace running all over the top of the board. The gate lead connects to the middle pad of each footprint, leaving the drain at the far end.

Component values are not critical. I've tried to calculate optimum values, but any resistor or capacitor value could be changed to the next higher or lower standard value without significant effect. The RF diodes are stocked by Down East Microwave. All other components are readily available from Digi-Key.[4] The cost of all components totals less than $15, not counting the enclosure box and connectors.

The PC board is sized to fit inside a small die-cast aluminum box since a shielding enclosure is highly desirable.

Application

There is enough flexibility in this circuit that using it requires some decisions; on the other hand, it should be possible to fit it to your system needs rather than forcing the system design to match the controller. The portions of the schematic diagram outside the two vertical rows of jacks show some of the possible functions.

The first decision is whether the transverter uses a single mixer, as shown in Fig 8, or separate mixers for transmit and receive, as shown in Fig 9. A single mixer would connect to J2; otherwise, the transmit mixer connects to J2 and the receive mixer connects to J2RX, which is a hole in the PC board next to C9. In this case, D6 and R15 must be removed, and Q3, C12 and R14 may be removed or used for another switching function, as described below.

The next decision involves the control signals. I usually provide inputs for both polarities of PTT using different connector styles (RCA phono for PTT-H, subminiature phone for PTT-L). The transmit-ready and fail-safe inputs can go to connectors if they are used. Otherwise, they should be jumpered to ground to avoid floating inputs. Finally, jumpers J3, J15 and J17 must be installed as described in the circuit description. The switch will not operate without these jumpers.

Finally, we must decide how to use the control outputs. I chose to only provide outputs grounded by FET switches, except for the floating relay contacts, to keep unwanted voltages off the board. The signals that drive the FET switches are labeled on the schematic to indicate function. Some possibilities that I have used are shown in the right-hand side of the schematic. The internal relay, RLY1, can be driven by jumpering J18 either to J9, timed to switch a coax relay, or to J12, timed at the transmit state of the sequence. An external coax or waveguide relay usually requires 28 V for operation, which can be provided from a +28-V supply and switched with the larger power FET Q14, or connected between +12 V and a −15-V supply and switched with the internal relay contacts since many transverters already generate a negative voltage internally.

Power for the transmit stages may be switched with the internal relay contacts or with a solid-state switch using a P-channel power FET like the IRF-9130 or IRF-9530, which can switch several amperes with a small voltage drop. Note that the P-channel FET is used "upside-down," with the positive voltage connected to the source, as shown in the schematic, since a P-channel FET operates using voltages opposite those of the N-channel FETs described above.

Receive stages and preamps may be switched in several ways. The schematic shows the simplest, using the internal relay to disconnect the voltage at the same time that the T/R relay operates. A more robust sequence would be to remove power when entering the safe state; FET Q10 would be the appropriate driver, with R24 replaced with a jumper so the connection is directly to J16. FET Q10 is turned off during receive and on in all other states. If the receive voltage is set by a variable three-terminal voltage regulator, connecting the adjust pin of the regulator to Q10 would turn off the regulator output. Another alternative, for transverters with separate mixers for transmit and receive, would be to use FET Q3, which turns on during receive and off in all other states, the inverse of Q10. In the two-mixer configuration described above, Q3 is not needed, and R14 and R15 are removed so their pads are available as connection points.

Of course, one needn't be constrained by the printed wiring. If one of the FET switches is not used for the function shown, it can be used for a different function by connecting its gate to the appropriate switching line. All it takes is a hobby knife to cut the trace and a soldering iron to add a wire.

LED indicators may be driven by any output and can be driven by the same FET that drives a relay since the additional current is small. The schematic shows a TRANSMIT READY LED in series with J6, so grounding the transmit-ready line draws enough current to light the LED. If there is no LED in this line, R25 could be much larger to reduce current drain.

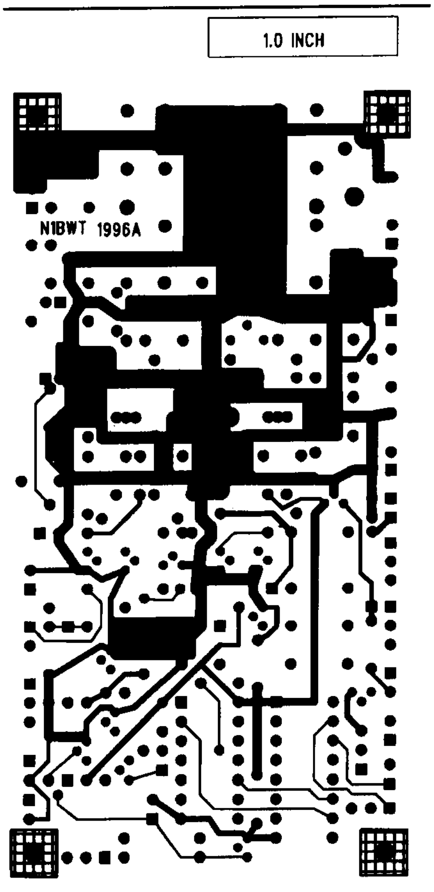


Fig 4—Top-layer printed-circuit board pattern.

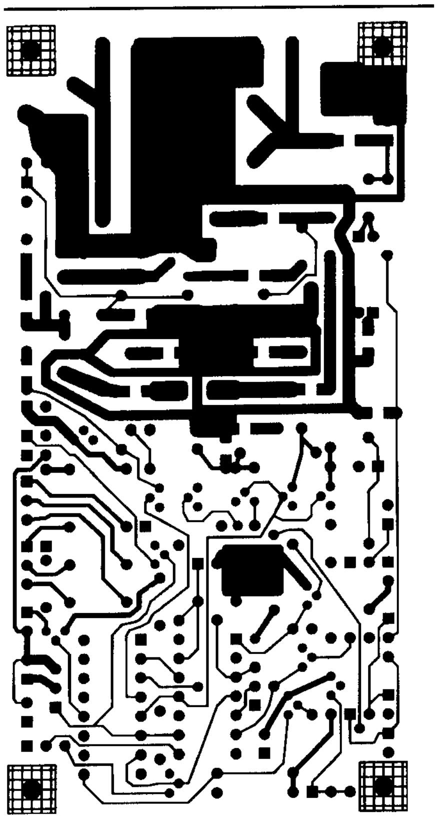


Fig 5—Bottom-layer printed-circuit board pattern. (Note: Board requires plated-through holes.)

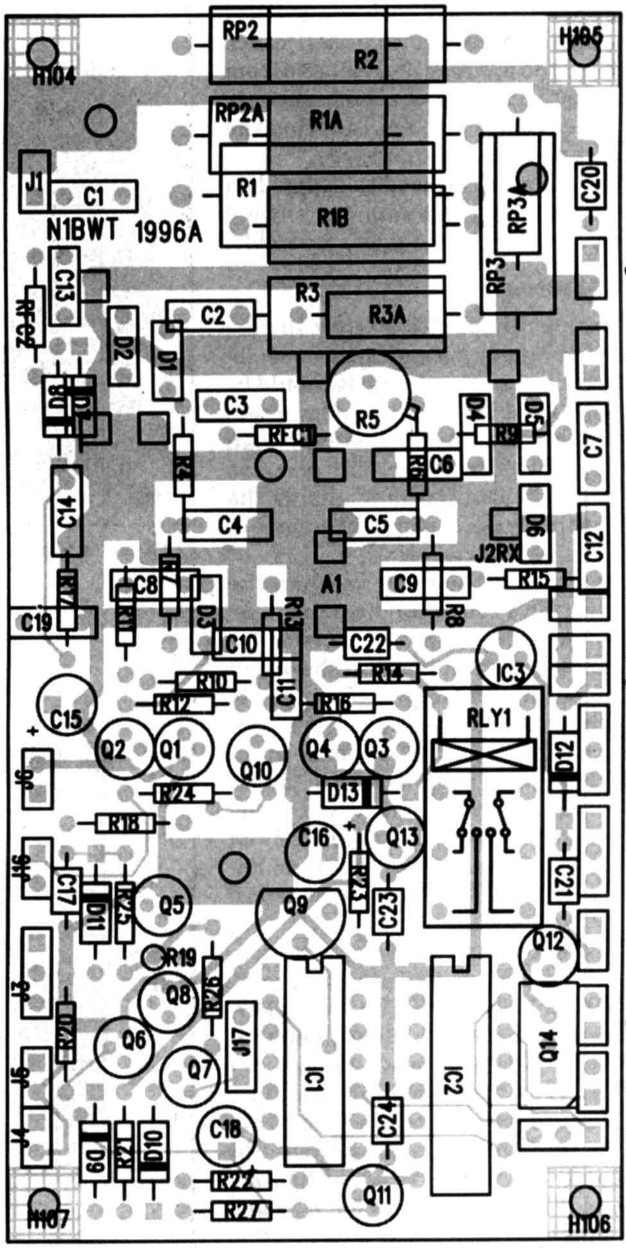


Fig 6—Parts-placement diagram for the printed-circuit board.

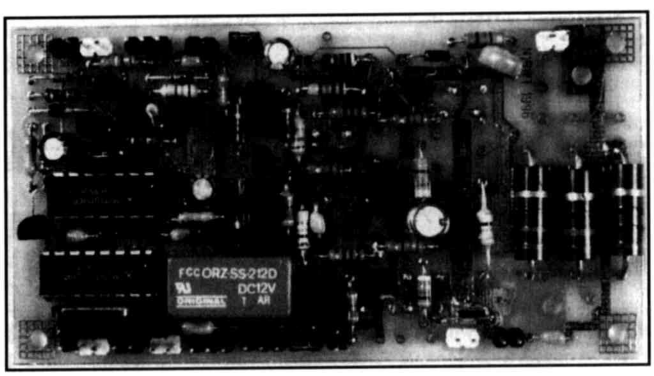


Fig 7—The top side of the PC board. The small power FETs are Motorola BS170.

The PIN-diode switch requires +6 V to operate, which may be obtained from a three-terminal regulator if not otherwise available. This regulator easily fits inside the die-cast box, as can be seen in Figs 8 and 9.

Finally, all lines entering and leaving the box should be properly filtered. I strongly recommend a bypass capacitor on the inside of the box and a ferrite bead on the wire between the capacitor and the PC board, plus a ground wire from box to board for each connection. I've seen equipment lacking these components unable to operate properly in the high-intensity RF environments found at many mountaintop sites. Listening to TV sync buzz all day is no fun!

The RF connections at J1 and J2 must have closely coupled grounds from box to board; twisted-pair or coax is preferred. The mounting standoffs do not provide an adequate ground path for RF.

Performance

I have built five of these switches and made RF measure-

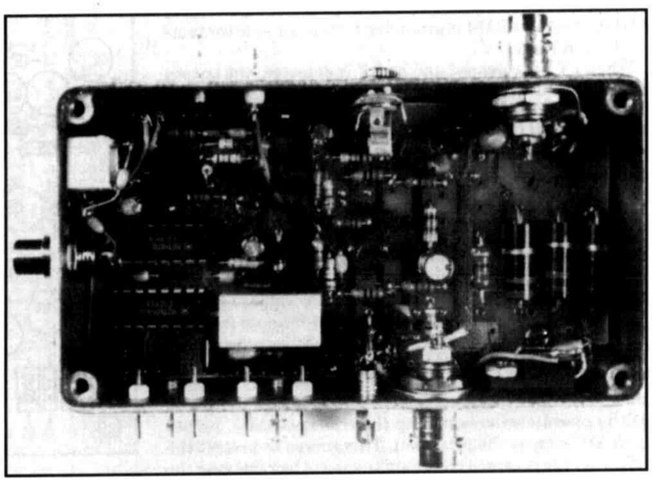


Fig 8—A completed unit, built for use with a single-mixer transverter. The small power FETs are Siliconix VN2222.

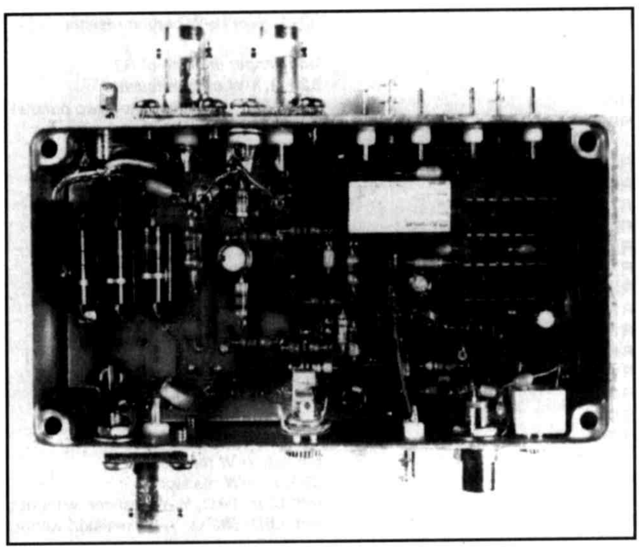


Fig 9—A completed unit, built for use with separate transmit and receive mixers. The small power FETs are Zetex BS170.

Table 1
Measured Performance

Frequency	*Receive Gain*	*Safe-Mode Gain*	*Transmit Gain*
30 MHz	5.5 dB	–60 dB	–31 dB
50 MHz	5.5 dB	–63 dB	–31 dB
144 MHz	6.0 dB	–59 dB	–30 dB
222 MHz	6.5 dB	–50 dB	–29 dB
432 MHz	7.0 dB	–35 dB	–26 dB

ments on three of them over the range of frequencies normally used for transverter IFs, with the trimpot set for 30 dB of transmit attenuation at 144 MHz. The results shown in Table 1 are typical.

Clearly, the PIN-diode switch works well at up to 222 MHz, with more than 50 dB of attenuation in the safe state and about 6 dB of gain in the receive state. The trimpot range for setting total transmit attenuation was from 20 to 38 dB at 144 MHz. The switch is still usable at 432 MHz as long as the voltage supply to the transmit amplifiers is sequenced to augment the reduced attenuation in the safe state.

The RF-detect circuit operates reliably with the output from an IC202 transceiver, roughly 2 to 3 W, switching smoothly and ignoring glitches like double-clicking the mike button by remaining in the safe state. I added attenuation between the IC202 and the switch to reduce power. The RF-detect circuit continued to operate with 15 dB of attenuation, at a power level of about 100 mW, but not with 21 dB of attenuation, or roughly 25 mW. This should be adequate margin for safe operation.

Conclusion

The IF switch described here is sequenced to provide fool-resistant operation and is flexible enough for most transverter applications. This combination should make microwave operation more reliable and successful and help protect the environment by reducing the amount of smoke being released from our equipment.

Notes

[1]Lau, Z., KH6CP, "A VHF/UHF/Microwave Transverter IF Switch," *QEX*, August 1988, pp 3-4.

[2]Wade, P., N1BWT, "Building VHF Power Attenuators," *QEX*, April 1994, pp 28-29.

[3]Down East Microwave, 954 Rte 519, Frenchtown, NJ 08825, tel: 908-996-3584, fax: 908-946-3072.

[4]Digi-Key Corporation, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel: 800-344-4539 (800-DIGI-KEY), fax: 218-681-3380. Also on the World Wide Web at **http://www.digikey.com/**.

Table 2
Parts List

A1	MAR-6 MMIC
C1,2,3,4,5,6,7,8,9,10, 11,12,14,15, 19	470 to 2000-pF chip capacitor
C13	2.2-pF disc capacitor (at 50-432 MHz)
C15	22-µF electrolytic capacitor
C16,18	10-µF electrolytic capacitor
C17,20,22,23,24	0.1-µF capacitor
C21	0.33-µF capacitor
D1,2,3,4,5,6	1SS103 PIN diode
D7,D8	1N5711, 1N5712 or HP5082-2035 hot-carrier diode
D9	1N914 or 1N4148 small-signal diode
D10,11	5.1 or 5.6-V Zener diode (1N751, 1N752, 1N5231 or 1N5232)
D12,13	1N4001 rectifier diode
IC1	CA4011 or MC14011
IC2	CA4049 or MC14049
IC3	78L08 8-V regulator
Q1,2,3,4,6,9,10, 11,12,13	BS170, VN2222 or VN10 small power switch FET (pinout varies—see text)
Q5	MPSA13 Darlington pair
Q7,8	2N3904, 2N2222, etc BJT
Q14	IRF841, IRF820, IRF830, etc N-channel power FET
T Attenuator	
R1	33-Ω, 2-W carbon resistor (two parallel 68-Ω, 1-W carbon resistors)
R2	22-Ω, 1-W carbon resistor
R3	33-Ω, $^1/_4$ or $^1/_2$-W carbon resistor
p Attenuator:	*add jumper in place of R3*
R1	120-Ω, 1-W carbon resistor
RP2	75-Ω, 2 W carbon resistor (two parallel 150-Ω, 1-W carbon resistors)
RP3	120-Ω, $^1/_4$ or $^1/_2$-W carbon resistor
R4,R6	68-Ω, $^1/_4$-W carbon resistor
R5	500-Ω small trimpot
R7,R8	10-Ω, $^1/_4$-W resistor
R9	300-Ω, $^1/_4$-W resistor
R10	5.6-kΩ, $^1/_4$-W resistor
R11,14	10-kΩ, $^1/_4$-W resistor
R12	2.7-kΩ, $^1/_4$-W resistor
R13	560-Ω, $^1/_4$-W resistor
R15	360-Ω, $^1/_4$-W resistor
R16	430-Ω, $^1/_4$-W resistor
R17	47-kΩ, $^1/_4$-W resistor
R18	8.2-kΩ, $^1/_4$-W resistor
R19	4.7-kΩ, $^1/_4$-W resistor
R20	100-kΩ, $^1/_4$-W resistor
R21,22,26	6.8-kΩ, $^1/_4$-W resistor
R23,27	33-kΩ, $^1/_4$-W resistor
R24	with LED: 1-kΩ, $^1/_4$-W resistor; without LED: 0-Ω jumper (wire)
R25	with LED: 680-Ω, $^1/_4$-W, resistor; without LED: 10-kΩ, $^1/_4$-W resistor
RFC1,2	1-µH RF choke, molded
RLY1	DPDT relay, Radio Shack #275-249
Enclosure	Bud CU-124 or Hammond 1590B die-cast box

The 1 dB Quest Revisited

By John Swiniarski, K1OR and Bruce Wood, N2LIV

(From *Proceedings Of 1996 Microwave Update)*

Introduction

In 1992 Zack Lau, W1VT (ex-KH6CP), published an article[1] describing the design and construction of a 10.368 GHz HEMT low noise amplifier. Zack's work with the NEC 32684 lead to a very reproducible LNA that paved the way for sub 1 dB noise figure (NF) preamps used in many amateurs' 10 GHz stations.

Today, due to the proliferation of 12 GHz DBS systems, the quest for 1 dB NF has become commonplace. However, with the advances in devices offering lower and lower NF, many devices have become obsolete and discontinued such as the NEC 32684. Fortunately, with manufacturers striving for high volume manufacturing capability, this has led to less expensive devices.

We have investigated the use of some of the newer devices available by several manufacturers. Our primary goal was to determine the suitability of these devices for use with Zack's NE326 design. Secondly we hoped that some of these newer devices might offer improved performance over the NE326. Our methodology was to model the performance of the devices using Zack's original NE326 circuit. If the modeling indicated reasonable performance, preamps would be built on Zack's circuit board using identical components and construction techniques. These preamps were measured for NF and gain at the 22nd Eastern VHF/UHF Conference in Vernon, Connecticut. All devices tested seemed to perform well with no tuning or optimization. We anticipate attempting to tune these preamps in the near future.

Presented here is the outcome of the modeling study, some comments on construction and the measured results of the preamps. As can be seen, the quest for 1 dB can easily be achieved.

History

After many of the amateurs in the Northeast had received NEC 32684 HEMT devices as "door prizes" at local VHF conferences, considerable interest was raised in reproducing Zack's design. With the help of Down East Microwave, a quantity of printed circuit boards was made available to members of the North East Weak Signal (NEWS) Group. Twenty-five or so preamps were constructed with the NE32684. It was then learned that NEC was replacing the '326 with the NE32584 that was shown to have a slightly lower NF. About the same time, a number of Fujitsu FHX 05 devices had become available. The FHX 05 is utilized in the Qualcomm 12 GHz receive LNAs and while it did not have as low a noise figure specification, its low cost made it an attractive candidate for study. In addition, an Avantek ATF 36077 was also made available for evaluation.

Modeling

EESOF Touchstone Software was utilized in the modeling study. Zack graciously provided a copy of his original circuit file. This eliminated the need to "reverse engineer" his circuit, an exercise that could lead to errors from making assumptions about the microstrip circuit elements. Having Zack's original model made the comparisons much more meaningful. A plot of the NF, Gain (S21) Input match (S11) and Output match (S22) of Zack's model is shown below:

The computed performance at 10.368 GHz is indicated by the markers on the chart. Using the same Touchstone circuit file, the S parameter and noise parameter file for the NE326 was replaced by the files for the NE325, FHX05 and the ATF360. Each new LNA was then "tuned" on the computer. This was done by changing the size and position of the open circuit stubs on the microstrip lines; much like we do with pieces of copper foil and a soldering iron on real hard-

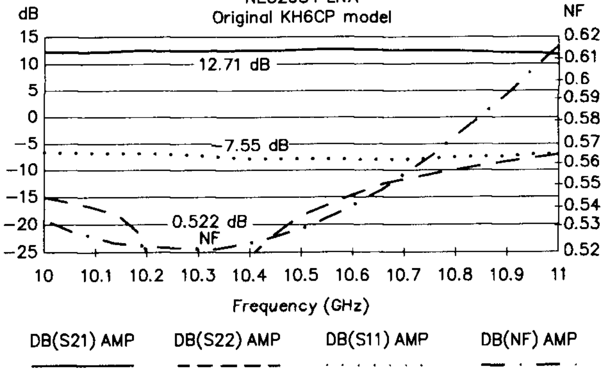


Fig 1

[1] Notes appear at the end of this section.

ware. The plots for the untuned condition and the corresponding tuned condition follows:

For visualization purposes, CAD drawings were made of the optimized LNAs. This might act as an aid when attempting optimization on the actual preamps. It can be seen that the changes are very subtle.

Construction

Samples of each of the devices were then built with identical connectors, capacitors, housings and bias circuits in an attempt to minimize variations. Assembly is fairly straight-forward, at least as far as microwave construction is concerned. Housings were fabricated using 0.500" by 0.025" hobby brass strips as described by Rus Healy, NJ2L.[2] This makes for a very compact device as the four hole SMA connectors used are also 0.500" high. The connectors used here were modified M/A-COM 2052-1215-00. This connector features a 0.085" diameter dielectric. The dielectric length was shortened to match the 0.025" brass wall thickness. This leaves a 0.025" center contact which interfaces nicely with the microstrip PCB.

Several authors have described various methods for achieving source grounding for HEMT devices. These methods include bending the device leads through the board (difficult with most new devices as they are designed for surface mounting and therefore have short leads) and using rivets. Here, we simply made two parallel cuts about 0.050" long on either side of the device and used a "u" shaped piece of ribbon. The ribbon is pushed up through one slot and down the other then soldered to the ground plane. This results in a low inductance ground pad in which to solder the source leads.

Another unique feature is the miniature active bias board mounted below the RF printed circuit board. This board uses surface mount technology (SMT) devices to realize a circuit that can fit in the small area available. It is held in place by the copper straps that also provide grounding for

LNA Results

		Computer Simulation	
Device	***Measured Result***	***Untuned Model***	***Tuned Mode***
NE326#1	0.95 dB NF @ 11.35 dB gain	N/A	0.52 dB NF @ 12.71 dB gain
NE326#2	0.87 dB NF @ 11.39 dB gain	N/A	0.52 dB NF @ 12.71 dB gain
NE325#1	0.92 dB NF @ 13.45 dB gain	0.54 dB NF @ 13.28 dB gain	0.49 dB NF @ 15.02 dB gain
NE325#2	0.85 dB NF @ 13.20 dB gain	0.54 dB NF @ 13.28 dB gain	0.49 dB NF @ 15.02 dB gain.
ATF36077	0.95 dB NF @ 11.21 dB gain	0.55 dB NF @ 13.38 dB gain	0.52 dB NF @ 14.24 dB gain
FHX05	DOA	0.78 dB NF @ 10.89 dB gain	0.78 dB NF @ 12.98 dB gain

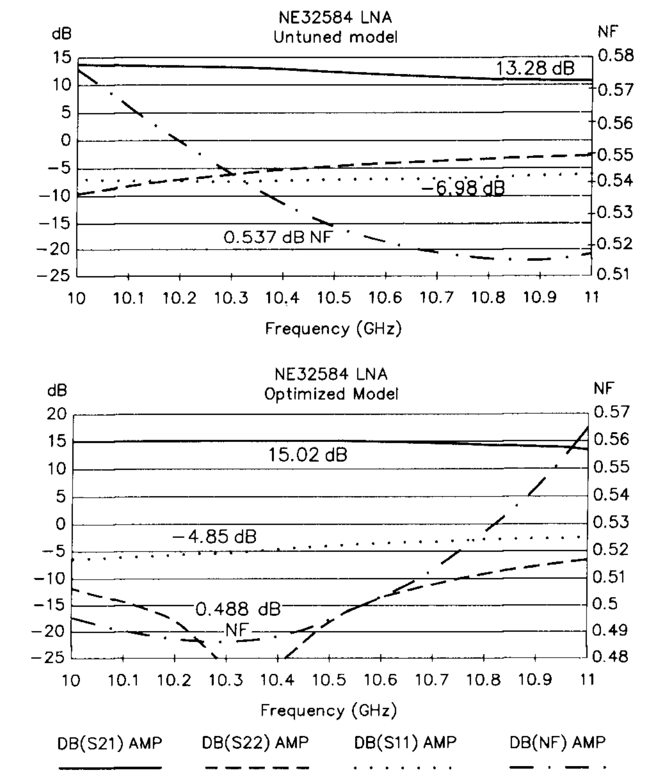


Fig 2

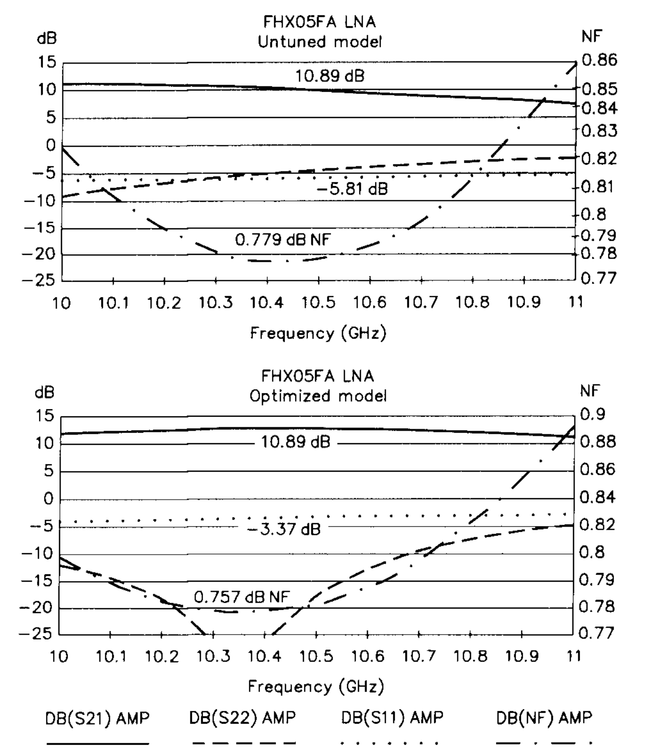


Fig 3

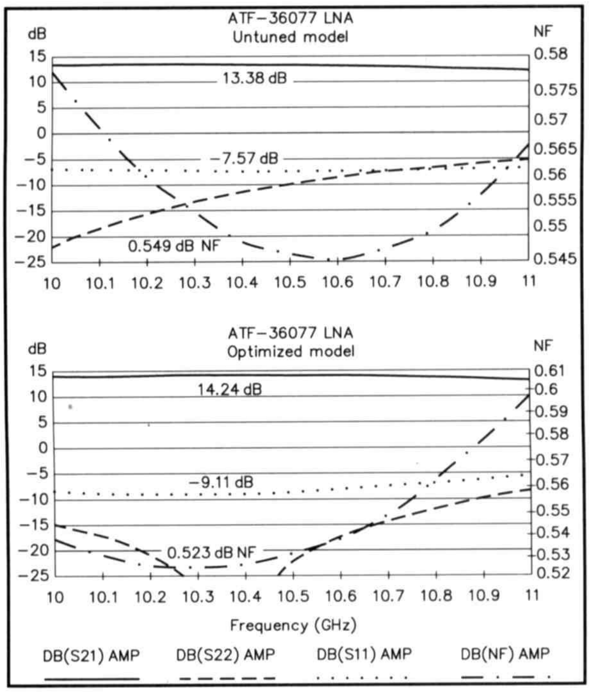


Fig 4

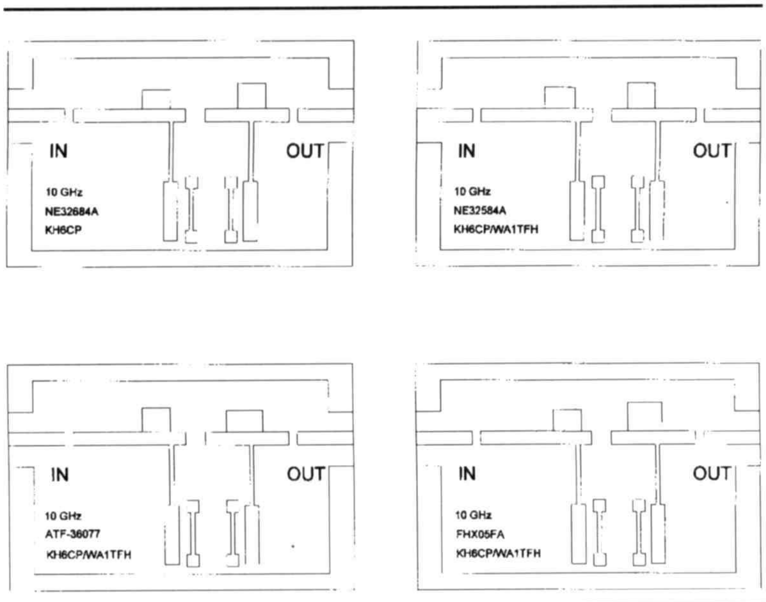


Fig 5

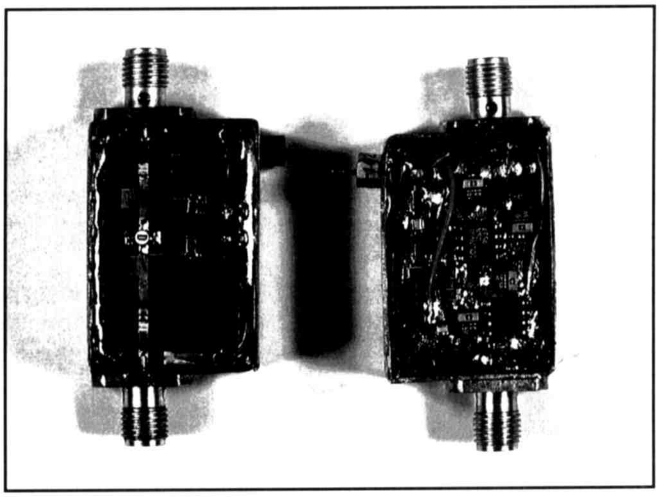
the board. The active bias circuit is similar to one described by Al Ward[3] and others. The –5 VDC supply is achieved using an ICL7660 voltage inverter chip. A spreadsheet was created to calculate the resistor values necessary for various bias conditions.

Results

The preamps were completed just in time for the 22nd Eastern VHF/UHF Conference in Vernon, Connecticut. Actually, several needed final wiring to the bias circuit, which was accomplished in the hotel room! The results obtained from the noise figure measuring session along with the computer simulation results are tabulated on the previous page. Unfortunately the Fujitsu FHX05 was DOA.

Conclusion

As the results here indicate, building a sub 1 dB NF preamp is readily achievable with several of today's hot devices. Each device tested resulted in less than 1 dB NF without optimization. The message appears to be "Use what you got." While the computer model seems to indicate much better NF performance should be possible, we are approaching the limits of what can be realized with readily available materials and construction techniques. We estimate at least 0.25 dB of loss exists ahead of the device. These are due to connector interfaces, component parasitics and dielectric losses. While it may be possible to tune these preamps to lower noise figures than measured, many builders of the original NE326 have been unsuccessful. The improvements indicated by the computer modeling of tuned vs. untuned performance hardly seem worth the trouble. Zack seems to have the magic, however. His original NE326 LNA always out performed all the others. His latest design based on the new NEC 329 device measured 0.70 dB NF at the conference. Perhaps his next quest should be for a 0.5 dB NF LNA for 10 GHz!



10 Ghz LNA Assemblies

Fig 6

References:

[1]Lau, Z., KH6CP, "The Quest for 1 dB on 10 GHz," *QEX*, December 1992, pp. 16-19

[2]Healy, R., NJ2L, "Building Enclosures for Microwave Circuits," *QEX*, June 1994, pp. 15-17

[3]Ward, A. J., WB5LUA, "Simple Low-Noise Microwave Preamplifiers," *QST*, May 1989, pp. 31-36

Noise Measurement and Generation

Quality weak-signal reception requires a low-noise system. Here's how to calculate and measure the noise performance of your system

By Paul Wade, N1BWT

(From *QEX*, November 1996)

As anyone who has listened to a receiver suspects, everything in the universe generates noise. In communications, the goal is to maximize the desired signal in relation to the undesired noise we hear. To accomplish this goal, it would be helpful to understand where noise originates, how much our own receiver adds to the noise we hear, and how to minimize it.

It's difficult to improve something unless we can measure it. Measurement of noise in receivers does not seem to be clearly understood by many amateurs, so I will attempt to explain the concepts and clarify the techniques, and to describe the standard "measure of merit" for receiver noise performance: *noise figure*. Most important, I will describe how to build your own noise generator for noise-figure measurements.

A number of equations are included, but only a few are needed to perform noise-figure measurements. The rest are included as an aid to understanding, with, I hope, enough explanatory text for everyone.

Noise

The most pervasive source of noise is thermal noise, which arises from the motion of thermally agitated free electrons in a conductor. Since everything in the universe is at some temperature above absolute zero, every conductor must generate noise.

Every resistor (and all conductors have resistance) generates an RMS noise voltage:

$$e = \sqrt{4kTRB} \qquad \text{Eq 1}$$

where R is the resistance, T is the absolute temperature in kelvins (K), B is the bandwidth in hertz, and k is Boltzmann's constant, 1.38×10^{-23} joules/K.

Converting to power, $P = e^2/R$, and adjusting for the Gaussian distribution of noise voltage, the noise power generated by the resistor, in watts, is:

$$P_n = kTB \qquad \text{Eq 2}$$

which is independent of the resistance. Thus, all resistors at the same temperature generate the same noise power. The noise is white noise, meaning that the power density does not vary with frequency, but always has a power density of kT watts/Hz. The noise power is directly proportional to absolute temperature T, since k is a constant. At the nominal ambient temperature of 290 K, we can calculate this power; converted to dBm, we get the familiar –174 dBm/Hz. Just multiply by the bandwidth in hertz to get the available noise power at ambient temperature. The choice of 290 K for ambient might seem a bit cool, since the equivalent 17° C or 62° F would be a rather cool room temperature, but 290 K makes all the calculations come out to even numbers.

The *instantaneous* noise voltage has a Gaussian distribution around the RMS value. The Gaussian distribution has no limit on the peak amplitude, so at any instant the noise voltage may have any value from $-\infty$ to $+\infty$. For design purposes, we can use a value that will not be exceeded more than 0.01% of the time. This voltage is four times the RMS value, or 12 dB higher, so our system must be able to handle peak powers 12 dB higher than the average noise power if we are to measure noise without errors.[1]

Signal-to-Noise Ratio

Now that we know the noise power in a given bandwidth, we can easily calculate how much signal is required to achieve a desired signal-to-noise ratio (S/N). For SSB, perhaps 10 dB of S/N is required for good communication; the ambient thermal noise in a 2.5-kHz bandwidth is –140 dBm, calculated as follows:

$$P_n = kTB = 1.38 \times 10^{-23} \times 290 \times 2500 = 1.0 \times 10^{-17}\ \text{W}$$

$$P_{dBm} = 10\log(P_n \times 1000) = -140\ \text{dBm}$$

(The factor of 1000 converts watts to milliwatts.) The signal power must be 10 dB greater than the noise power, so a minimum signal level of –130 dBm is required for a 10 dB S/N. This represents the noise and signal power levels at the antenna. We are then faced with the task of amplifying the signal without degrading the signal-to-noise ratio.

[1] Notes appear at the end of this section.

Noise Temperature

Any amplifier will add additional noise. The input noise N_i per unit bandwidth, kT_g, is amplified by gain G to produce an output noise of kT_gG. The additional noise added by the amplifier, kT_n is added to the input noise to produce a total noise output power N_o:

$$N_o = kT_gG + kT_n \qquad \text{Eq 3}$$

To simplify future calculations, we pretend that the amplifier is noise-free but has an additional noise-generating resistor of temperature T_e at the input, so that all sources of noise are inputs to the amplifier. Then the output noise is:

$$N_o = kG\left(T_g + T_e\right) \qquad \text{Eq 4}$$

and T_e is the *noise temperature* of the excess noise contributed by the amplifier. The noise added by an amplifier is then kGT_e, which is the fictitious noise generator at the input amplified by the amplifier gain.

Cascaded Amplifiers

If several amplifiers are cascaded, the output noise N_o of each becomes the input noise T_g to the next amplifier. We can create a large equation for the total. After removing the original input noise term, we are left with the added noise:

$$N_{added} = \left(kT_{e1}G_1G_2...G_N\right) + \left(kT_{e2}G_2...G_N\right) + ... + \left(kT_{eN}G_N\right) \qquad \text{Eq 5}$$

Substituting in the total gain $G_T = \left(G_1G_2...G_N\right)$ results in the total excess noise:

$$T_e = T_{e1} + T_{e2}/G_1 + T_{e3}/G_1G_2 + ... + T_{eN}/G_1G_2 ... G_{N-1} \qquad \text{Eq 6}$$

with the noise of each succeeding stage reduced by the gain of all preceding stages.

Clearly, if the gain of the first stage, G_1, is large, then the noise contributions of the succeeding stages are not significant. This is why we concentrate our efforts on improving the first amplifier or preamplifier.

Noise Figure

The noise figure (NF) of an amplifier is the logarithm of the ratio (so we can express it in dB) of the total noise output of an amplifier with an input T_g of 290 K to the noise output of an equivalent noise-free amplifier. A more useful definition is to calculate it from the excess temperature T_e:

$$NF = 10\log\left(1 + \frac{T_e}{T_0}\right) \text{dB at } T_0 = 290 \text{ K} \qquad \text{Eq 7}$$

If the NF is known, T_e may be calculated after converting the NF to a ratio, F:

$$T_e = (F-1)T_0 = \left[10^{(NF/10)} - 1\right]T_0 \qquad \text{Eq 8}$$

Typically, T_e is specified for very low-noise amplifiers, where the NF would be a fraction of a dB, and NF is used when it seems a more manageable number than a T_e of thousands of kelvins.

Losses

We know that any loss or attenuation in a system reduces

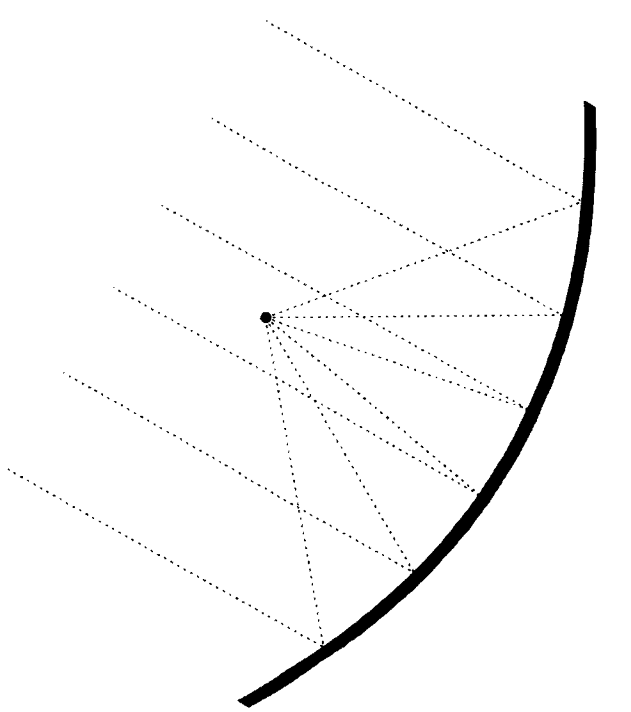


Fig 1—A parabolic dish aimed at a satellite.

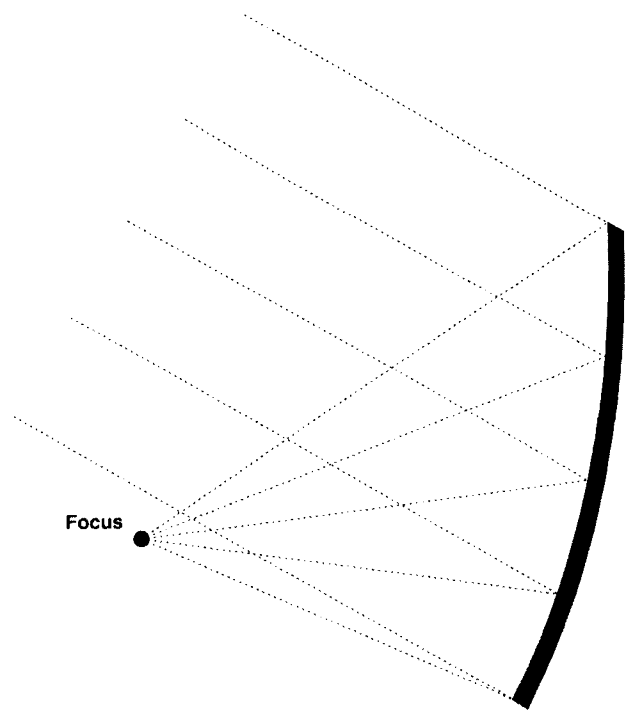


Fig 2—An offset-fed parabolic dish antenna aimed at a satellite.

the signal level. If attenuation also reduced the noise level, we could suppress thermal noise by adding attenuation.

Intuitively, this can't be true. The reason is that the attenuator—or any lossy element—has a physical temperature, T_x, that contributes noise to the system while the input noise is being attenuated. The output noise after a loss L (expressed as a ratio, not in dB) is:

$$T_{g'} = \frac{T_g}{L} + \left(\frac{L-1}{L}\right)T_x \qquad \text{Eq 9}$$

If the source temperature T_g is higher than the attenuator temperature T_x, the noise contribution is the familiar result found by simply adding the loss in dB to the NF. However, for low source temperatures the degradation can be much more dramatic. If we do a calculation for the effect of 1 dB of loss (L = 1.26) on a T_g of 25 K:

$$T_{g'} = \frac{25}{1.26} + \left(\frac{0.26}{1.26}\right) \times 290 = 80\ \text{K}$$

The resultant $T_{g'}$ is 80 K, *a 5 dB increase in noise power* (or a 5-dB degradation of signal-to-noise ratio). Since noise power = kT and k is a constant, the increase is the ratio of the two temperatures 80/25, or in dB, 10 log (80/25)=5 dB.

Antenna Temperature

How can we have a source temperature much lower than ambient? If an antenna, assumed to be lossless, is receiving signals from space, rather than the warm earth, then the background noise is much lower. The background temperature of the universe has been measured as about 3.2 K. An empirical number for a 10-GHz antenna pointing into clear sky is about 6 K, since we must always look through the attenuation and temperature of the atmosphere.[2] The figure will vary with frequency, but a good EME antenna might have a T_g of around 20 K at UHF and higher frequencies.

A couple of examples of actual antennas might bring all of this together.[3]

1. A 30-inch conventional dish at 10 GHz, with a measured gain of 36.4 dBi and efficiency of 64%. The estimated spillover efficiency is 87% for a 10-dB illumination taper. With the dish pointing at a high elevation, as shown in Fig 1, perhaps half of the spillover is illuminating Earth at 290 K, which adds an estimated 19 K to the 6 K of sky noise, for a total of 25 K. In a 500-Hz bandwidth, the noise output is −157.6 dBm.

2. An 18-inch DSS offset-fed dish at 10 GHz, with measured gain of 32.0 dB and efficiency of 63%. The spillover efficiency should be comparable, but with the offset dish pointing at a high elevation, as shown in Fig 2, far less of the spillover is illuminating warm Earth. If we estimate 20%, then 8 K is added to the 6 K of sky noise, for a total of 14 K. In a 500-Hz bandwidth, the noise output is −160 dBm.

The larger conventional dish has 2.4 dB higher noise output but 4.4 dB higher gain, so it should have 2.0 dB better signal-to-noise ratio than the smaller offset dish when both are pointing at high elevations.

However, while the offset dish is easy to feed with low loss, it is more convenient to feed the conventional dish through a cable with 1 dB of loss. Referring back to our loss example above, the noise temperature after this cable loss is 80 K. In a 500-Hz bandwidth, the noise output is now −152.6 dBm, 7.4 dB worse than the offset dish. The convenience of the cable reduces the signal-to-noise ratio by 5 dB, making the larger conventional dish 3 dB worse than the smaller offset dish. Is it any wonder that the DSS dishes sprouting on rooftops everywhere are offset-fed?

If the dishes are pointed at the horizon for terrestrial operation, the situation is much different. At least half of each antenna pattern is illuminating warm Earth, so we should expect the noise temperature to be at least half of 290 K, or about 150 K. Adding 1 dB of loss increases the noise temperature to 179 K, a 1 dB increase. At higher noise temperatures, losses do not have a dramatic effect on signal-to-noise ratio. In practice, the antenna temperature on the horizon may be even higher since the upper half of the pattern must take a much longer path through the warm atmosphere, which adds noise just like any other loss.

Image Response

Most receiving systems use at least one frequency-converting mixer that has two responses, the desired frequency and an image frequency on the other side of the local oscillator. If the image response is not filtered out, it will add additional noise to the mixer output. Since most preamps are broadband enough to have significant gain (and thus, noise output) at the image frequency, the filter must be placed between the preamp and the mixer. The total NF including image response is calculated:

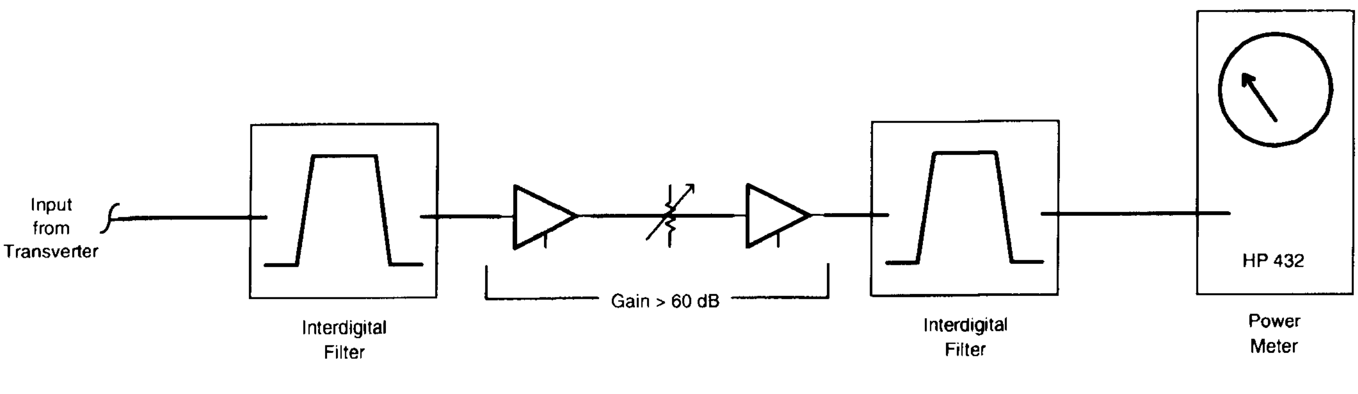


Fig 3—System for measuring sun noise.

$$NF = 10\log\left[\left(\frac{1+T_e}{T_0}\right)\left(1+\frac{G_{image}}{G_{desired}}\right)\right] \quad \text{Eq 10}$$

assuming equal noise bandwidth for desired and image responses. Without any filtering, $G_{image} = G_{desired}$ so $1+\left(G_{image}/G_{desired}\right)=2$, doubling the noise figure, which is the same as adding 3 dB. Thus, without any image rejection, the overall noise figure is at least 3 dB *regardless of the NF of the preamp*. For the image to add less than 0.1 dB to the overall NF, a quick calculation shows that the gain at the image frequency must be at least 16 dB lower than at the operating frequency.

Noise Figure Measurement

So far we have discussed the sources of noise and a figure of merit for evaluating the receiving system's response to noise. How can we measure an actual receiver?

The noise figure of a receiver is determined by measuring its output with two different noise levels, T_{hot} and T_{cold}, applied to its input. The ratio of the two output levels is referred to as the *Y-factor*. Usually, the ratio is determined from the difference in dB between the two output levels, Y_{dB}:

$$Y_{(ratio)} = 10^{\left(Y_{dB}/10\right)} \quad \text{Eq 11}$$

Then the receiver T_e may be calculated using $Y_{(ratio)}$:

$$T_e = \frac{T_{hot} - Y \times T_{cold}}{Y-1} \quad \text{Eq 12}$$

and converted to noise figure:

$$NF = 10\log\left(1+\frac{T_e}{T_0}\right) dB \quad \text{Eq 13}$$

where T_0 = 290 K

The two different noise levels may be generated separately, for instance by connecting resistors at two different temperatures. Alternatively, we could use a device that can generate a calibrated amount of noise when it is turned on. When such a device is turned off, it still generates noise from its internal resistance at T_{cold}, the ambient temperature (290 K); usually this resistance is 50 Ω, to properly terminate the transmission line that connects it to the receiver. When the noise generator is turned on, it produces excess noise equivalent to a resistor at some higher temperature at T_{hot}. The noise produced by a noise source may be specified as the Excess Noise Ratio (ENR_{dB}), the dB difference between the cold and the equivalent hot temperature, or as the equivalent temperature of the excess noise, T_{ex}, which is used in place of T_{hot} in Eq 12. If the ENR is specified, then the calculation is:

$$NF_{dB} = ENR_{dB} - 10\log\left(Y_{(ratio)} - 1\right) \quad \text{Eq 14}$$

The terms T_{ex} and ENR are used rather loosely; assume that a noise source specified in dB refers to ENR_{dB}, while a specification in "degrees" or kelvins refers to T_{ex}.

An automatic noise-figure meter, sometimes called a PANFI (*precision automatic noise-figure meter*), turns the noise source on and off at a rate of about 400 Hz and performs the above calculation electronically.[4] A wide bandwidth is required to detect enough noise to operate at this rate; a manual measurement using a narrow-band communications receiver would require the switching rate to be less than 1 Hz, with some kind of electronic integration to properly average the Gaussian noise.

Noise-figure meters seem to be fairly common surplus items. The only one in current production, the HP 8970, measures both noise figure and gain but commands a stiff price.

AIL (later AILTECH or Eaton) made several models; the model 2075 measures both NF and gain, while other models are NF only. The model 75 (a whole series whose model numbers start with 75) shows up frequently for anywhere from $7 to $400, typically $25 to $50, and performs well. Every VHFer I know has one, with most of them waiting for a noise source to be usable. Earlier tube models, like the AIL 74 and the HP 340 and 342, have problems with drift and heat, but they can also do the job.

Another alternative is to build a noise-figure meter.[5]

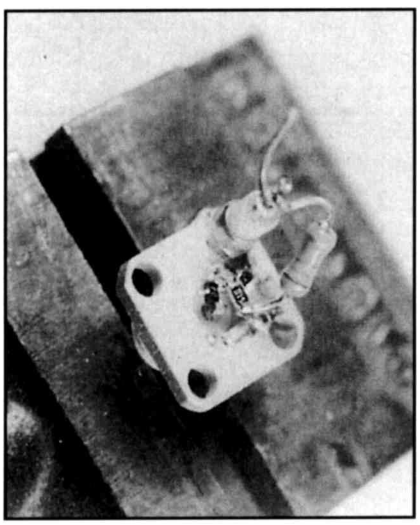


Fig 4—A noise source built on an SMA connector.

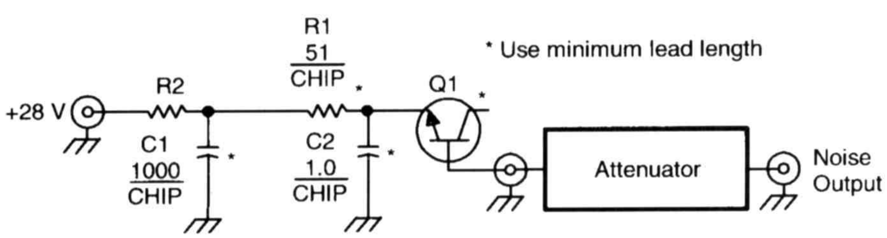


Fig 5—Schematic diagram of a noise source built on an SMA connector.

Q1—Tiny silicon NPN RF transistor such as NEC 68119.

R2—Select to set current (see text). 1 kΩ minimum, 1/4 W.

Using the Noise-Figure Meter

I'll describe the basic procedure using the model 75; others are similar, but the more complex instruments will require studying the instruction manual.

Input to almost all noise-figure meters is at 30 MHz, so a frequency converter is required (some instruments have internal frequency converters; except for the HP 8970, I'd avoid using this feature). Most ham converters with a 28-MHz IF work fine, unless the preamp being measured is so narrowband that a megahertz or two changes the NF. The input is fairly broadband, so LO leakage or any other stray signals can upset the measurement—this has been a source of frustration for many users. There are two solutions: a filter (30-MHz low-pass TVI filters are often sufficient) or a tuned amplifier at 30 MHz. Since a fair amount of gain is required in front of the noise figure meter, an amplifier is usually required anyway.

A noise source (which we will discuss in detail later) is connected to the rear of the instrument: a BNC connector marked DIODE GATE provides +28 V for a solid-state noise source, and high-voltage leads for a gas-tube noise source are also available on many versions. The noise-figure meter switches the noise source on and off. The noise output coax connector of the noise source is connected to the receiver input.

The model 75 has four function pushbuttons: OFF, ON, AUTO, and CAL. The OFF and ON positions are for manual measurements: OFF displays the detector output with the noise source turned off, and ON displays the detector output with the noise source turned on. If all is working, there should be more output in the ON position, and a step attenuator in the IF line may be used to determine the change in output, or Y-factor, to sanity-check our results. The knob marked GAIN is used to get the meter reading to a desirable part of the scale in the OFF and ON positions only; it has no effect on automatic measurements.

The AUTO position causes the instrument to turn the noise source on and off at about a 400-Hz rate and to calculate the NF from the detected change in noise. The model 75 has a large green light near the meter which indicates that the input level is high enough for proper operation—add gain until the light comes on. Then the meter should indicate a noise figure, but not a meaningful one, since we must first set the ENR_{dB} using the CAL position. The lower scale on the meter is marked for from 14.5 to 16.5 dB of ENR; adjust the CAL ADJ knob until the reading in the CAL position matches the ENR of the noise source.

If the ENR of your noise source is outside the marked range, read the section below on homebrew noise sources.

Now that we have calibrated the meter for the ENR of the noise source, we may read the noise figure directly in the AUTO position. Before we believe it, a few sanity checks are in order:

- Manually measure and calculate the Y-factor.
- Insert a known attenuator between the noise source and preamp—the NF should increase by exactly the attenuation added.
- Measure something with a known noise figure (known means measured elsewhere; a manufacturer's claim is not necessarily enough).

Finally, too much gain in the system may also cause trouble if the total noise power exceeds the level that an amplifier stage can handle without gain compression. Gain compression will be greater in the on state, so the detected Y-factor will be reduced, resulting in erroneously high indicated NF. The Gaussian distribution of the noise means that an amplifier must be able to handle 12 dB more than the average noise level without compression. One case where this is a problem is with a microwave transverter to a VHF or UHF IF followed by another converter to the 30-MHz noise-figure meter, for too much total gain. I always place a step attenuator between the transverter and the converter,

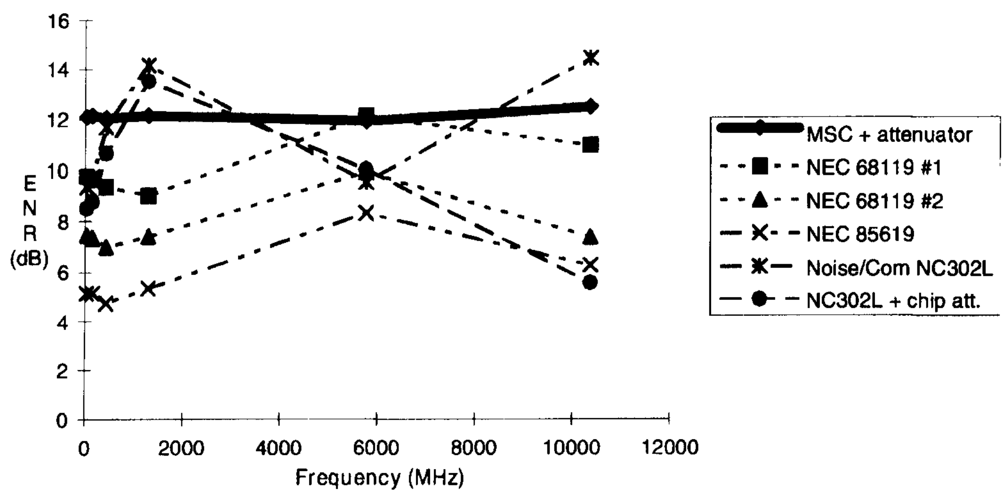


Fig 6—ENR of several versions of the noise source of Fig 5.

which I adjust until I can both add and subtract attenuation without changing the indicated noise figure.

One final precaution: noise-figure meters have a very slow time constant, as long as 10 seconds for some of the older models, to smooth out the random nature of noise. If you are using the noise-figure meter to "tweak" a receiver, *tune very slowly!*

Sky-Noise Measurement

Another way to measure noise figure at microwave frequencies is by measurement of sky noise and ground noise.[3,6] Sky noise is very low, around 6 K at 10 GHz, for instance, and ground noise is due to the ground temperature, around 290 K, so the difference is nearly 290 K. At microwave frequencies we can use a manageable antenna that is sharp enough that almost no ground noise is received, even in sidelobes, when the antenna is pointed at a high elevation. A long horn would be a good antenna choice.

The antenna is pointed alternately at clear sky overhead, away from the sun or any obstruction, and at the ground. The difference in noise output is the Y-factor; since we know both noise temperatures, the receiver noise temperature is calculated using the $Y_{(ratio)}$ and Eq 12.

The latest version of my microwave antenna program, *HDLANT21*, will make this calculation.[3] Since the measured Y-factor will be relatively small, this measurement will only be accurate for relatively low noise figures. On the other hand, they are the most difficult to measure accurately using other techniques.

A system for measuring sun noise was described by Charlie, G3WDG, that also works well for measuring noise figure from sky noise.[7] He built a 144-MHz amplifier with moderate bandwidth using MMICs and helical filters that amplifies the transverter output to drive a surplus RF power meter. The newer solid-state power meters are stable enough to detect and display small changes in noise level, and the response is slow enough to smooth out flicker. Since my 10-GHz system has an IF output at 432 MHz, duplicating Charlie's amplifier would not work. In the junk box I found some surplus broadband amplifiers and a couple of interdigital filters, which I combined to provide high gain with bandwidth of a few megahertz, arranged as shown in Fig 3. I found that roughly, 60 dB of gain after the transverter was required to get a reasonable level on the power meter, while the G3WDG system has somewhat narrower bandwidth so more gain is required.

Several precautions are necessary:

• Peak noise power must not exceed the level that any amplifier stage can handle without gain compression. Amplifiers with broadband noise output suffer gain compression at levels lower than found with signals, so be sure the amplifier compression point is at least 12 dB higher than the indicated average noise power.

• Make sure no stray signals appear within the filter passband.

• Foliage and other obstructions add thermal noise that obscures the cold sky reading.

• Low-noise amplifiers are typically very sensitive to input mismatch, so the antenna must present a low VSWR to the preamp.

A noise-figure meter could also be used as the indicator for the sky-noise measurement, but a calibrated attenuator would be needed to determine the Y-factor. Using different equipment gives us an independent check of noise figure so we may have more confidence in our measurements.

W2IMU suggested that the same technique could be used for a large dish at lower frequencies.[8] With the dish pointing at clear sky, the feed horn is pointing at the reflector, which shields it from the ground noise so it only sees the sky noise. If the feed horn is then removed and pointed at the ground, it will see the ground noise.

Noise-figure meters are convenient, but if you don't have one, the equipment for measuring sun and sky noise could also be used indoors with a noise source. The only complication is that the Y-factor could be much larger, pushing the limits of amplifier and power meter dynamic range.

Noise Sources

The simplest noise source is simply a heated resistor—if we know the temperature of the resistor, we can calculate exactly how much noise it is generating. If we then change

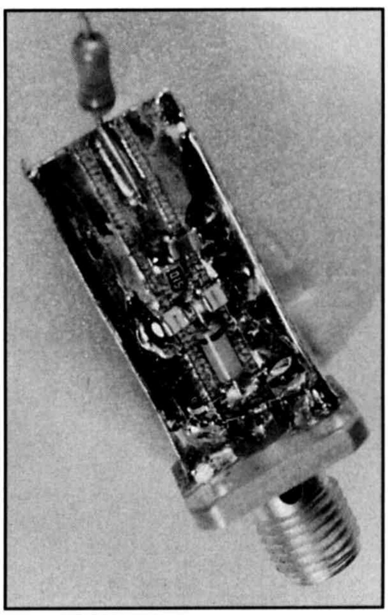


Fig 7—The homebrew noise source of Fig 8.

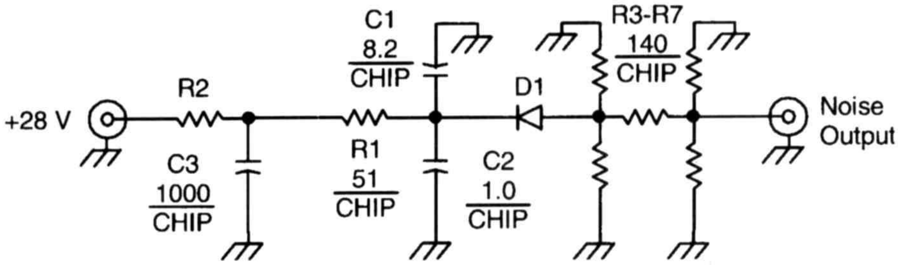


Fig 8—Schematic diagram of the homebrew noise source using a chip-resistor attenuator. C1 and C2 are microwave chip capacitors, ATC or equivalent.

the temperature, the noise output will change by a known amount. This would work if we could find a resistor with good RF properties whose value does not change with temperature, an unlikely combination. There are commercial units, called *hot-cold noise sources*, with two calibrated resistors at different temperatures and low VSWR. Typically, one resistor is cooled by liquid nitrogen to 77.3 K (the boiling point of nitrogen), while the other is heated by boiling water to 100° C, or 373.2 K. The preamp is connected to first one resistor, then the other; the difference in noise output is the Y-factor.

Since the boiling point of pure liquids is accurately known, this type of noise generator can provide very accurate measurements. However, they are inconvenient to use, since the receiver must be connected directly to alternate resistors (the loss in an RF switch would significantly reduce the noise output and accuracy). Also, few amateurs have a convenient source of liquid nitrogen.

Three types of noise sources are commonly available and convenient to use:

1. Temperature-limited vacuum tube diode. The noise output is controlled by the diode current but is only accurate up to around 300 MHz due to limitations of the vacuum tube. These units generate around 5 dB of excess noise.

2. Gas tube sources. The noise is generated by an ionized gas in the tube, similar to a fluorescent light—homebrew units have been built using small fluorescent tubes. The noise tubes use a pure gas, typically argon, to control the noise level. These units typically generate about 15 dB of excess noise.

Coaxial gas tube sources work up to around 2.5 GHz, and waveguide units to much higher frequencies. One problem using these is that a high voltage pulse is used to start the ionization (like the starter in a fluorescent light) which is coupled to the output in the coaxial units and is large enough to damage low-noise transistors. Since a noise-figure meter turns the noise source on and off continuously, pulses are generated at the same rate.

Since waveguide acts as a high-pass filter, the starting pulses are not propagated to the output, so wave-guide gas-tube noise sources are safe to use, though bulky and inconvenient. However, they could be used to calibrate a solid-state noise source.

Another problem with all gas tubes is that the VSWR of the noise source changes between the on and off states. If the source VSWR changes the noise figure of an amplifier, as is almost always the case, the accuracy of the measurement is reduced.

3. Solid-state noise sources. Reverse breakdown of a silicon diode PN junction causes an avalanche of current in the junction that would rise to destructively high levels if not limited by an external resistance. Since current is "electrons in motion," a large amount of noise is generated. If the current density of the diode is constant, the average noise output should also be constant; the instantaneous current is still random with a Gaussian distribution, so the generated noise is identical to thermal noise at a high temperature. Commercial units use special diodes designed for avalanche operation with very small capacitance for high-frequency operation, but it is possible to make a very good noise source using the emitter-base junction of a small microwave transistor.

Typical noise output from an avalanche noise diode is 25 dB or more, so the output must be reduced to a usable level, frequently 15 dB of excess noise to be compatible with gas tubes or 5 dB of excess noise for more modern equipment. If the noise level is reduced by a good RF attenuator of 10 dB or more, the source VSWR (seen by the receiver) is dominated by the attenuator, since the minimum return loss is twice the attenuation. Thus, the change in VSWR as the noise diode is turned on and off is minuscule. Commercial noise sources consist of a noise diode assembly and a selected coaxial attenuator permanently joined in a metal housing, calibrated as a single unit.

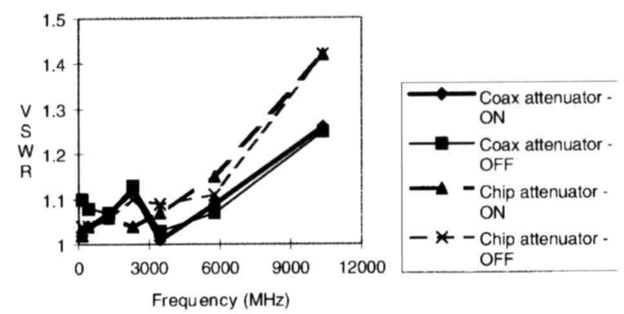


Fig 9—Measured VSWR of homebrew noise sources of Figs 4 and 7.

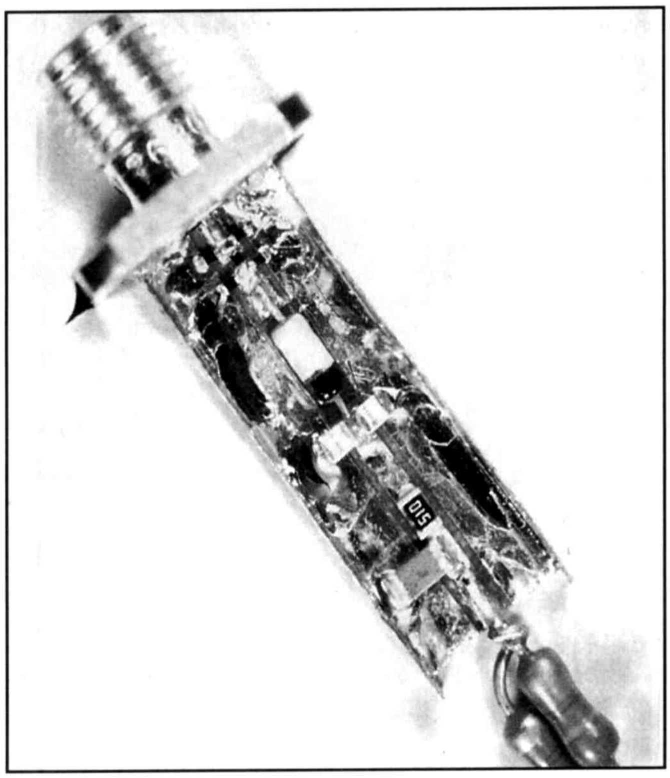


Fig 10—The noise source of Fig 8, constructed on a photographically printed circuit board.

Homebrew Noise Sources

There are three components of a noise source: a noise generator, an attenuator and the calibration data—the ENR at each frequency. The most critical component is the attenuator; it is very important that the noise source present a very low VSWR to the preamp or whatever is being measured since low-noise amplifiers are sensitive to input impedance, and even more important that the VSWR does not change significantly when the noise source is turned on and off since a change causes error in the measurement. Because an attenuator provides twice as many dB of isolation as loss (reflections pass through a second time), 10 dB or more of attenuation will reduce any change in VSWR to a very small value.

Commercial solid-state noise sources occasionally appear in surplus sources, usually at high prices but occasionally very cheap if no one knows what it is. I have found two of the latter, and one of them works! It produces about 25 dB of excess noise, which is too much to be usable. I went through my box of hamfest attenuators and found one that has excellent VSWR up to 10 GHz and 13 dB of attenuation. Mated with the noise source, the combination produces about 12 dB of excess noise—a very usable amount. Finally, I calibrated it against a calibrated noise source for all ham bands between 50 MHz and 10 GHz; not exactly NTIS traceable, but pretty good for amateur work.

While noise sources are hard to locate, noise-figure meters are frequent finds. If we could come up with some noise sources, all the VHFers who have one gathering dust could be measuring and optimizing their noise figure.

Several articles have described construction of homebrew noise sources that work well at VHF and UHF but not as well at 10 GHz.[9,10,11] All of them have the diode in a shunt configuration, with one end of the diode grounded. When I disassembled my defective commercial noise source (even the attenuator was bad), I found a bare chip diode in a series configuration—diode current flows into the output attenuator. Obviously I could not repair a chip diode, but I could try the series diode configuration. I found the smallest packaged microwave transistor available, some small chip resistors and capacitors, and soldered them directly on the gold-plated flange of an SMA connector with zero lead length, as shown in the photograph, Fig 4. We've all soldered components directly together in "dead-bug" construction; this is more like "fly-speck" construction. The schematic is shown in Fig 5, and it works at 10 GHz! I built several versions to evaluate reproducibility and measured them at several ham bands from 30 MHz to 10 GHz, with the results shown in Fig 6. All units were measured with the same 14-dB attenuator, so the diode noise generator output is 14 dB higher.

(Later I found that the MIT Radiation Laboratory had described a noise source with a series diode 50 years ago, so we aren't giving away anyone's trade secrets.)[12]

I then remembered that I had a commercial noise diode, a Noise/Com NC302L, which was used in a noise source described in *QST*, with the diode in the shunt configuration.[11] The diode is rated as working to 3 GHz, so, in the amateur tradition, I wanted to see if I could push it higher, using the series configuration. Since I didn't expect to reach 10 GHz, I increased the value of the bypass capacitor, but otherwise, it looks like the units in Fig 5. When I measured this unit, it not only worked at 10 GHz, but had more excess noise output than at lower frequencies, probably due to an unexpected resonance. The performance is shown in Fig 6 along with the other units.

Also shown in Fig 6 is the output of my pseudo-commercial noise source; even with the external attenuator, the excess noise output is pretty flat with frequency. Commercial units are typically specified at ±0.5 dB flatness. In Fig 6, none of the homebrew ones are that flat, but there is no need for it; as long as we know the excess noise output for a particular ham band, it is perfectly usable for that band.

All the above noise sources rely on a coaxial microwave attenuator to control the VSWR of the noise source. Attenuators are fairly frequent hamfest finds, but ones that themselves have good VSWR to 10 GHz are less common, and it's hard to tell how good they are without test equipment. An alternative might be to build an attenuator from small chip resistors. I used my PAD.EXE program to review possible resistor values, and found that I could make a 15.3-dB π attenuator using only 140-Ω resistors if the shunt legs were formed by two resistors in parallel, a good idea to reduce stray inductance.[13] I ordered some 0402-size (truly tiny) chip resistors from Digi-Key and more NC302L diodes from Noise/Com, and built the noise source shown in Fig 7 on a bit of Teflon PC board, cutting out the 50-Ω transmission line with a hobby knife. The schematic of the

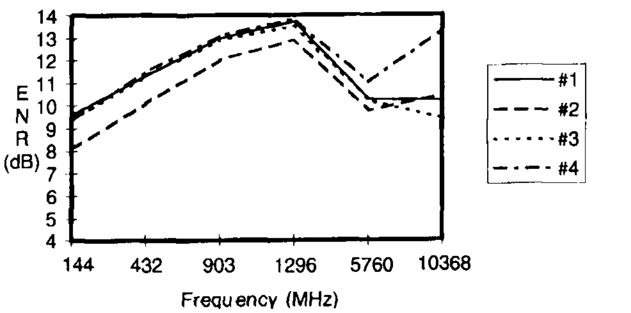


Fig 11—ENR of the PC-board noise sources (Fig 10).

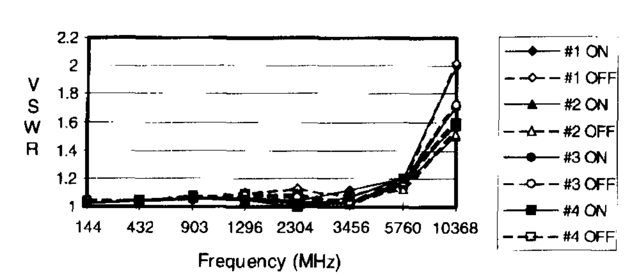


Fig 12—Measured VSWR of the PC-board noise sources (Fig 10).

Enhance Miniature Construction with Optical Feedback

Many microwave construction projects, and most modern equipment, use extremely small surface-mount components. Working with these parts requires steady hands and good vision. And as we get older, our vision usually deteriorates—I got my first bifocals last year.

I'm convinced that the key to working with tiny parts is to see them well. When I built the first noise source with the tiny 0402-size (1 × 0.5 mm) chip resistors, it was frustrating trying to get the resistors soldered where I wanted them. After that experience, I was on the lookout for a surplus stereo microscope and finally located one at a reasonable price. These microscopes are commonly used for microelectronics assembly work, providing moderate magnification at a long working distance.

After I set up the microscope on my workbench with adequate illumination, I was ready to build some more noise sources. Now the tiny chip resistors were clearly visible, and I was able to hold them in place with tweezers while soldering them exactly where I wanted them. On another project, I wanted a clearance hole in the ground plane around a hole drilled though a Teflon PC board. The hole is 0.025 inches in diameter, and I was able to cut an octagon around it with a hobby knife; the length of each side of the octagon is about the same as the hole diameter. Then I lifted the unwanted copper with the point of the knife. Magnification makes miniature work feel precise and easy instead of clumsy and frustrating.

What the microscope does is add gain to the feedback loop from the eyes to the hand. Our hands are never perfectly steady, but adding this feedback steadies them under the microscope, as the brain takes input from the eyes and automatically compensates (after a bit of practice).

A microscope is an elegant solution for very small parts, but any optical magnification helps. I have also used magnifiers, jewelers loupes and "drugstore" reading glasses. If the reading glasses are stronger than you need, they will provide additional magnification; just don't try walking around wearing them.

Other aids to miniature work are tweezers, fine-point soldering irons and lots of light. When an object is magnified, proportionally more light is required for the same apparent brightness. Tweezers help in holding small objects—I prefer the curved #7 style Swiss tweezers, of stainless steel so solder won't stick. Finally, a temperature-controlled soldering iron prevents overheating, which can destroy the solder pads on surface-mount components; 700° F tips are hot enough. All the tools I've mentioned came from hamfests, surplus places or flea markets, at reasonable prices.

So, even if you think that microwave project with tiny parts is beyond your capability, use a magnifier and give it a try. I'll bet you surprise yourself.—*N1BWT*

complete noise source is shown in Fig 8.

The chip-resistor attenuator works nearly as well as an expensive coaxial one. The measured VSWR of two noise sources, one with the chip attenuator and the other with a coaxial attenuator, is shown in Fig 9. Curves are shown in both the off and on states, showing how little the VSWR changes. The VSWR of the chip attenuator unit is 1.42 at 10 GHz, slightly over the 1.35 maximum specified for commercial noise sources, but still fine for amateur use.

Still, I wondered if I could do even better. The hand-cut board used is 0.031-inch-thick Teflon material, which is a bit thick at 10 GHz. I obtained some 0.015-inch-thick material and made a photo mask to print an accurate 50-Ω line. Then I carefully assembled the components under a microscope (see sidebar: Enhance Miniature Construction with Optical Feedback). Fig 10 shows the construction: the thin PC board is supported by a thin brass strip soldered along each side to create miniature I-beam, a much sturdier structure. The brass strips also connect the top and bottom ground areas of the board.

I built four units like the one shown in Fig 10, with encouraging results. The ENR of these units, shown in Fig 11, was higher at 10 GHz than the hand-cut one and reasonably flat with frequency—and consistent from unit to unit. The VSWR, however, was still high at 10 GHz, as shown in Fig 12. It is difficult to make a really good coax-to-microstrip transition at 10 GHz! Ordinarily, in an amplifier, we simply tune out the slight mismatch as part of the tuning procedure, but broadband tuning is much more difficult. As a final improvement, I dug up some 5-dB SMA attenuators from the swap session at Microwave Update last year. Adding one of these to the worst unit in Fig 12 reduced the VSWR to 1.18 at 10 GHz (below 1.10 at lower frequencies) and the ENR to 5.0 dB. This performance is every bit as good as a very expensive commercial noise source, lacking only NTIS-traceable calibration.

Noise-Source Alignment

The only alignment requirement for a solid-state noise source is to set the diode current; the current is always set at the highest frequency of interest. A noise figure meter must be set up with converters, etc, for the highest frequency at which the noise source might be used and set to display the detector output (OFF position on a model 75). Then voltage from a variable dc power supply is applied to the noise diode through the 1-kΩ current-limiting resistor. The detector output should increase as the voltage (diode current) increases, reach a peak, then decrease slightly. The optimum current is the one that produces peak output at the highest frequency (I set mine at 10 GHz). Then additional resistance must be added in series with the current-limiting resistor so that the peak output occurs with 28 volts applied, so that the noise source may be driven by the noise-figure meter. Once the proper resistor is determined and added, the dc end of the noise source is connected to the diode output of the noise-figure meter and the meter function is set to ON. This should produce the same detector output as the power supply.

Then the meter function is set to AUTO and the meter should produce some noise-figure indication, but not yet a calibrated one. However, it is good enough to tune up preamps—a lower noise figure is always better, even if you don't know how low it is.

Noise-Source Calibration

Much of the high price of commercial noise sources pays for the NTIS-traceable calibration. Building a noise source only solves part of the problem—now we need to calibrate it.

The basic calibration technique is to measure something with a known noise figure using the new noise source, then calculate what ENR would produce the indicated noise figure.

Fortunately, the calculation is a simple one involving only addition and subtraction; no fancy computer program required. Simply subtract the indicated noise figure, $NF_{indicated}$, from the known noise figure, NF_{actual}, and add the difference to the ENR for which the meter was calibrated, ENR_{cal}:

$$ENR_{(noise\ source)} = ENR_{cal} + (NF_{actual} - NF_{indicated}) \qquad \text{Eq 15}$$

This procedure must be repeated at each frequency of interest; at least once for each ham band should be fine for amateur use.

The known noise figure is best found by making the measurement with a calibrated noise source, then substituting the new noise source so there is little opportunity for anything to change. Next best would be a sky noise measurement on a preamp. Least accurate would be to measure a preamp at a VHF conference or other remote location, then bring it home and measure it, hoping that nothing rattled loose on the way. If you can't borrow a calibrated noise source, it would be better to take your noise source elsewhere and calibrate it. Perhaps we could measure noise sources as well as preamps at some of these events.

Using the Noise Source

Now that the ENR of the noise source has been calibrated, the noise-figure meter calibration must be adjusted to match. However, the model 75 in the CAL position has only 2 dB of adjustment range marked on the meter scale. Older instruments have no adjustment at all. However, we can just turn around the equation we used to calculate the ENR and calculate the NF instead:

$$NF_{actual} = NF_{indicated} + \left(ENR_{(noise\ source)} - ENR_{cal}\right) \qquad \text{Eq 16}$$

There is a short cut. My noise source has an ENR around 12 dB, so I set the CAL ADJ in the CAL position as if the ENR were exactly 3 dB higher, then subtract 3 dB from the reading. Even easier, the meter has a +3 dB position on the ADD TO NOISE FIGURE switch. Using that position, I can read the meter starting at 0 dB. Any ENR difference from 15 dB that matches one of the meter scales would also work—rather than an involved explanation, I'd urge you to do the noise figure calculations, then try the switch positions and see what works best for quick readout.

Reminder: Noise-figure meters have a very slow time constant, as long as 10 seconds for some of the older models, to smooth out the random nature of noise. *Tune slowly!*

Don't despair if the ENR of your noise source is much less than 15 dB. The optimum ENR is about 1.5 dB higher than the noise figure being measured.[1] The fact that today's solid-state noise sources have an ENR around 5 dB rather than the 15 dB of 20 years ago shows how much receivers have improved.

Conclusion

The value of noise-figure measurement capability is to help us all to "hear" better. A good noise source is an essential part of this capability. Accurate calibration is not necessary but helps us to know whether our receivers are as good as they could be.

Notes

[1]Pettai, R., *Noise in Receiving Systems*, Wiley, 1984.

[2]Graves, M. B., WRØI, "Computerized Radio Star Calibration Program," *Proceedings of the 27th Conference of the Central States VHF Society*, ARRL, 1993, pp 19-25.

[3]Wade, P., N1BWT, "More on Parabolic Dish Antennas," *QEX*, December 1995, pp 14-22. The *HDLANT21* program may be downloaded from **http://www.arrl.org/qexfiles**.

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More Pipe-Cap Mania

By John Sortor, KB3XG

(From *QEX*, April 1996)

I needed a band-pass filter to reduce the noise bandwidth of my 3456-MHz receiver lineup. I tried the "beg, borrow or steal" method with no success. Bill, W3HQT, gave me a copy of a 5760-MHz transverter article which used pipe-cap filters on the mixers. I scaled the dimensions for a 3456 version and it just so happens that a 1 1/2-inch pipe cap works fine.

I used a 2-inch square piece of 0.063-inch G-10 fiberglass board, two pieces of 0.085-inch semi-rigid coax with SMAs, a 1 1/2-inch 10-32 brass screw and two brass nuts. The probe spacing equals 1/4 (0.854 inch at 3456 MHz) and the probe length equals 1/8 (0.427 inch at 3456 MHz).

The semi-rigid shield should protrude 1/16 inch into the cavity side. Solder the coax on both sides of the PC board for mechanical strength. Measure the length of the probes after completing the soldering process. The heat may cause the Teflon to walk out of the copper jacket. Drill a hole in the top of the pipe cap and solder a brass nut to the cap. Use the other nut as a jam nut to assure a good electrical contact to the pipe cap.

Secure the pipe cap to the PC board using two small C-clamps. Connect your Down East no-tune transverter or equivalent to the filter. Move the pipe cap around to minimize the filter loss. Simultaneously re-tweak the tuning screw to find the optimum spot for the shell and screw. Tighten the C-clamps and solder the pipe cap to the PC board.

I can't believe this thing works as well as it does. I got about 1 dB of loss at 3456 MHz with a 3-dB bandwidth of about 50 MHz. The filter dropped off rapidly below 3.4 GHz, but flat topped above 3.4 GHz. There was another peak at 7 GHz where the probes are 1/4 waves. This works fine for us since we are trying to reject our LO signal. WA3NUF says that shortening the probes will increase the loss but sharpen the filter skirts. Feel free to experiment at your own expense. The pipe caps are two bucks a piece, so you don't have to feel bad about scrapping a cap or two.

Frequency	***Loss***
GHz	*dB*
2.0	–54
2.5	–43
3.0	–31
3.42	–4.4
3.45	–1.1
3.48	–4.0
4.0	–20
4.5	–22
5.0	–27

Temperature Compensation or Crystals

By Dave Mascaro, WA3JUF
(From *QEX*, April 1996)

Since my "Hints & Bits" article in June 1994 *Cheesebits*, several homebrewers have asked questions about ovens and thermistors. Which unit is easiest to use? How are they wired into an existing unit? Does something as simple as a thermistor really work?

Why is temperature compensation a concern? The frequency of your transverter's LO can drift many kilohertz when you go roving or during seasonal temperature changes in your basement. Few amateurs think about the temperature drift of their transverters or commercially made transceivers when they take the equipment out to the field. I measured a 75-kHz shift on the nonovenized LO of a 10-GHz transverter. Reducing (if not eliminating) the frequency variable when working a weak station is very significant.

The easiest, cheapest and fastest way to temperature compensate a crystal is to use a PTC thermistor. The leaded thermistor KC004P is $2.31 from Digi-Key (800-344-4539) and can be added to any crystal circuit in 10 minutes. The unit is connected directly to 12 V dc. The Yaesu G9090019 thermistor comes with its own holder that slides over an HC-25/U crystal. The number for Yaesu USA is 310-404-2700.

First, quickly unsolder one lead of the thermistor without damaging the metalization. Solder tin the side of the crystal. Solder the flat side of the thermistor to the crystal case. Solder the remaining lead to the +12-V line. Solder a small gauge wire from the case of the crystal to dc ground to complete the dc path. The thermistor is nominally 50 Ω at 25° C, so it draws several hundred milliamps for a short while, then settles down to an idling current of less than 30 mA when it reaches its operating temperature.

I found the 60° C unit to be hot enough, even for the temperature excursions my equipment sees in the attic. Even in winter the crystal temperature stabilizes after about five minutes, instead of drifting for hours. Adding a small styrofoam or insulated cover over the crystal will further stabilize the frequency.

Don't get hung up on netting crystals to an exact frequency. Adding a heater will age the crystal in addition to moving its frequency, so allow the transverter to stay on for several days to complete the aging process. You may not be able to pull it back to the original frequency with the crystal trimmer. You can either make a note of the exact frequency and use it that way, or figure out how far the heated crystal moved and order another crystal based on that frequency delta and heater temperature.

I have added thermistor compensation to several SSB Electronic transverters, which all adjusted back to the original frequency. The thermistors work great on reference crystal oscillators for PLL sources. Adding the $2.31 thermistor to a DEM no-tune transverter produces a rock-solid LO that won't budge even in the hot sun.

Mixers, Etc, for 5760 MHz

A modular building block transverter

5760 MHz may be the least utilized amateur microwave band. Few construction articles have been published and very little equipment is available. Yet there is a wealth of surplus components available which are usable on this band as building blocks for a transceiver. The missing pieces, readily constructed by the microwave experimenter, are described and integrated into a system.

By Paul Wade, N1BWT

(From *Proceedings of 1992 Microwave Update*)

Introduction

5760 MHz may be the least utilized amateur microwave band. Few construction articles have been published and very little equipment is available. Yet there is a wealth of surplus components available which are usable on this band as building blocks for a transverter. The missing pieces are readily constructed by the microwave experimenter, so that more stations could get on the air. In the week after my transceiver became operational, I contacted five other stations; For four of us, it was our first contact on this band.

The wavelength for this band is 6 cm, so a dish antenna will have only 3 dB less gain than on 10 GHz—large gains are available from modest size antennas. On the other hand, amplifiers capable of providing several watts of power are much more readily available for 5760 than for 10 GHz. Combining them gives a modest station a really impressive ERP, so DX possibilities are good.

Mixer

The heart of any transverter system is the mixer, and there are few choices available for 5760 MHz. Unlike other components, surplus mixers for this band are scarce, so homebrewing is necessary. One option, the KK7B no-tune transverter[1] for this band has a simple bilateral mixer, used for both receive and transmit, so that switching is needed to utilize separate power amplifiers and receive preamps. Having separate mixers for transmit and receive is preferable so that each path may be optimized for its function.

[1] Notes appear at the end of this section.

The KK7B transverter has a 1296 MHz IF, probably because of the difficulty of making sharp filters, or any other high-Q circuit, reproducibly on a printed-circuit board at this frequency. The dimensions are too small and critical for normal printed-circuit tolerances.

Another transverter, with separate transmit and receive mixers, was described[2] (in German) by DJ6EP and DCØDA, and subsequently reprinted[3] in *Feedpoint* and *73*. The latter also described a modification to use a surplus Phase-Locked Microwave Source as the local oscillator, and made PC boards available, making it even more attractive. A unit was assembled, with abysmal results. There was no apparent mixing taking place and the only output was strong LO leakage. Closer examination of the mixer circuit suggested that it might be a harmonic mixer, operating with a half-frequency LO. This was confirmed when we located someone who could fake enough German to translate the article. At 5.6 GHz, the LO input impedance is effectively a short circuit, and measures exactly that.

It was obviously time for a new design. Some time ago, I designed and built a series of balanced mixers[4,5] using 90° hybrid couplers[6] from 1296 to 5760 MHz. Since these had worked well as receivers, two mixers were integrated with a third 90° hybrid coupler as a power splitter on a small Teflon PC board. The layout is shown in Figure 1. As expected, it worked well as a receive mixer, with about 7 dB of conversion loss. However, it worked poorly as a transmit mixer, with transmit conversion loss of around 25 dB.

This nonreciprocal performance was a mystery until

Rick, KK7B, steered me to an article explaining why a 90° hybrid-coupler works as a downconverter but not as an upconverter. I had only worked out the downconverter case and assumed that it would be reciprocal.

One reason for choosing the 90° hybrid coupler is that it is a low-Q structure and uses wide, low impedance transmission lines, so that dimensions are not extremely critical and performance should be reproducible.

The KK7B mixer[1] used a 6/41 "rat-race" coupler,[6] so the next version, shown in the photograph of Figure 2 used this structure for the transmit mixer. Line widths are somewhat narrower than the 90° hybrid coupler, but it is still a low-Q structure, so it should still be reproducible. This unit had much better transmit performance, about 8 dB of conversion loss, but its noise figure was not quite as good as the original receive mixer, so the original receive mixer was retained.

The final version integrates the "pipe-cap" filters like those in the DJ6EP[2] transverter onto the mixer board. These are copper plumbing pipe caps for three-quarter inch copper tubing, with probes 7/32 inch long and tuned with an 8-32 screw; dimensions are from the measurements made on individual filters. PC board layout is shown in Figure 3, and the only other components on the board are the mixer diode pairs and a 51 ohm chip resistor termination. IF attenuators like those in some of the no-tune transverters would also fit, and are recommended for the transmit side. No through holes are needed for grounding—the radial transmission line stub acts as a broadband RF short. The diodes I used (Hewlett-Packard HSMS-8202) are inexpensive Ku-band mixer diode pairs; they and the mixer boards are available from Down-East Microwave.

Local Oscillator

Microwave local oscillators normally start with a crystal in the 100 MHz range, followed by a string of multipliers. For 5760 MHz, a multiplication factor of 50 to 60 is necessary—not an easy task. Fortunately, there are many surplus Phase-Locked Microwave Sources (often called PLO bricks) available, made by companies such as Frequency West and California Microwave. These units were used in the 5.9-6.4 GHz communications band, and provide more than enough LO power for the mixer (a 6 dB attenuator was needed with mine). Some units have an internal crystal oven; after a few minutes warm-up, stability is comparable to a VHF transceiver. Operation and tune-up of these units has been described by KØKE,[8] WD4MBK,[9] and AA5C.[10] The sources can be used unmodified to provide high-side LO injection, above 5760 MHz, or modified[11] to operate below 5760 for normal low-side injection. Unless you are obsessive about direct digital readout, high-side injection, using LSB and reverse tuning, is perfectly acceptable. For CW operation, there is no difference.

Most of the available sources operate on –20 volts. This is only a problem for portable operation. WB6IGP has described[12] a +12 volt to –24 volt converter, and surplus potted converters are occasionally found. A three-terminal regulator IC provides the –20 volts. In order to prevent switching noise generated by the converter from reaching the LO, this is all contained in a metal box with RFI filtering on both input and output.

Filter

A good filter is essential for a serious microwave station, particularly for mountaintop operation. Most accessible high places are crowded with RF and microwave sources, so the RF environment is severe. I've seen unfiltered no-tune transverters for other bands fold up and quit in mountaintop environments.

The best filter I've tried is a waveguide post filter, such as the 10 GHz ones described by N6GN.[13] It is easily built using only a drill, tuning is smooth and non-critical, and the performance is excellent. Glenn was kind enough to calculate dimensions for 5760 using standard waveguide and hobby brass tubing, as shown in Figure 4. Dimensions are for WR-137 waveguide for 5800 MHz; reducing the spacings a hair will give a little more tuning range. I built two units; the second, with careful fit and flux cleaning had 0.4 dB of loss, while the first, with sloppy fit and soldering, had 0.5 dB of loss. Both units measured as shown in Figure 6, with steep skirts (135 MHz wide at 30 dB down) and no spurious responses detectable (>70 dB down). Tuning was smooth and easy; with high-side LO injection, the LO and image frequencies are out of the tuning range, so 5760 is the only output that can be found while tuning.

Construction Hint:

Make sure the holes are carefully measured and centered in the waveguide. Centerpunch lightly. Using a drill press, start the holes using a center drill, then drill them out a few drill sizes undersize. Then enlarge them one drill size at a time until the tubing is a snug fit. Solder on a hotplate using paste rosin flux.

Power amplifier

Semiconductor devices capable of good power and gain at this frequency are not as readily available as complete surplus amplifiers. Both TWT (Traveling Wave Tube) and solid-state amplifiers in the one to ten watt power range are fairly common. TWT operation was recently described by KD5RO.[14] These units usually have 30+ dB of gain and typically require one milliwatt (0 dBm) drive power.

Since our mixer has a maximum linear output of around –7 dBm, a small intermediate amplifier is needed. A GaAsFET stage could be used, but a broadband MMIC amplifier is also usable — some of the common MMICs have usable gain left at this frequency. WØPW[15] has described several amplifiers using MAR-8 devices. After seeing several unpublished papers about cures for oscillating MAR8s in 3456 MHz no-tune transverters, I chose to use the MSA-0986, which is unconditionally stable, at the cost of reduced low frequency gain. It has a flat gain of approximately 7 dB from about 0.1 to 4 GHz, rolling off to zero at about 8 GHz. At 5760, the two-stage amplifier shown in Figure 6 has 8.3 dB of gain; the 1/8 inch disc on the input line is adjusted to peak the output at 5760, yielding an additional dB, then soldered in place. Gain is still broadband, about 14 dB over the 903 to 3456 MHz bands, falling to zero at 8 GHz.

Construction is with minimal lead length on a scrap of Teflon PC board, with a soldered sheet brass enclosure. Keep the enclosure dimensions smaller than WR-137 waveguide cross-section and it won't oscillate. The blocking capacitors marked with an asterisk should be high quality chip capacitors.

Preamplifier

The WB5LUA GaAsFET preamps[16] are completely satisfactory—they work. My attempt to design a single stage amplifier was noteworthy only because it oscillates within 50 MHz of 5760.

Switching

KH6CP/1 has described two[17,18] transverter switching units that are a good starting point. Since most surplus coax relays operate on 28 volts, I tap off some negative voltage from the PLO supply through a 7915 three-terminal regulator IC to provide –15 volts to one side of the relay. The other side is connected to the +12 volt input to activate it with a total of ≈28 volts.

Surplus

There is a wealth of surplus components available that were originally used in the 5.9-6.4 GHz communications band. Most of these are broad enough to cover 5760 as well. As many commercial systems are upgraded, some of it appears dirt-cheap at hamfests—I've seen scrap metal dealers with truck-loads.

Particularly useful components are circulators and isolators,[9] which allow RF to flow in only one direction, protecting amplifiers from load mismatch (and often prevent oscillation).

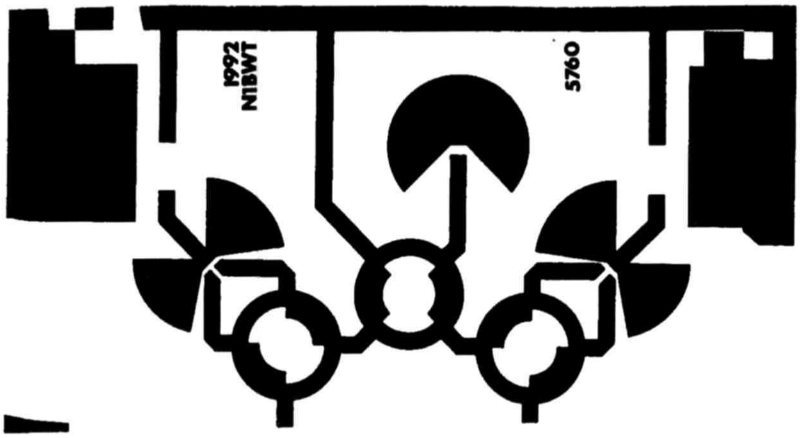

Fig 1—First dual mixer layout.

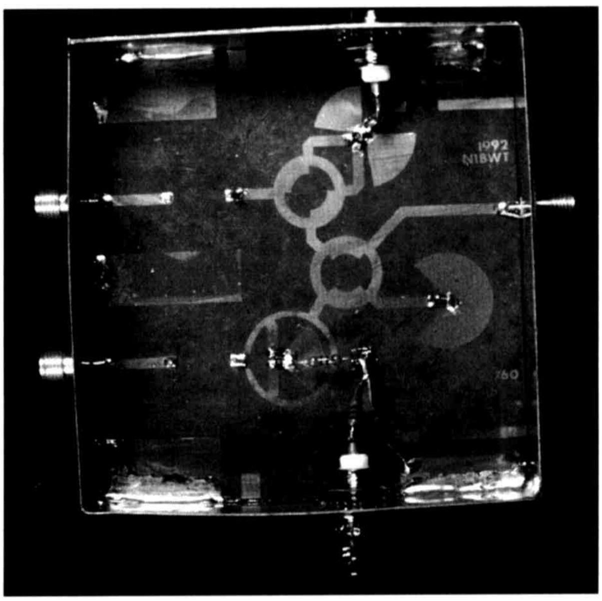

Fig 2—Revised dual mixer.

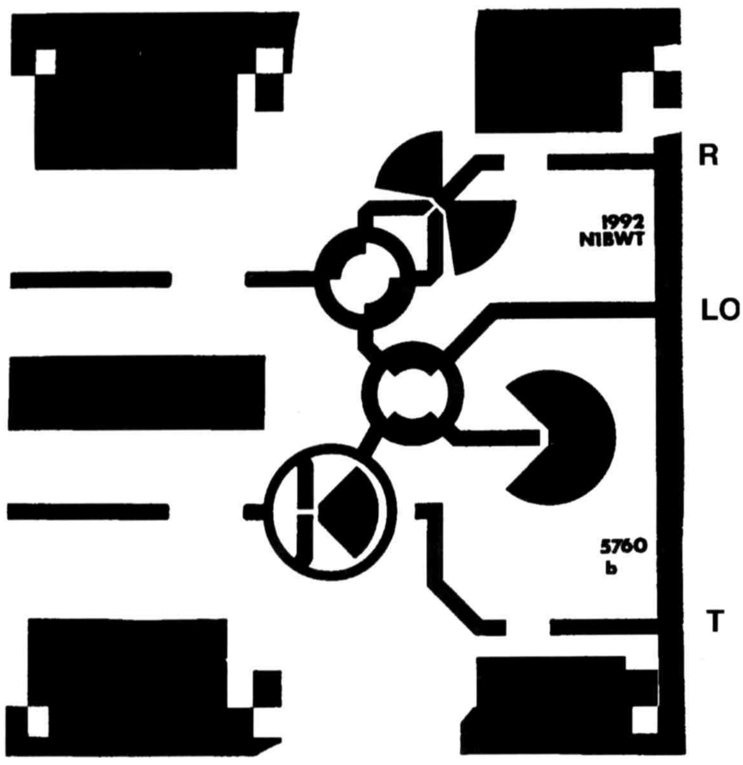

Fig 3—Layout of dual mixer with filters.

I use isolators between the LO and mixer, between the preamp and filter, and, of course, between the power amp and antenna. Other useful components are waveguide sections, useful for making filters, horn antennas, directional couplers, attenuators, and diode detectors, both coax and waveguide. Some filters can be retuned to 5760; I found a waveguide filter for 6.2 GHz that just barely reached 5760 at the tuning limit, but it had more loss than the homemade ones. Slotted lines are great for measuring VSWR and impedance. Even the cabinet for my transverter originally held a now-defunct TWT amplifier.

Block Diagram

Fig 7 is a block diagram of the current state of my transverter. Of course, with the building block approach, any of the blocks can be changed to improve the system.

Weak Signal Source

On a lightly populated band, having a weak signal source available greatly improves your confidence. I have been using an 1152 MHz signal source[20] based on the no-tune LO scheme, with a waveguide diode detector acting as a multiplier to 5760 MHz (or 10368 MHz). Recently I put together a much simpler version, based on an 80 MHz crystal oscillator of the type used in computers. Multiplying by 72 equals 5760 MHz, with no other harmonics near microwave calling frequencies. As shown in Fig 8, the oscillator drives an MMIC. K1TR gave me some MWA320 MMICs, which run fine on 5 volts like the oscillator, and I threw together a prototype on a piece of perfboard.

The oscillator output is a 5 volt square wave, so it is seriously overdriving the MMIC. As you would expect, the output waveform looks terrible, probably full of harmonics! Connecting the output to a piece of waveguide (which acts as a high-pass filter) gives an S9 signal across the basement at 5760.150 MHz.

Higher frequency MMICs, like the MAR or MSA series, and better microwave construction would probably work even better.

Conclusion

The transverter system I have described started out as a mixer, then other modular building blocks were added to make a system. Many of the blocks are from surplus sources, others homebrewed as needed. In its current state, it's a good portable system for mountaintopping, but not capable of moonbounce. However, the building block approach can be used to assemble a system at any level of performance without limiting future enhancements.

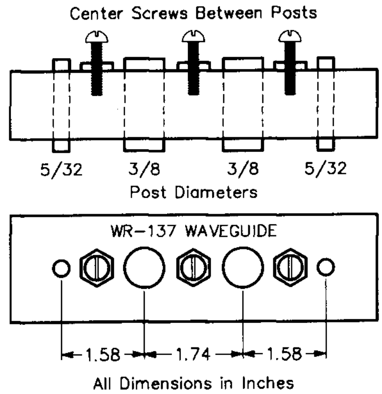

Fig 4—Waveguide filter for 5760 MHz.

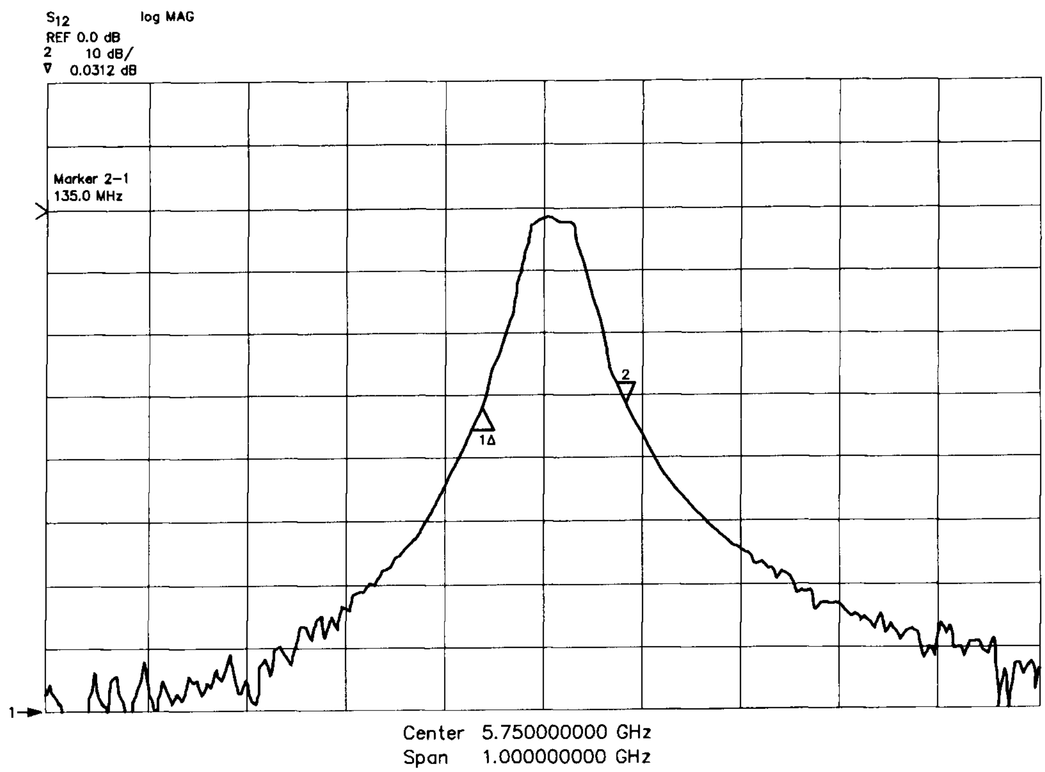

Fig 5—Measured performance of waveguide filter.

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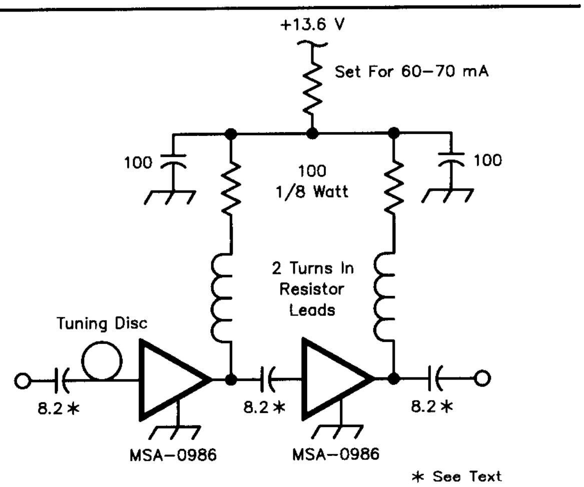


Fig 6—MMIC amplifier.

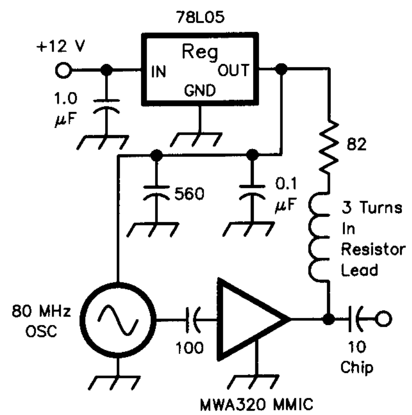


Fig 8—5760 MHz weak signal source.

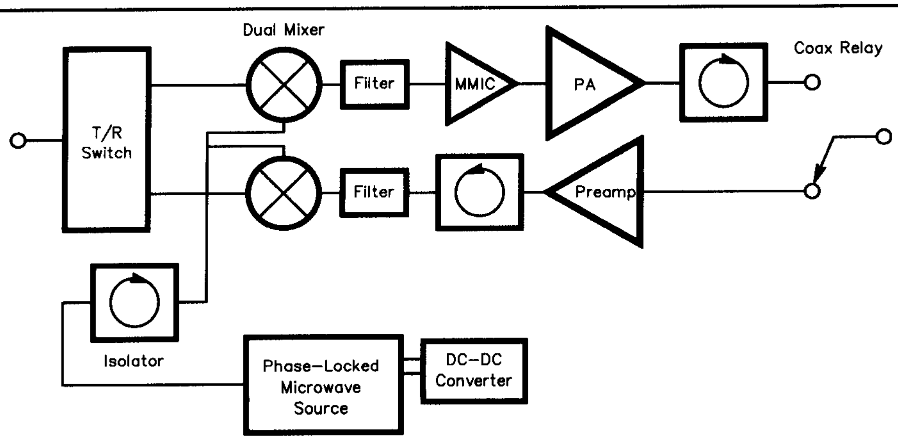


Fig 7—5760 MHz transverter block diagram.

Building Blocks for a 10 GHz Transverter

By Paul Wade, N1BWT

(From *Proceedings of The 19th Eastern VHF/UHF Conference*)

Introduction

This is not a description of a complete no-tune transverter for 10 GHz SSB and CW, but rather a series of building blocks that comprise the critical parts of a transverter. These will ease the development of a working narrowband system, but some skill and experience is required at 10 GHz. There are still enough vagaries at this frequency that a single-board transverter would probably not work as assembled—building blocks can be individually checked before assembly into a system.

One obstacle to the development of reproducible 10 GHz equipment is the lack of affordable devices to provide gain, so the first building block is a two-stage amplifier to provide adequate gain in the several places required to make a working system. Other blocks include a dual mixer, with separate mixers for transmit and receive, an LO multiplier to generate a 10 GHz local oscillator, and a power amplifier for modest power output, plus some blocks that others have previously described that don't need reinventing. All these blocks can be put together as a transverter, such as the block diagram in Figure 13 below.

Background

My previous 10 GHz narrowband system used a surplus waveguide balanced mixer, which performs reasonably well with only an LO, filter, and antenna as a low-power system. However, the low power limits range, and we have not made any contacts beyond the range of a good wideband FM Gunnplexer system. The next step, integrating the waveguide mixer into a high-performance system with transmit and receive amplifiers would require multiple switches and waveguide to coax transitions, so packaging it for portable operation would be unwieldy. A new system, using printed microstrip circuitry, was in order. At Microwave Update '92 in Rochester,[1] I presented a description of my modular building-block approach to assembling a transverter for 5760 MHz.

Since the 5760 transverter—designed with inexpensive, available parts—works well, I wondered whether a 10 GHz system could be made with the same approach.

As this project neared completion, another 10 GHz transverter, by KH6CP, appeared in *QST*. Zack's design[2] is also constructed as a series of building blocks, and someone contemplating building a 10 GHz system would do well to study both this version and Zack's, and perhaps choose the block that best fit available components and talents.

Two-Stage Amplifier

One obstacle to the development of reproducible 10 GHz equipment is the lack of affordable devices to provide gain. The no-tune transverters for the lower microwave bands use low-cost silicon MMICs for the gain elements, but these don't work at 10 GHz and gallium arsenide MMICs are not cheap. WB5LUA has described excellent low-noise preamps,[3] which are also useful as low-power amplifiers, but the GaAsFETs that Al used are rather pricey. Inexpensive GaAsFETs are available, so a reproducible design could be used as a gain block like an MMIC.

Amplifier Design

The design goal for this amplifier was simple: to utilize the cheapest readily available device that provides good performance at 10 GHz. The best I found is the Avantek ATF-13484 at $4 from Down East Microwave. A bit of computer calculation with S-parameters yielded a straightforward design that could be printed on common 1/32-inch thick Teflon PC board. The circuit has a transmission line and a stub at the output, plus bias decoupling networks. A layout with two identical stages is shown in Figure 1.

A proper bias network is one key to disaster-free operation. I prefer the active bias circuit described by WB5LUA[3] for his low-noise GaAsFET preamps. The schematic diagram in Figure 8 shows the two-stage amplifier with active bias circuits. The component values shown result in an operating point of about 3.5 volts at 35 mA for each stage.

[1] Notes appear at the end of this section.

Amplifier Performance

Measured gain, shown in the curves of Figure 2, is about 15 dB at 10368 MHz. This performance was achieved after minor trimming of the stubs nearest to the connectors, probably to compensate for the connector mismatch. The one-dB compression point was around +13 dBm, a modest amount of power at 10 GHz. Although the amplifier was designed to work with the ATF-13484, other available small GaAsFETs also work fine. Units were assembled using the Mitsubishi MGF-1302 and MGF-1412, which both provided more gain than the original with no tuning, about 21 and 19 dB respectively. Curves for these amplifiers are also shown in Figure 2.

Noise figure measured around 5 dB at operating current, and would probably be lower at lower current with a bit of retuning. This is hardly a super front end, but it is respectable for the cost, and much better than the bare mixers many of us have been running.

Figure 2 also shows the predicted ATF-13484 performance for lossless components. Measured gain for the ATF-13484 unit is a bit lower, as might be expected with real components at this frequency.

Amplifier Construction

Construction is with minimal lead length on Teflon PC board, with a soldered sheet brass enclosure. Components marked with an asterisk are chip capacitors or resistors, and the blocking capacitors should be high quality chip capacitors. Parts placement can be seen in the photograph, Figure 7.

Achieving minimum lead length takes a bit of effort, but is worth it. I do it by bending the ground leads down right at the package, as illustrated in the sketch by DB6NT[4] shown in Figure 9. Then I drill a hole in the PC board, which just fits the package body, and file the sides just enough for the ground leads to slip through. Then I push the device down until the input and output leads start to bend up, so the ground leads are as short as possible. The ground leads are then bent over and soldered to the ground side of the board right at the edge of the hole.

Stability

The key to stability in these amplifiers is minimum ground lead inductance, achieved by keeping the leads as short as possible. In one of the units, I assembled MGF-1412 devices in smaller holes drilled for MGF-1302 (I dropped my last 1302 on the floor and couldn't find it). This resulted in slightly longer ground leads, perhaps 0.5 mm, and an extremely unstable amplifier, which couldn't be tamed.

Computer simulation with additional ground inductance confirmed this behavior, which was only cured by getting more MGF-1302s and installing them. Computer analysis with minimum ground lead inductance predicts sta-

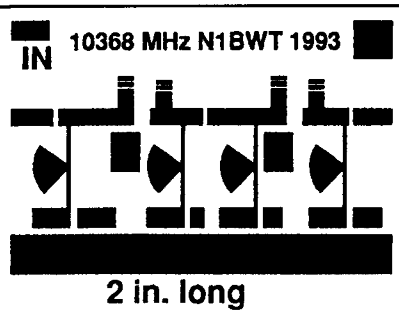


Fig 1

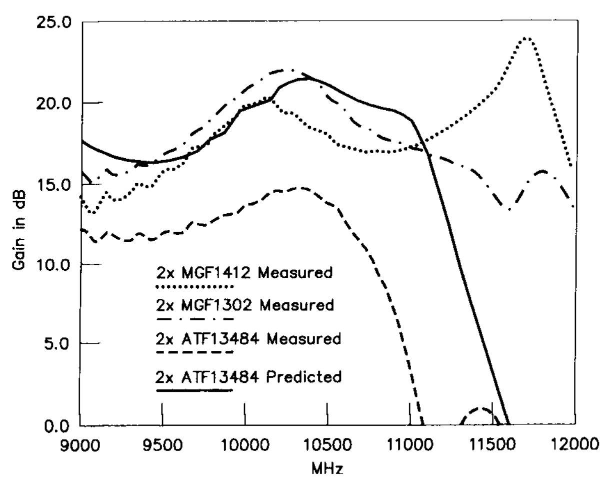


Fig 2

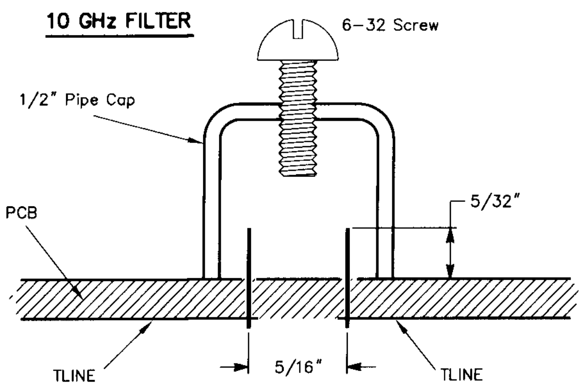


Fig 3

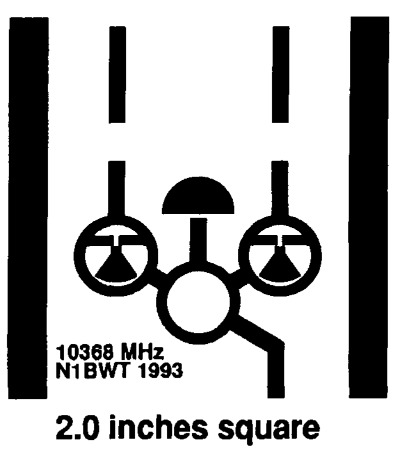


Fig 4

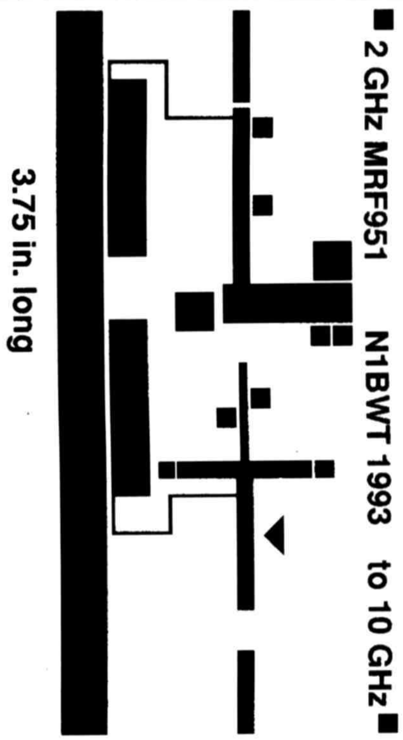


Fig 5–PC layout for 10 GHz LO multiplier.

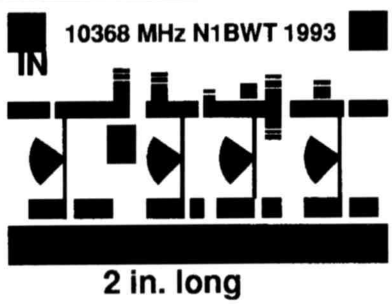


Fig 6–PC layout for 10 GHz power amplifier.

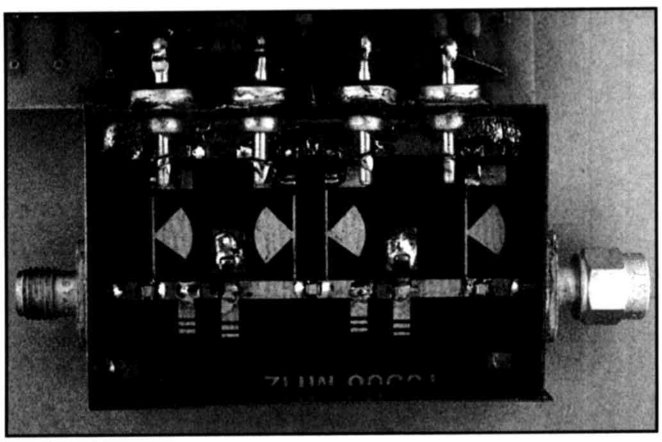


Fig 7—Photograph of two-stage GaAsFET amplifier for 10 GHz.

bility above 4 GHz, but shows a potential instability at lower frequencies, which is readily cured by adding resistance from gate to ground. A 220 ohm chip resistor was adequate for stability and does not appear to affect the 10 GHz gain; the only drawback is that it draws current from the negative bias supply, so that some of the smaller voltage inverter circuits may not provide enough current.

Dual Mixer

The 5760 transverter uses a dual mixer, with separate mixers for transmit and receive, so that each path may be optimized for its function, and separate amplifiers for transmit and receive are easily added. In fact, for 5760 I used different types of mixers for the two paths. However, when I tried to scale this design to 10 GHz, the 90° hybrid coupler balanced mixer used for the receive was no longer practical—the wide, low impedance transmission lines would be wider than their length. One alternative would be the use of thinner Teflon PC board material, which would make the lines proportionally narrower; however, this would make the high impedance line sections very narrow. Since I wanted to use the more common $^{1}/_{32}$ inch thick material, the other alternative was to change the receive mixer design to the 6/4 1λ "rat-race" coupler type used in the transmit mixer. Now the linewidths are reasonable for my basement PC board techniques.

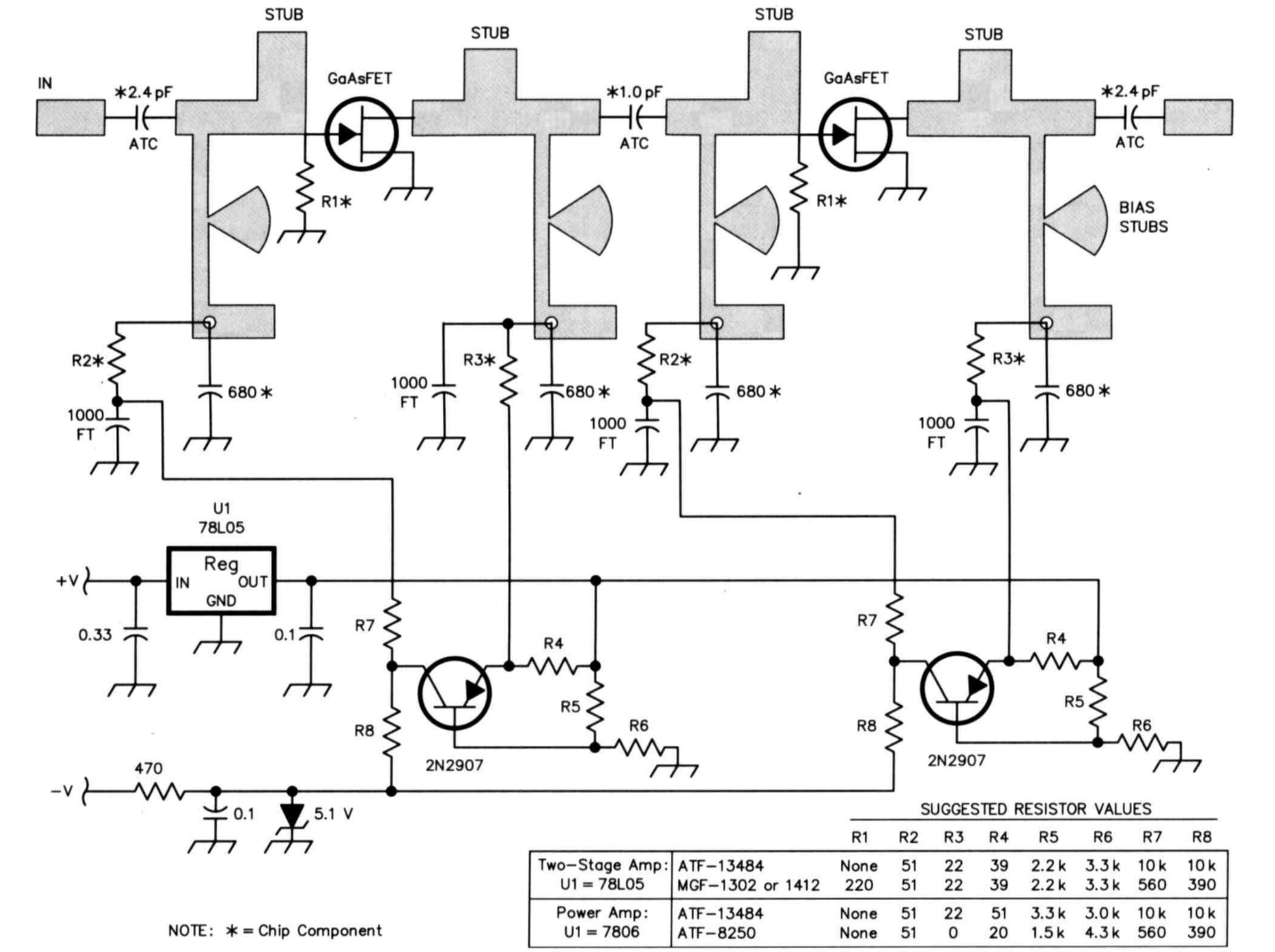
The complete mixer integrates the "pipe-cap" filters onto the mixer board, as originally described[5] (in German) by DJ6EP and DCØDA, in a transverter for 5760 MHz. The filters are made from copper plumbing pipe caps for $^{1}/_{2}$-inch copper tubing, tuned with a 6/32 brass screw, with probes of #24 wire projecting $^{5}/_{32}$-inch above the ground plane inside the filter. The probes are $^{5}/_{16}$-inch apart A cross-section sketch of a pipe-cap filter is shown in Figure 3. I started with dimensions from the measurements WA5VJB[6] made on individual filters. However, the difference in construction techniques seems to make a significant difference in performance at 10 GHz, so I chose the length after making the measurements plotted in Figure 10, for a 3 dB bandwidth of about 140 MHz with loss of about 4 dB.

PC board layout is shown in Figure 4. Other than the filters, the only components on the board are the mixer diode pairs and a 51 ohm chip resistor termination. IF attenuators like those in some of the no-tune transverters would also fit, and are recommended for the transmit side. No through holes are needed for grounding—the radial transmission line stub acts as a broadband RF short. The diodes I used (Hewlett-Packard HSMS-8202) are inexpensive Ku-band mixer diode pairs.

Mixer Performance

Performance of the 10 GHz dual mixer is not as good as the 5760 version—we are pushing the limits of these simple techniques. Conversion loss is around 13 dB, partly due to the 4 dB filter loss. This is part of the compromise for making it simple and inexpensive. However, we can overcome the loss with the amplifier described above.

Mixer performance is dependent on the local oscillator



		R1	R2	R3	R4	R5	R6	R7	R8
Two–Stage Amp: U1 = 78L05	ATF–13484	None	51	22	39	2.2 k	3.3 k	10 k	10 k
	MGF–1302 or 1412	220	51	22	39	2.2 k	3.3 k	560	390
Power Amp: U1 = 7806	ATF–13484	None	51	22	51	3.3 k	3.0 k	10 k	10 k
	ATF–8250	None	51	0	20	1.5 k	4.3 k	560	390

Fig 8—Schematic diagram of 10 GHz amplifier.

power. The HSMS-8202 diodes require less LO power than most, which is usually an advantage. Low LO power will limit the maximum output power from the transmit mixer, while excess LO power will increase the receive mixer noise figure, so the optimum point is a compromise. Since the LO power splitter on the mixer board is not perfect, one mixer may get more LO than the other, so test both sides as both transmit and receive.

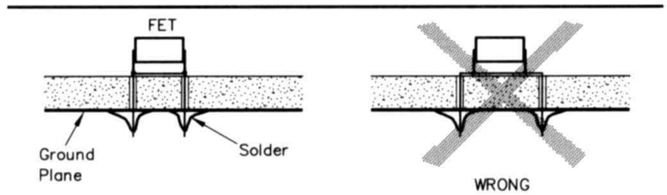


Fig 9—Achieving minimum lead length (by DB6NT).

Mixer Construction

Construction is with minimal lead length on a Teflon PC board, with soldered sheet brass around the perimeter for SMA connector attachment. This is the procedure I use: The copper pipe-cap filter should be installed first, on the ground-plane side of the board. In preparation, I drill tight-fitting holes for the probes and make clearance holes in the ground plane around the probe holes. Then I measure from the holes and scribe a square on the ground plane that the pipe cap just fits inside. Next I prepare each pipe cap by drilling and tapping (use lots of oil) the hole for a tuning screw, then flattening the open end by sanding on a flat surface. Then I apply resin paste flux lightly to the open end and the area around the screw hole. A brass nut, added to extend the thread length, is held in place by a temporary stainless-steel screw (Solder won't stick to it). Then I center the open end in the scribed square on the PC board—the flux holds it in place. Finally, I fit a circle of thin wire solder around the base of each pipe cap and nut, push down gently, and heat each pipe cap for few seconds with a propane torch until the solder melts and flows into the joints.

After everything cools, the temporary stainless-steel screw should be replaced with 3/4-inch long brass tuning screws and locknuts. The remainder of the assembly is performed with a soldering iron, using the photograph of Figure 11 as a guide.

Local Oscillator

Microwave local oscillators normally start with a crystal

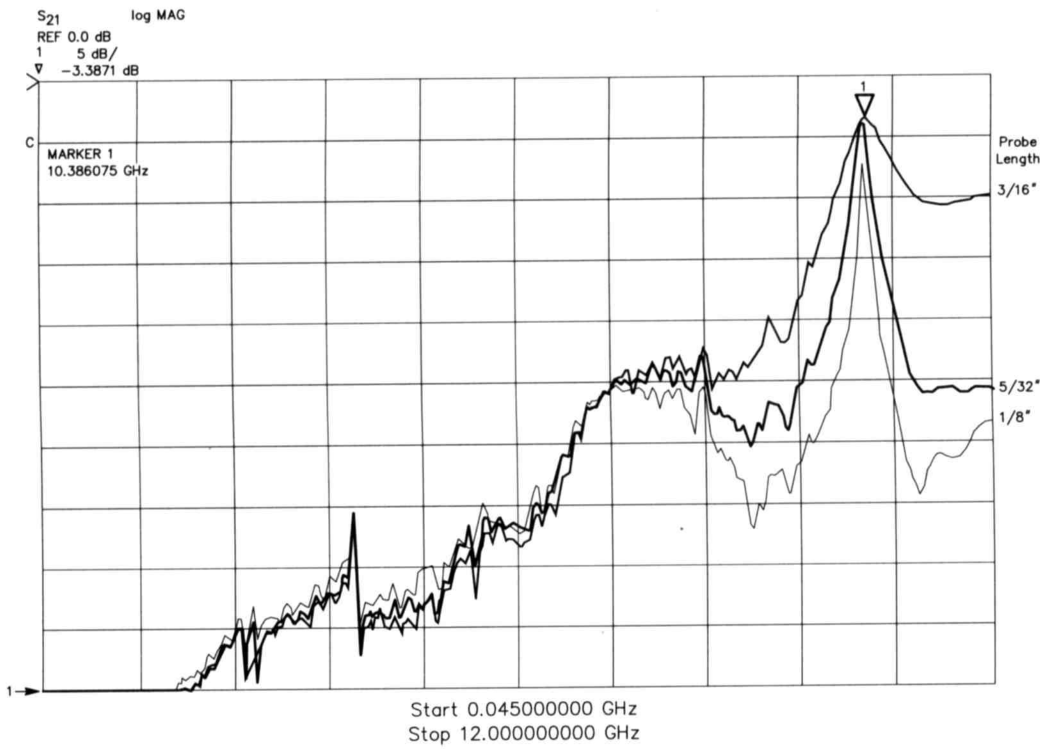


Fig 10—Measurements on ½" pipe-cap filter.

in the 100 MHz range, followed by a string of multipliers. For 10 GHz, a multiplication factor of 100 or more is necessary—not an easy task. However, most of the work has been already done in making local oscillators for the lower microwave bands, so only one more stage of multiplication is needed. I will describe this approach as well as the surplus alternative.

Phase-Locked Microwave Sources

If you are fortunate enough to find them, there are many surplus Phase-Locked Microwave Sources (often called PLO bricks) available, made by companies such as Frequency West and California Microwave. Most of the available surplus units were used in the 11-12 GHz band, and provide adequate enough LO power for the mixer. Some units have an internal crystal oven; after a few minutes warm-up, stability is comparable to a VHF transceiver. Operation and tune-up of these units has been described by KØKE,[7] WD4MBK,[8] and AA5C.[9] The sources can be used unmodified to provide high-side LO injection, above 10368 MHz, or modified[10] to operate below 10 GHz for normal low-side injection. Unless you are obsessive about direct digital readout, high-side injection, using LSB and reverse tuning, is perfectly acceptable. For CW operation, there is no difference.

Most of the available PLO sources operate on –20 volts. This is only a problem for portable operation. WB6IGP has described[11] a +12 volt to –24 volt converter, and surplus potted converters are occasionally found. A three-terminal regulator IC provides the –20 volts. In order to prevent switching noise generated by the converter from reaching the LO, all this should be contained in a metal box with RFI filtering on both input and output.

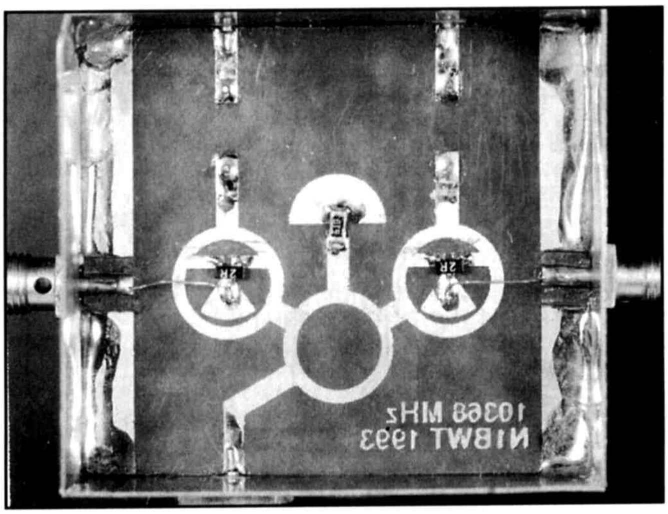


Fig 11—Photograph of 10 GHz dual mixer.

LO Multiplier to 10 GHz

The PLO bricks are relatively scarce, and they do require a substantial amount of power, so a multiplication from a lower frequency is an alternative. One attractive choice is the 2160 MHz LO for 2304 MHz by KK7B; 2160 multiplied by five equals 10800 MHz, which is perfect for a 432 MHz IF at 10368 MHz. Since the pipe-cap filters have a bandwidth of about 140 MHz, using 432 as the IF will provide much more LO rejection.

I've never had much success with GaAsFET frequency multipliers, so a diode multiplier seemed like the way to go. A recent article in *Microwave Journal* described using back-to-back diodes to favor odd harmonics in millimeter-wave frequency multipliers. Why not at lower frequencies? I used the same diode pair (HSMS 8202) used in the mixer, with both diodes connected from a 50 ohm line to ground to form a back-to-back pair. The diodes are installed by cutting a small slot, slightly larger than the diode package, through the board right next to the transmission line. The diode leads are bent so that the single common lead is soldered to the transmission line and the pair of leads on the opposite side of the package are bent and soldered to the ground side of the board.

The diodes are followed by a pipe-cap filter, just like the ones on the mixer board. With approximately +20 dBm applied at 2160 MHz, output was about –5 dBm at 10800 MHz; our two-stage amplifier will bring this up to an adequate LO level. The multiplier output is relatively insensitive to the input power. To test the output spectrum, I turned the tuning screw in the pipe-cap filter; the only other response was at 6480 MHz, the third harmonic, so the back-to-back diodes do favor odd harmonics. Replacing them with a single diode reduced the output at 10800 MHz by about 6 dB, further proving the effectiveness of this configuration.

The KK7B[12] 2160 MHz source has an output of about +10 dBm, so some amplification is required to drive the multiplier. W3HQT suggested the MRF-951 bipolar transistor for this stage, and it works well, providing more than +20 dBm output with about 9 dB of gain, operating at 8 volts and 50 mA. PC board layout, shown in Figure 5, includes the pipe-cap filter; the multiplier diode is opposite the printed arrow. The schematic diagram is shown in Figure 12.

Construction is like the two-stage amplifier, except that everything is much larger! For tune-up, I assemble everything but the diodes and pipe-cap filter probes, then bridge the filter with a brass strap, and test the amplifier at 2160 MHz (The amplifier alone makes a nice power amplifier for a 2304 MHz no-tune transverter). After verifying that the amplifier works, I remove the strap and install the diodes and probes. Then the filter is tuned for output power at 10800 MHz (the first response as the screw is turned inward) and the amplifier tweaked for maximum output.

Power Amplifier

The two-stage amplifier described above has a one-dB compression point around +13 dBm, or 20 milliwatts. This is hardly QRP at 10 GHz, since we have been using barefoot mixers with 20 dB less output, but more power is always attractive. I reviewed the available higher-power devices, and most of them seemed to have little gain left at 10 GHz, so many stages would be required. One possibility was the AT-8250 (ATF-25170), offering moderate power with usable gain. A two-stage amplifier was designed with a ATF-13484 as a driver for increased gain, with the layout shown in Figure 6.

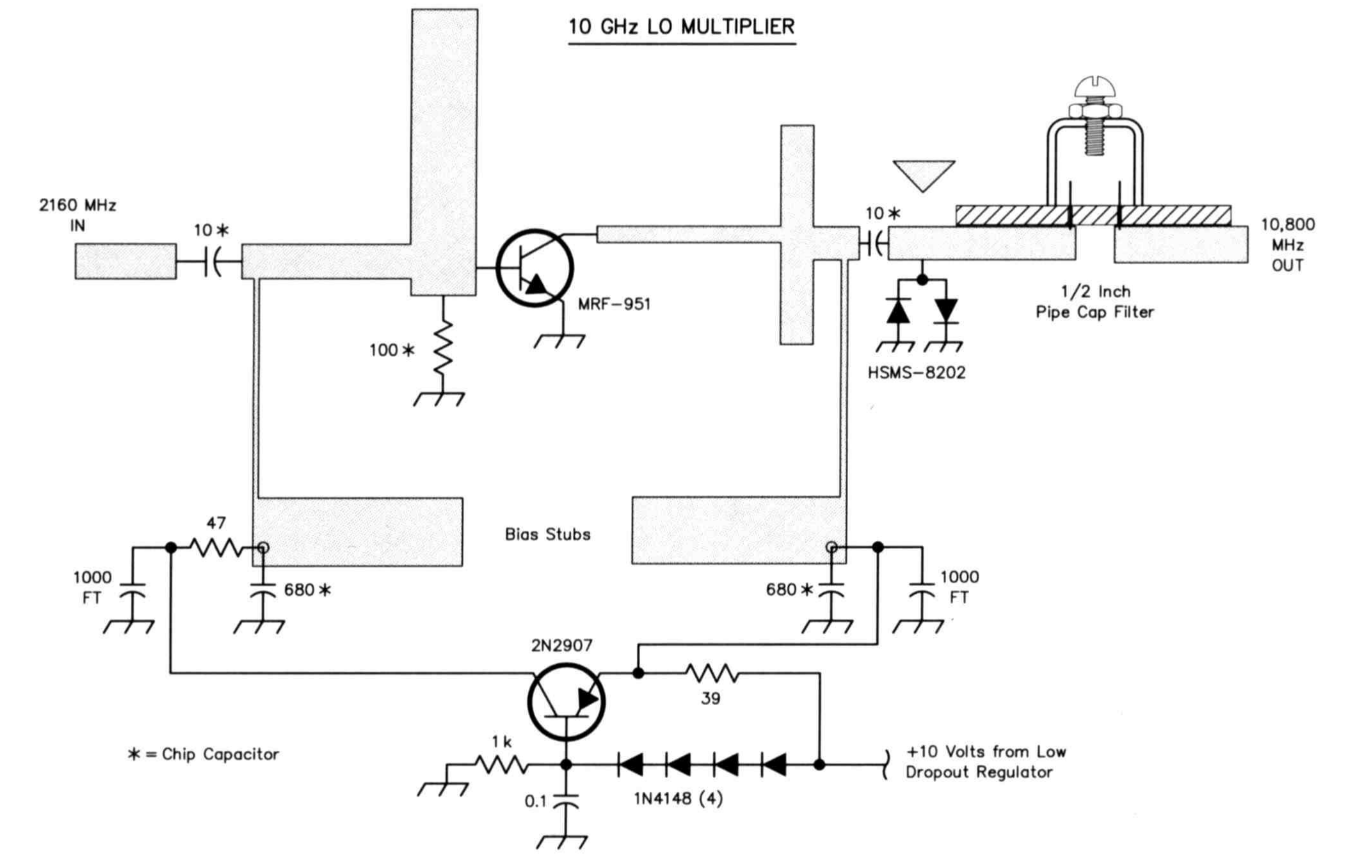


Fig 12—Schematic diagram of 10 GHz LO multiplier.

The amplifier has about 13 dB of gain and a one-dB compression point of about +19 dBm; a bit more drive yields 100 mW output—respectable power at 10 GHz.

To provide the additional power, the AT-8250 is biased to 5 volts and 50 mA. by changing component values in the bias network, as shown in Figure 8. Otherwise, it is so similar to the two-stage amplifier that a separate schematic is unnecessary.

Tune-Up

This is not a no-tune design. Components and PC boards are just not consistent enough for that. However, they are close enough that things will work if assembled correctly, but need some tuning for optimum performance. Obviously, the tuning screws in the pipe-cap filters must be tuned to frequency, but the amplifiers and mixers also benefit from fine tuning. This is accomplished by sliding little pieces of thin copper or brass around on the printed transmission lines, looking for "hot-spots" which significantly change the output. If the output increases, the metal piece can be soldered in place to make it permanent. On the other hand, if the output decreases, a bit of the printed metal may be trimmed off with a sharp knife. Make small changes! It doesn't take much at 10 GHz, and a few small improvements can add up.

Transverter Block Diagram

Figure 13 illustrates one possible block diagram for a transverter using the building blocks described above. Expected signal levels are shown at each point. The isolators[13] shown are optional, but recommended to ensure stable operation if you have them—they frequently appear at our local flea markets at very reasonable prices. An isolator will allow RF to flow in only one direction, protecting amplifiers from load mismatch (and often prevent oscillation). For the preamp, I'd recommend the WB5LUA[3] design.

An output filter is highly recommended, particularly if mountaintop portable operation is contemplated. Most accessible high places are teeming with RF microwaves, so a good filter will keep them out of your transverter. One excellent one is the waveguide post filter described by N6GN,[14] which is easily built in a section of surplus waveguide and provides performance far superior to the pipe-cap filters. Since most 10 GHz antennas are fed with waveguide anyway, the filter fits well at the output.

Conclusion

Not surprisingly, working at 10 GHz is more difficult than the lower microwave bands. Everything is more critical—some things can be tuned by tightening an SMA connector (the only kind that works). Test equipment is necessary; you don't have to have a fancy lab, but having some way of measuring power and frequency is essential. All the equipment I used for tune-up has been obsolete for years, and most was found at flea markets. Later I was able to use a fancy Network Analyzer to generate some of the plots in this paper.

What was encouraging is that a circuit design can be analyzed on a computer, fabricated using basement printed-circuit techniques, built with tinsnips and soldering iron, and work rather reproducibly if built with care. On the other hand, the amplifier instability described above, which was caused by a little extra ground lead length, was a real eye-opener.

The building blocks described should make it possible for someone wishing to get on 10 GHz to assemble a transverter by building one piece at a time, checking it out, making contacts, adding on for increased performance, and learning more with each step. Also, all PC boards and critical parts are available from Down East Microwave, eliminating the often daunting task of finding that one last elusive part.

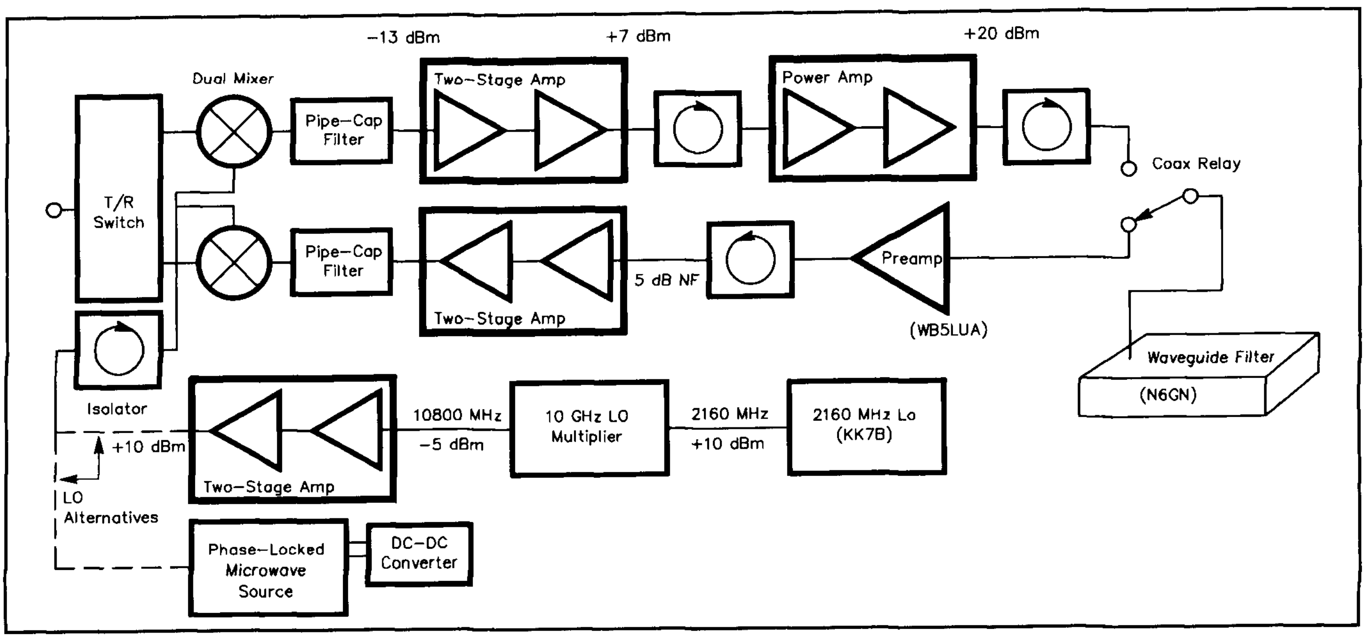


Fig 13—10 GHz transverter block diagram

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Just About as Cheap as You Can Get 10 GHz

The guts from a RADAR detector, a Ramsey Electronics FR10, and a 2N2222 modulator puts you on the air.

By Kent Britain, WA5VJB

(From *Proceedings Of 1996 Microwave Update*)

The Ramsey FR10 is easily converted into a very sensitive 30 MHz wideband FM microwave IF. The FR-10 has two filters in the IF section, simply bypassing the second filter and letting the first 10.7 MHz filter do its job, makes the FR10 a 200 kHz wide receiver. I chose to reduce the tuning range of the receiver to about .5 MHz, but this was a personal choice. The NE-602 Mix/Osc IC can be crystal controlled. See the February '92 issue of *73 Magazine*, page 21 for some suggested circuits:

Figure 1 is the only mod you have to do; replacing FL2 with a .01 mfd cap makes the FR10 a wideband receiver. The .01 goes from pin 3 to pin 5 of U2.

Figure 2 is the audio stage. The FR10 is designed to drive a speaker and is far too hot for headphones. (On Gunnplexer systems you need headphones to prevent feedback.) Omitting C35 (it was optional anyway!) and adding the 330 ohm resistor worked well.

Figure 3 was my tuning range mod. Replacing C12 with a 12-15 pF and adding another 12-15 pF across L3 tightened up the tuning range. Replacing C12 with a 22 pF will also work giving a 1 MHz tuning range and the ultimate would be the *73 Magazine* mod using a 40.7 MHz xtal. Alignment is quick. Put the tuning control in the middle of its range and listen to a 30.0 MHz signal while adjusting L3.

Figure 1 also has a mod suggested by Al Ward. The 3.9K resistor across the quadrature detector coil broadens the FM detector response and improves audio response with highly deviated signals.

The RF head in Figure 4 is the guts from a RADAR Detector. To retune the Gunn oscillator from 11.5 to 10.25 GHz you'll need either a Wavemeter, Spectrum Ana-

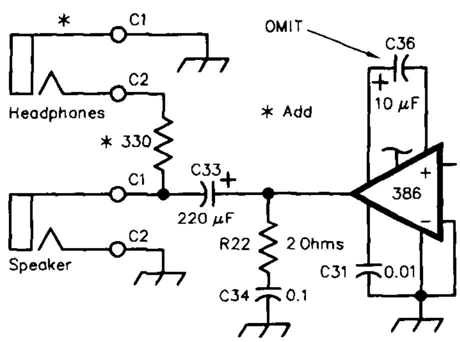

Fig 2—Ramsey FR-10 mods.

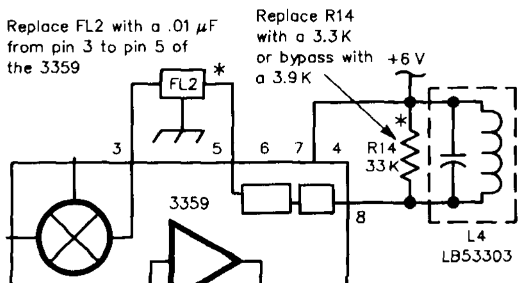

Fig 1—Ramsey FR-10 mods.

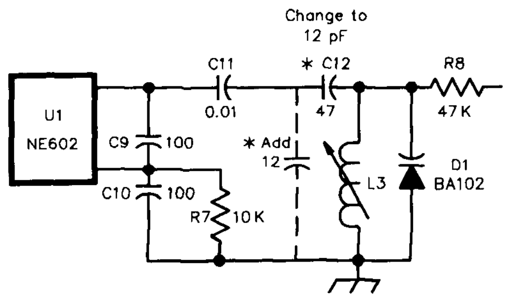

Fig 3—Ramsey FR-10 mods.

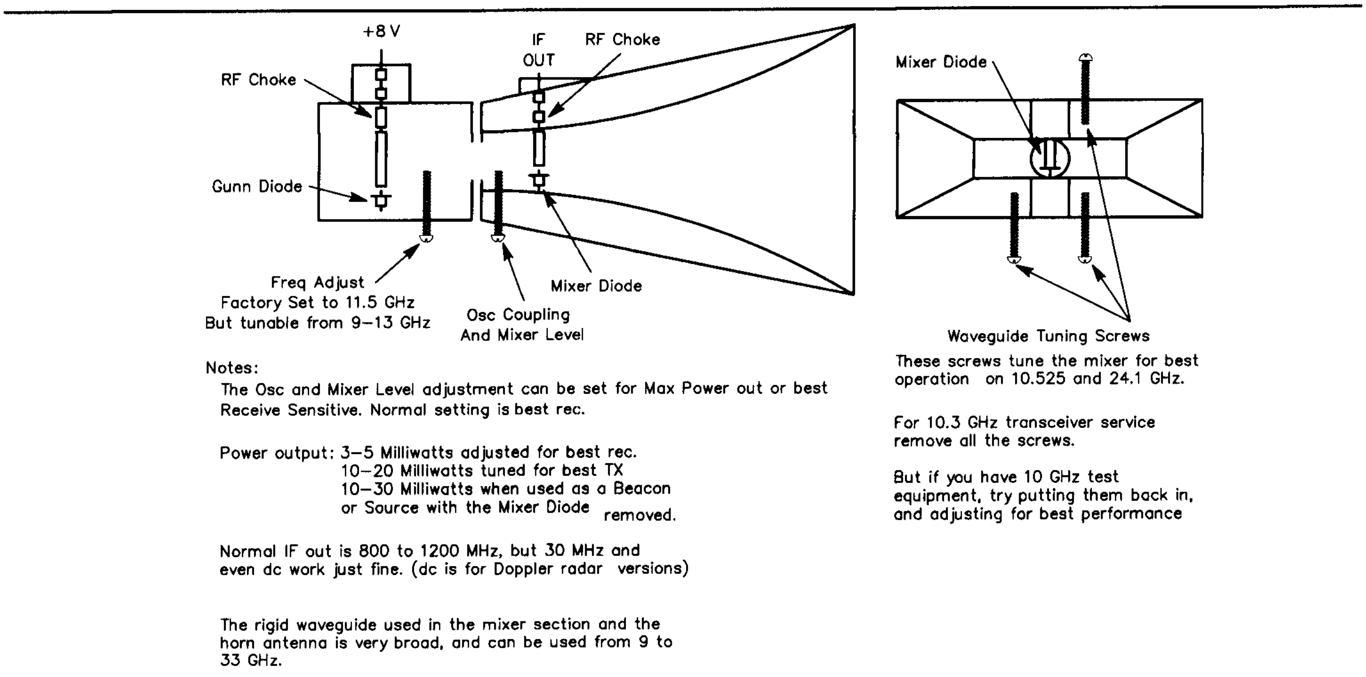

Fig 4—Typical radar detector RF section.

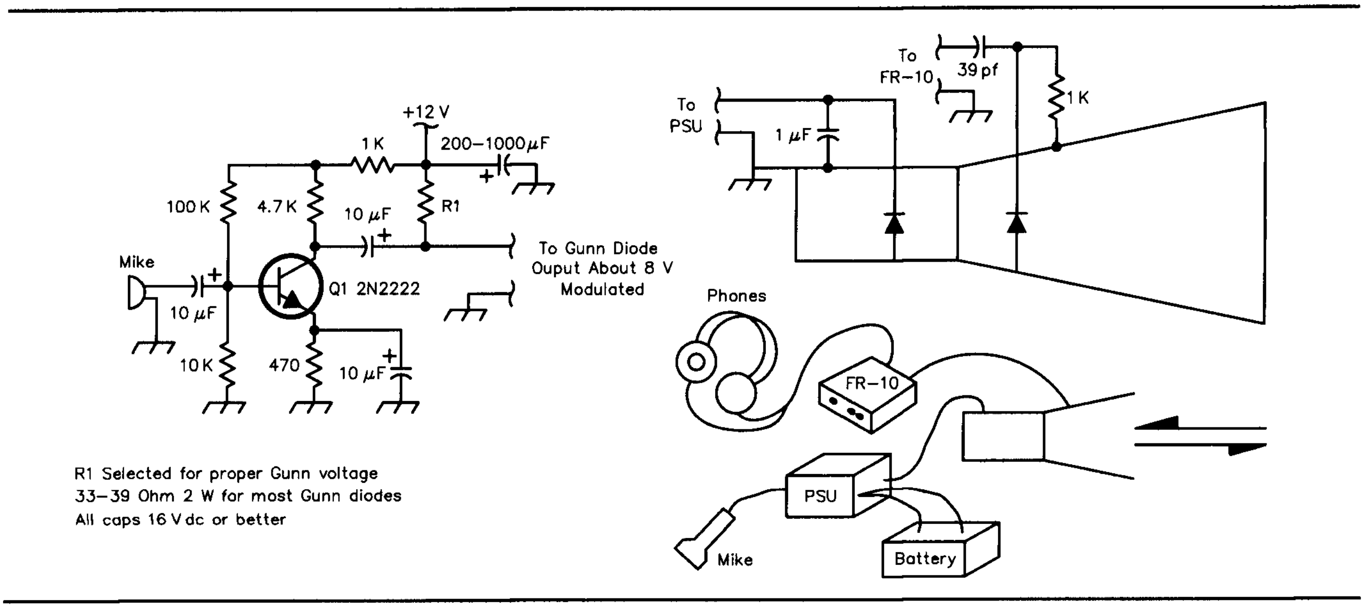

Fig 5—Power supply and modulator connections.

lyzer, or an EIP type microwave frequency counter. A friend with test gear is very valuable at this point.

The 1K resistor and 39 pF capacitor in Figure 5 are typical values. I usually use a 10-80 pF trimmer and tweak for best sensitivity. The 0.01 to 0.1 μF cap across the Gunn diode keeps down noise and VHF parasitics.

It only takes 0.1 volt of audio to wideband FM modulate a Gunn oscillator. This modulator is simply a 2N2222 audio amp taken directly out of the *ARRL Handbook*. No ALC or compressor circuits are used. On 10 GHz FM you hear your own signal coming back, audio weak—speak up, too loud — back off the mike.

The 39 ohm 2 watt resistor drops the 12 volts to 8 volts at the Gunn diode. This can be tweaked for best results and is a typical value. Varying the 12 volt source will give you a limited tuning range.

Use shielded wire between the modulator and Gunn diode to keep down hum and RF rectification.

More complicated and versatile PSUs (Power Supply Modulators) are described in *The RSGB Microwave Newsletter Technical Collection* and the *RSGB Microwave Handbooks*. This design was kept as simple as possible. PS: Works with 24 GHz Gunns as well.

See ya on 10 GHz!

A "Tri-Band Booster"—or "Improving 2, 3 and 5 GHz at WA1MBA"

By Tom Williams

(From *Microwave Update '97*)

Problems with SHF from the Home Station

As some of you know, I run a home station that is active on all UHF and SHF bands (24 GHz to come soon). After being on the air for several years, and upgrading all aspects of the station, I have been bothered by the limited DX on 2, 3 and 5 GHz. Sure, I can work the mountaintoppers out to 100 miles or more, but there are good stations that I can't work during contests or other times. Simply put, this station could do a lot better in transmit power and receive noise figure.

As you can see, in Figure 1, there is a lot of cable resulting in significant losses. My original intent with this station was to put the LNAs as close as possible to the feed. This configuration is quite acceptable on the "low bands," but at 2304 and up the loss between the top-of-the-tower box and the feed is too great. I have about 7 dB of loss ahead of the 5 gig LNA, 4 dB on 3 gig and about 3 dB on 2 gig. This results in unacceptable system noise figures of 4 to 8 dB, whereas a good terrestrial station should achieve under 1 dB on all these bands. On the transmit side, antenna power levels are 2 watts on 2 gigs, and about a hundred milliwatts on 3 and 5 gigs. I figured that about 10 watts on each band would be enough to work the DX that I am missing. The only way to fix all of this is to put power amps and LNAs as close as possible to the feed.

The Plan

The perfect system would have a wideband high power amp (2 to 6 GHz) and a wideband LNA. This way, two components would be needed, along with simple relays to boost the transmit and receive signals, and my problem would be solved with an elegant system. But elegance is not in the cards.

Unfortunately, high power amplifiers that cover that range are very tricky to build and many thousands of dollars to buy. There are TWTAs that would cover the range, but this would mean having high voltages and a sensitive and expensive system up top. If I owned two miniature 10 W TWTAs (one for a spare), I would probably use them. If I ever design a solid state 10 W amplifier that covers all these bands, I will certainly share it with the community.

On the receive side, wideband preamps do exist. LNAs that cover these frequencies can be built with noise figures below 1.5 dB, but are very expensive, and no designs that I know of have been made public. One can achieve about 2+ dB with a GaAs MIMIC. I wanted to achieve under 1 dB noise figure, preferably 0.8 dB at 5 GHz, and 0.4 dB at 2 GHz. Fortunately, single band designs and parts that achieve this performance level are readily available.

Given all the above considerations, I set out to build separate amplifiers and preamplifiers for all three bands, and came up with a concept for the booster, based on connecting them with seven-port coaxial relays and appropriate power switching. I also included output power monitoring and a bypass connection in case any device becomes inoperative (or suspect).

Table 1

Power Loss and Gain Needed to Achieve 10 W at the Feed

Band	*Power in Shack*	*Power at Feed*	*Cable Loss*	*Gain Needed to Achieve 10 W*
2304	41 dBm (10+ W)	34 dBm (2+ W)	7 dB	6 dB
3456	33 dBm (2 W)	18.5 dBm (65 mW)	14.5 dB	21.5 dB
5760	37 dBm (5 W)	20.5 dBm (0.1+ W)	16.5 dB	19.5 dB

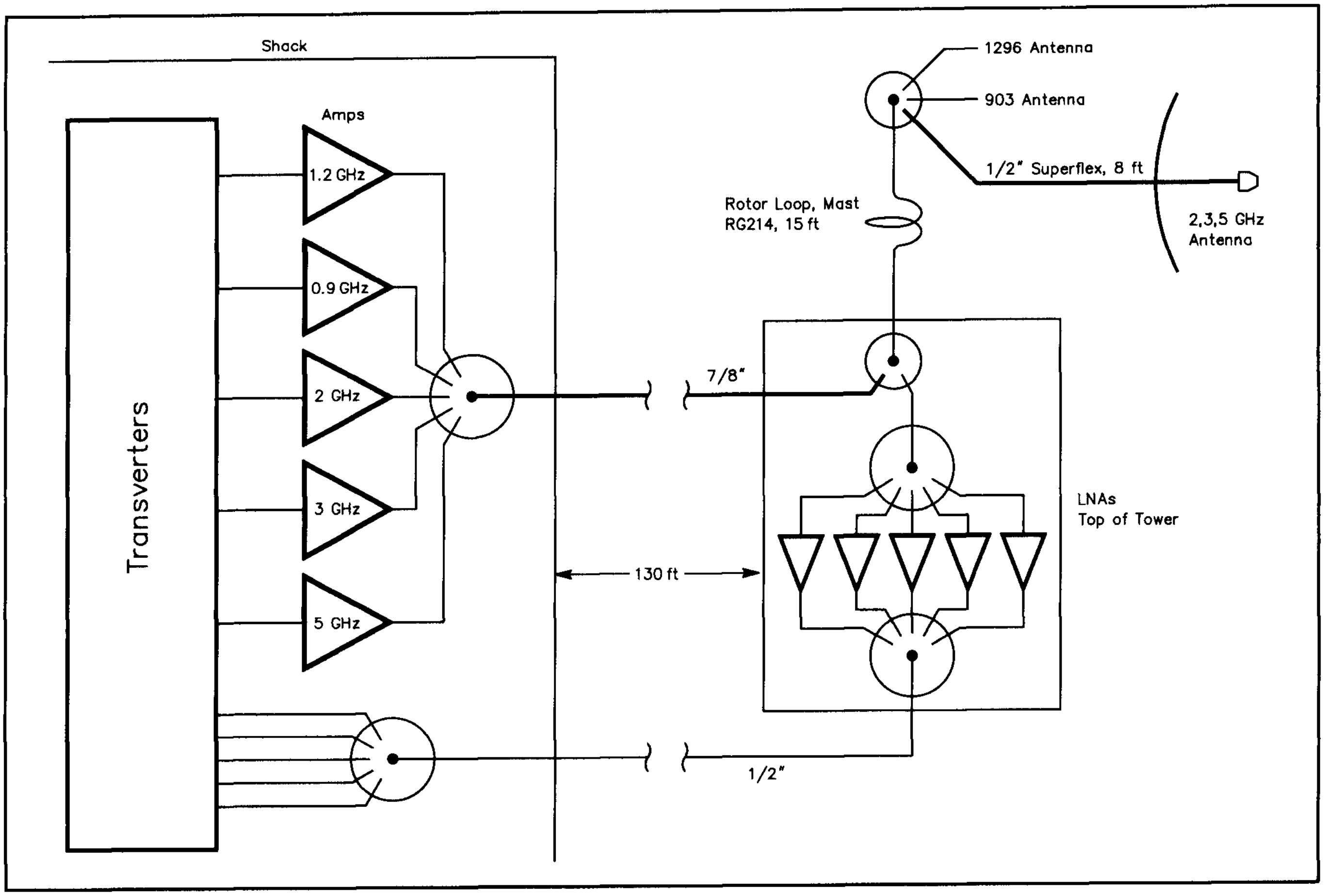

Fig 1—WA1MBA 0.9 through 5.7 GHz before booster.

The Experience

Decisions, decisions. . .

First, I had to decide whether to buy or build. It turns out that there are some sources of ready-made amplifiers for these bands, the higher power ones are very (very) expensive new, but there are occasional surplus amplifiers showing up at flea markets at a fraction of their original price. The flea market prices I've seen for 10 W 5 gig amps is $40 to $150. The Avantek ones take –28 Vdc and are very heavy. They are TWT replacements, having lots of low level gain so that they can be driven by as little as 0 dBm. Although excessive gain isn't a problem, I decided that the weight, and difficulty in making it a lot smaller would nix using this fine amplifier. The $150 five watt surplus unit I found is much smaller, and runs on +12 Vdc. Very nice. But I really wanted an even smaller package than this unit, and I wanted to build as much myself as possible, in order to gain experience. I put my five watter in service driving the transmit coax from the shack. I've also found 3 and 2 gig amps being sold from time to time, but they all were either too large or too expensive—and often both.

I decided to build a 2304 amp that WA3JUF (now W3KM) had described that uses a bipolar device normally designed for high power in class C operation. His design uses a heavy bias supply, resulting in class AB operation (clean signal) with about 10 dB of gain and an output around 10 W. This was just what I needed for this band. I asked Dave for some help, and he was glad to assist me with a spare device and a design. I found a dead amplifier (originally built for 903) that had a milled out slot for the transistor that was an exact fit. The resulting 2304 amplifier is quite small. The only disadvantage is that it needs heavy current at +28 V and –5 V.

For 3 GHz, I went with a DB6NT design that Eisch Electronic of Germany had advertised. It is based on MGF (I think) 904 and 905 FETs, which are not expensive. They had the board, an aluminum slab for a base, and a parts list. This looked like a good next project for me, not too difficult to step up to 3 gig after 2 gig. It is described in the DUBUS Technik books. It has an extra input stage that would allow it to achieve full output with about 50 mW of drive. Its specifications were just what I needed for the band.

The 5 GHz band starts to get tricky. Construction detail and quality become even more important than on the lower bands. I decided to try a DB6NT design. I had no base plate, so I was going to have to learn simple machining with no experience.

For receive LNAs, I knew that I had to copy an LUA or a KH6CP design in order to get the performance I desired.

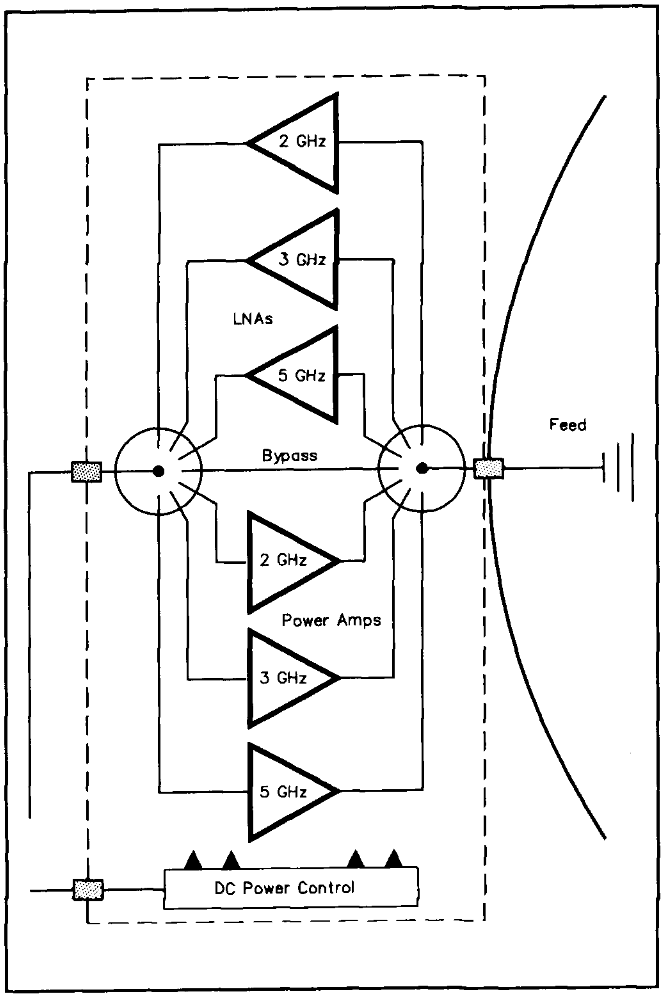

Fig 2—Three-band, back-of-the-dish booster.

Down East Microwave had begun to make kits of the LUA designs (around the ATF-36077). I picked up the 5 gig board and parts, and found a suitable enclosure in the junk box. For the 2 and 3 gig units. I ordered the enclosure from DEM as well.

Acquiring the Parts

Before building anything, it's a good idea to acquire all the parts. In the case of the 2 gig amp, I already had all the parts. For the 3 gig amp, I decided to order the board and the aluminum base plate from Eisch Electronic in Germany. They take credit cards and when I called I found that they would speak English (a lot better than I could speak German!). Eventually, I placed two orders by fax (I didn't include everything on the first order). It took about three weeks to get each order. Because of the added costs, I only ordered the parts that I could not readily obtain in the US. I got most of the standard parts (surface mount electrolytics, integrated circuits, voltage regulator) from Digi-Key, Mouser, and/or Newark. I obtained most critical RF components (surface mount resistors, ATC capacitors and some active devices) from Down East Microwave. Altogether, I spent about $85 on the parts for the 3 gig amp. This cost included overseas shipping and money exchange. Some of the parts used were from my shack supply, so if I had to buy everything new, including the heat sink, it would have been more like $150 for parts.

For the 5 gig amp, I had been keeping my eyes open for about 3 years for devices, and happened to find a supply of Avantek 5964-3, which are 3+ watt devices, and some Mitsubishi MGFC39V-6472A, which are 8+ watt devices. The first could drive the second quite well. Neither of these devices is designed to work at 5760, although the 5964-3 is typically tested as low as 5700. The 6472 is designed to run (as you might assume from its part number) from 6.4 to 7.2 GHz, so I expected that it would need a little tuning to bring it to power at 5760 MHz.

Table 2

Free Advice for Novice Machinists

Item	*Free Advice*
Placing a hole	Use a fine punch to scribe first, then a larger punch.
Drilling a hole, yes ANY hole	Drill a first pilot with a #55 or the smallest your press will hold, then drill a larger pilot hole, then drill the final hole. Hassle? Yes. Worth it? YES!
. . . continued	Use a tapping or drilling fluid on the bit, such as Aluma-Tap. After each 1/8 inch of depth, clean the bit and add fluid. Your bits will last forever.
Tapping	Use a tapping fluid or wax. Use it every time you insert the tap.
. . . continued	Rotate forward 1/2 to 3/4 turns, and then back 1/4 to 1/2 to break the thread chips. Never force the tap. Always rotate without any sideways force.
. . . continued	Remove the tap every 5 turns or so to clean and re-lubricate.
Cutting	Measure and mark at least twice, cut slightly big and file the edge down.
Goofing up	Take a walk, watch TV, get on the air for a while. Sometimes long projects just need time off in order to clear the mind.

At first I wanted to use a straightforward design by Al Ward for a single stage. It would need to be changed to a two stage and have a bias/regulator power supply added. Since I had not actually made a printed circuit board in about 15 years (and even then I was not working on RF boards), I decided to obtain a DB6NT, two-stage ready-made board as well. This turned out to be a good idea. To make my own board, I tried a laser-printer/iron-on technique from a supplier of a pattern transfer kit. There were many descriptions of what can go wrong and what to do about it. After three tries, I thought I had good transfers (onto an expensive piece of Duroid) and went ahead with etching. The results were not good. Next time I try to make my own RF boards. I will either directly lay tape on the board. or use a photoetch technique. This time, I would use the DB6NT board. That board is two stage, where the first FET is meant to be a much smaller one than the 5964, so I had to squeeze it in.

Parts for the 5 gig amp were obtained from Digi-Key and Down East. There was no baseplate, so I decided to take a simple approach to creating one. This is described below. I obtained metals and hardware (those tiny 080 hex cap screws, tapping tools, etc) from Small Parts, Inc.

The biggest problem that I have with acquiring parts is that I always forget something. Usually this means another minimum order, which can mean $25, even if I only forgot a $0.32 resistor. At first I thought that I would get better at making sure that I would order everything, and perhaps I have become better. Lately I have adopted another technique, and I find this quite satisfactory. I keep a list of items that I would like to have, but don't absolutely need. Then, when I goof and have to place a small order, I first ask my friends if they need any small parts (to pad the order) and if I need more, I order from my "wish list." The other advice I have is to be as generous as possible with those hard to find parts in the stash. Generosity is usually paid back with interest.

Building

Oh my, isn't construction fun! I get both great frustration and satisfaction from building things myself. There are several aspects of building an SHF amplifier that can each be a challenge. The two construction challenges are soldering those tiny things (just where you want them), and the second is drilling, tapping and cutting aluminum (just how you want it).

After years of practice, I am getting to an acceptable level (although far from the best) at soldering—actually making the electronics. The only challenge for the 5760 amp was that there was no description of this amplifier in any of the texts I had from DUBUS. I was either missing the issue, or this particular amp wasn't published. In any case, the only important unknown was the placement of the power supply parts, and it turned out to be a fairly standard design, so that was not a problem.

When it comes to the mechanical parts, however, I am really a novice. Ken, W1RIL, has figured out how to convert a drill press into a small milling machine, appropriate for this kind of work. He has helped me by making a very nice enclosure and base for a 10 GHz amplifier. Someday I hope I do machining as well as he, but I don't know if I have the patience. I built the 3 GHz amplifier enclosure by bending hobby brass in two "Ls" to surround the aluminum base. The transistors for that amp are shallow enough that slots do not need to be cut in the base. It worked! In the process I learned all about drilling and tapping 6-32, 4-40, 2-56, and yes 0-80 holes. I've broken several taps, drilled some holes in the wrong spot, and made a few cuts at a bad angle or too short. It took me a while to learn some of the following things, that I thought I would summarize to help you if you are also a novice amplifier builder.

I don't doubt that you could learn these and more from a machinist, or in a good high school shop, but somehow I missed that year of shop—we seemed to always be working with wood. I doubt that walnut would make a good amplifier base. Anyway, I'm not the expert, so I offer the following as "free advice" (Table 2) and hopefully it will be worth more than it costs.

Because I have not converted my drill press into a mill, I could not cut slots into the base plate for the transistors in this amp. Instead, I made a second, "false base" of thin aluminum between the board and the real base (see Figure 3). By cutting slots into this false base with a nibbler tool, and drilling it with "through holes" for all the screws that hold down the board, the power FETs could be recessed as needed to keep their leads at the level of the striplines on the board. Well, this was the idea anyway.

Not following my own advice, I calculated the wrong depth for the transistors, and so the 0.032 aluminum sheet was a little too thin to achieve the proper depth. So, the power FET leads had to be bent down to the board approximately 0.02 inch. This sounds small, but I felt pretty lucky that I didn't end up with instability. Two layers of 0.032 would have been just a little too much. I do not recommend this approach unless you absolutely cannot get the slots machined out properly. If you decide to use this approach, try to use material of the right thickness, and "measure twice" with a caliper.

It is often recommended that you use silver loaded epoxy to glue down the board so that a good RF contact is made to the base, resulting in a good match to the transistors and the RF connectors. I agree that this is good advice, but not always needed. At 10 GHz and above, this is probably always the case. At 5 GHz and below. I recommend constructing without the epoxy. Then, when testing, press down hard around the FETs and the connector parts of the board

Table 3

Input/Output of the 2-Stage 5760 Amplifier

dBm input	*dBm output*
16.5	36.5
17.5	37.4
18.5	38.2
19.5	39.0
20.5	39.4

Saturation is reached at 20.5 dBm input, the maximum drive power available.

to see if the contact resistance is contributing to any loss of performance. If it is, then silver epoxy is in order. Otherwise, don't bother. My board had plenty of screws in those critical places, and the epoxy was not needed.

By the way, this epoxy is also quite expensive and has a short shelf life. Unfortunately, it doesn't take a lot of epoxy to glue down a board. Because the material tends to harden in the tube, and any exposure to air will make this worse, I would recommend using the entire epoxy kit at one time. So, to reduce waste, it's a good idea to store up all the projects that need silver epoxy and do them all in one sitting.

Testing

One should always perform a dc test before placing the power FETs into an amp. This is true of LNAs as well. Make sure that the amplifier can handle the power load without significant change in voltage. To do this, temporarily substitute for the FETs power resistors that will draw the same current. Set the bias supplies to −1.5 V. If you have a 'scope, try to make sure that the bias comes on before the $+V_{dd}$ supply. Then install the power FETs. The next step is to set bias. With a load on the input and output of the amp, adjust bias for the right idle current for each device. The big-boy 10 watters can take as much as 2.4 A of drain current at the proper bias setting. This represents over 20 watts of dissipation! Things get hot quickly. Once bias is adjusted, it's time to tune up the amplifier.

On 2 gig amps, it's common to have variable piston capacitors as part of the tuning process. On all stripline solid-state amplifiers (even on amplifiers with tuning capacitors), the best gain and power can be achieved through some "snowflake" tuning. This kind of tuning amounts to placing small pieces of copper foil along the striplines of the power FETs and soldering them in place. It's a process that can take hours, even several sittings, before the output and gain are maximized. Usually, the best results are obtained with two pieces of foil along each section of stripline. Unfortunately, there is sometimes not enough space to place them exactly right, so three pieces might be used. I have found that in some cases only one piece or rarely, no pieces, are needed. Anyway, here is the process.

Set up a known input level at about 10 dB below the expected input level needed to drive the amp to saturation. It is best to have a dial attenuator in the input line. Attach an output level meter. Use a known (calibrated) directional coupler and a good quality load. If a load with a great return loss at the frequency you are testing isn't available, then a long length of good quality coax with a less-than-excellent load at the end will suffice. Make sure that the coupling factor plus the maximum reading of the power meter exceeds the maximum output of the amplifier—in other words, don't set the system up in a way that might burn out your power head! If there is excess power, then insert attenuator(s) between the coupler and the power head. Now you are set up for gain/power measurement (see Figure 4).

To tune a solid-state microwave amplifier, you need to make a set of snow-flake tools. Find some insulating sticks, such as pieces of Teflon or plastic rod, or tuning tools, or wooden Q-tips, glue different size pieces of copper foil to the flattened tips, ranging from about 0.05 to about 0.15 inch on a side. With the unit powered and RF applied, place different size foil along the striplines to find the "hot spots," where the output jumps up. I usually go from the input end to the output, but end up going back several times to all hot spots anyway. Once you have decided where to place or change a piece of foil, shut off power, solder a best size piece to a hot spot, reapply power and RF, and see what the improvement is. Always keep track of the power level to be sure that things are improving. Try to extend the size of the piece you just added with a small piece. If it improves, shut it off, replace with a bigger piece, and try again. If it does not improve, you can go on to the next hot spot. Sometimes, I find that I can go back after tuning all the hot spots and reduce or enlarge an earlier placed piece to improve the system. During all of this you have to be very, very careful to never short out a bias line. Doing this could easily fry the power FET. Once the tuning is pretty well maxed, increase the RF input power to see how well it performs. It is often

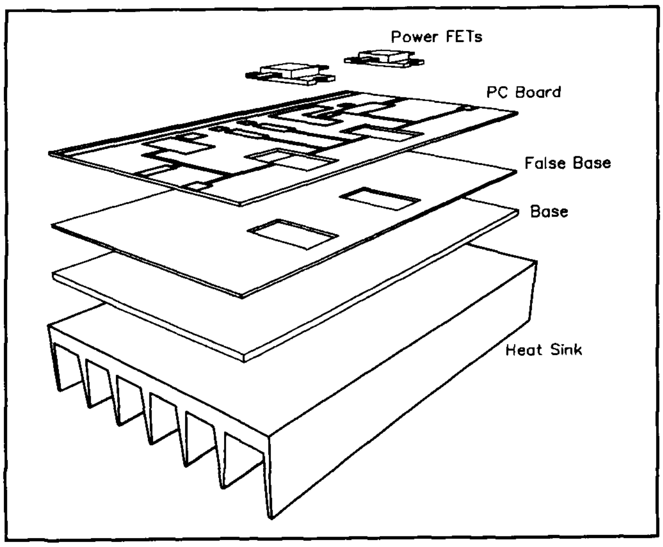

Fig 3—The method used for building the 5 GHz amp.

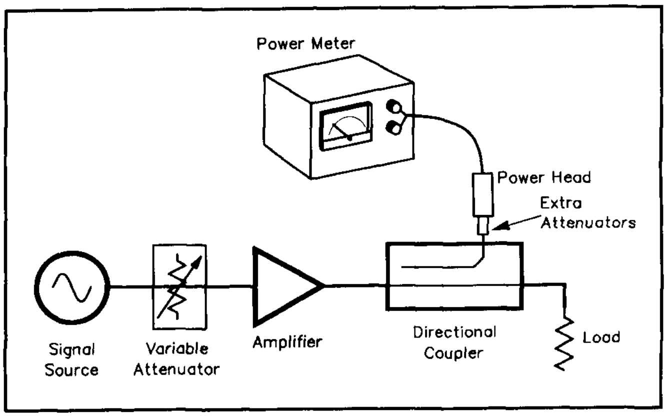

Fig 4—Testing setup.

necessary to readjust the bias at full RF, and then retune the snowflakes for maximum output.

After the amplifier is tuned, I make a chart of the input and output levels, and if there is a power monitor, I record this voltage as well. Although 2 dB increments are sufficient, 1 dB will give a good indication of saturation/power compression. This all helps immediately when determining best input power levels, and later if you want to evaluate an amplifier to determine if there is something failing. I was fortunate, in that the amplifier reached its saturation point at exactly the maximum level that is available at the input to the booster.

I ran each amplifier at full power (key down) for five minutes to determine if there is enough heat sink. Because I wanted these all to fit in a fairly small enclosure, and am also concerned about weight, I used heat sinks that are really too small for the power dissipation needed. I compensated by adding miniature fans to the amps. This kept them all quite cool. Within the enclosure, the air will circulate, and eventually will heat up, so key down time will surely be limited in practical operation. I decided to include a remote temperature sensor in the box to monitor for this condition. Figure 5 shows the components and the box prior to assembly.

In summary, unless you want to use waveguide, the best bet is to mount both power and preamps as close to the feed as possible on all bands 2 GHz and above. I thought that doing this kind of work was strictly for the professionals. Many hams who publish microwave designs are indeed professionals in communications. But many are just hobbyists, like me. If you are willing to learn, you probably can build your own.

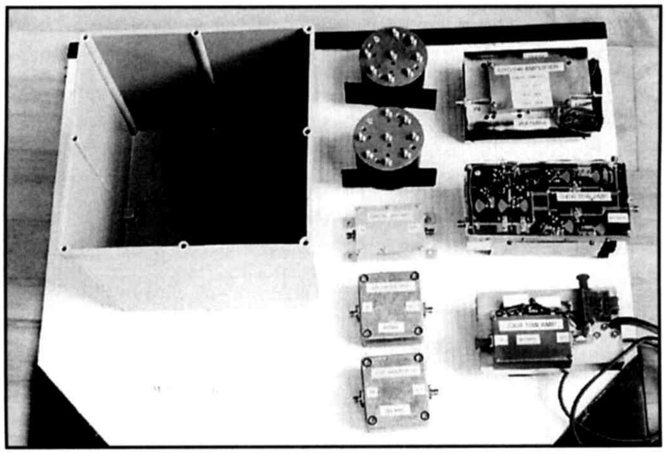

Fig 5—Power amps, LNAs, coaxial relays and box prior to assembly.

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Down East Microwave, 954 Rte 519, Frenchtown, NJ 08825 (Tel 908-996-3584). **http://www.downeastmicrowave.com/**

Digi-Key Corporation, 701 Brooks Ave S, Thief River Falls, MN 56701-2757 (Tel 218-681-6674, toll free order line 800-344-4539), **http://www.digikey.com/**

A High RF-Performance 10-GHz Band-Pass Filter

By Zack Lau, W1VT

(From *QEX*, July 1997)

Why do I call this a high-performance filter? It offers a very high ratio of Q to size and wastes very little volume in coax transitions or empty space. This is quite important when building a compact portable transverter.

The design is based on one of the very first 10-GHz filters I built—the 2-pole band-pass filter in the TNT, or Tuesday Night Transverter.[1] Unfortunately, the authors omitted two important details—the input/output coupling of the filters and the measured insertion loss. While their design had the potential for great performance, I doubt that many people realized it—even those who built one!

This filter is made from a 1.5-inch section of WR-90 rectangular waveguide. An enclosed cavity is formed by soldering brass sheet stock to the open ends of the waveguide. A pair of posts down the center line creates a pair of coupled cavities. The diameter of the posts determines the coupling—narrower posts increase the coupling between the cavities.

The filter is capable of surprisingly low insertion loss, if you can get the coupling adjusted properly. I chose to use the center conductors of captivated contact SMA connectors as probes. The hard part is adjusting them—just ask anyone who has spent hours filing away! My solution was pretty obvious—use shim stock. After all, isn't this what they sell shim stock for? I made a bunch of spacers of different heights—this allowed me to quickly adjust the height of the coupling probes. With this design, it appears necessary to adjust the probes and resonator tuning for lowest insertion loss. The tuning interacts with the probes, so it didn't appear terribly useful to look at the passband shape when setting the probe coupling. This differs from most other filters, where looking at the passband is an excellent guide.

Using a mixer and isolator to measure the insertion loss, the design shown in Fig 1 has 0.6 dB of insertion loss with a 3-dB bandwidth of 106 MHz. The 144-MHz IF image rejection is 33 dB. This should offer excellent performance for receive applications.

For transmit, it might be useful to trade a little insertion loss for better image rejection. With the original design's center coupling using a pair of 3/16-inch posts, the insertion loss increased to 1.3 dB with an image rejection of 47 dB. The 3-dB bandwidth is 36.7 MHz.

Construction

The first step is to get the waveguide, brass metal stock and SMA connectors. Either 2- or 4-hole flange SMA connectors will work, but it is important to use connectors with captivated contacts. Otherwise, the probes will move around, changing the filter alignment. I suggest using brass tubing since it's easier to solder. On the other hand, it might be practical to tap solid brass rod after it's soldered in place—so you can easily mount the filter with screws. Don't forget the clearance required for the 4-40 tuning screws.

As a starting point, I'd try 100-mil probes for the wider filter and 80-mil probes for the narrower filter. These lengths are shortened a bit by the waveguide walls—only 30 to 50 mils actually sticks into the cavity. Adding shim stock reduces the length of the probes, so you might cut the

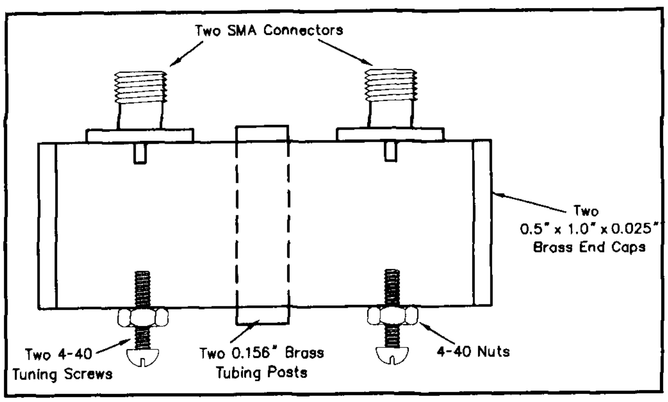

Fig 1—Diagram of the 2-cavity 10-GHz band-pass filter.

[1] Notes appear at the end of this section.

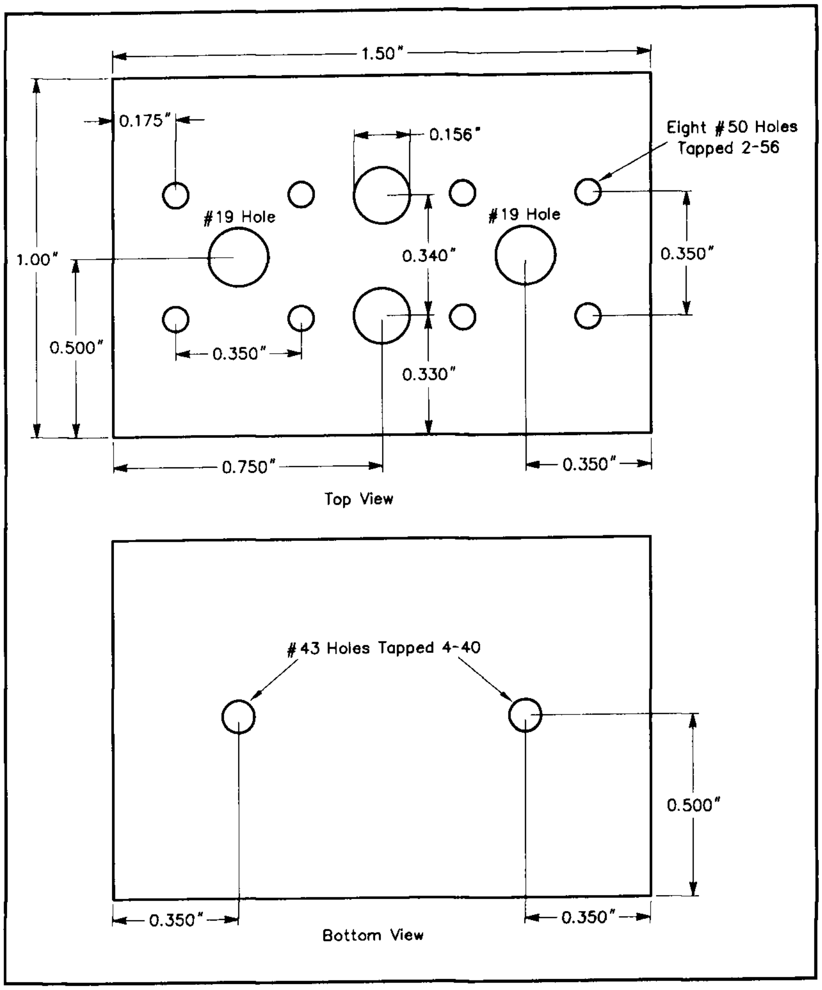

Fig 2—WR-90 waveguide drilling template.

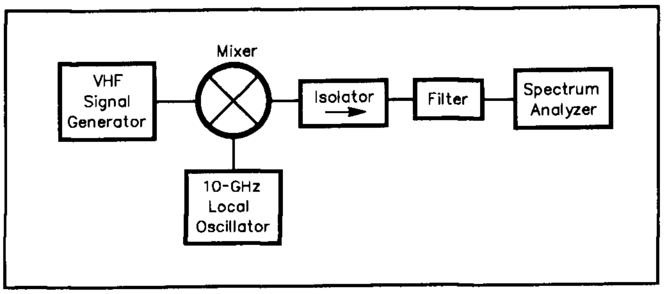

Fig 3—Diagram of a filter alignment test fixture.

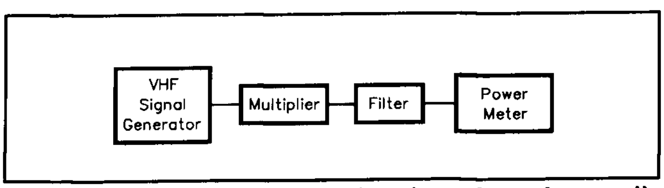

Fig 4—A simpler test fixture that doesn't work as well.

probes long and add shim stock to get the 100-mil starting lengths. The narrower bandwidth version with 3/16-inch posts required shorter probes—about 80 mils long.

Once you have the SMA connectors prepared, you can cut and drill the WR-90 waveguide. I filed the cut waveguide edges smooth. If you have 2-hole flange connectors, you need just half of the 2-56 mounting holes shown in the drilling guide. When drilling the holes for the posts, use several drill sizes to get up to the desired hole size. This will help to ensure a tight-fitting round hole that matches the tubing.

After deburring the holes and polishing up the waveguide, I soldered the posts into place. If you do a good job of drilling tight-fitting holes, it shouldn't be difficult to solder the end plates into place without having the posts unsolder themselves, even if you use a propane torch. I use a C-clamp to hold the end plates in place during soldering. I used ordinary 60/40 rosin-core solder.

To tune up the filter, I used a10-GHz mixer/local oscillator to up-convert a signal generator to provide a suitable signal source, as shown in Fig 3. I prefer this technique because I have a spectrum analyzer available to sort out the various mixing products. It is a considerable improvement over the simpler setup shown in Fig 4 using a frequency multiplier and a power meter. The frequency multipliers that I have built are rather frequency sensitive, requiring a calibration plot. Because most filters reflect rather than absorb unwanted signals, the reflections often disturb the operation of mixers and multipliers. The isolator works well with a mixer since all the big signals are near the same frequency. In contrast, a multiplier may have strong signals distributed over a relatively wide frequency range. A resistive attenuator may work better with a multiplier than an isolator designed to work over just an octave.

I estimate that the tuning screws were inserted 0.10 inches into the cavities. Obviously, the exact tuning will vary due to construction tolerances. Adjust the length of the probes and the tuning for minimum insertion loss.

Fig 5 shows a dual-mixer circuit board designed for use at 10 GHz.

It uses 15-mil 5880 Duroid (ε_r=2.2). It's a slight improvement over the one in the June 1993 *QST*, since it uses less board area. This article explains how to tune these mixers for best performance.

A brass frame is soldered around the mixer board using 0.5 × 0.025-inch brass strips. The strips are drilled and tapped to hold SMA connectors. I use 2-hole flange connectors to offset the center pins of the IF connectors so they

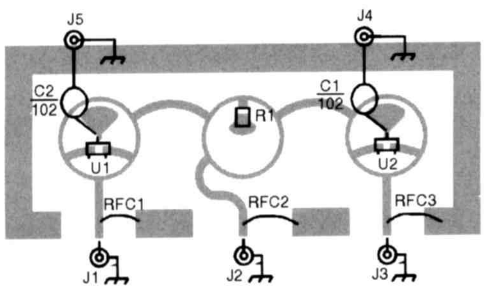

Fig 5—Dual 10-GHz mixer on 5880 Duroid.
C1, C2—0.001 μF, 50 V ceramic disc capacitors.
J1-5, SMA panel-mount female jacks.
R1—51 Ω 1/10-W chip resistor.
RFC1-3—0.21-inch #28 wires.
U1, U2—HSMS 8202 diode pair.

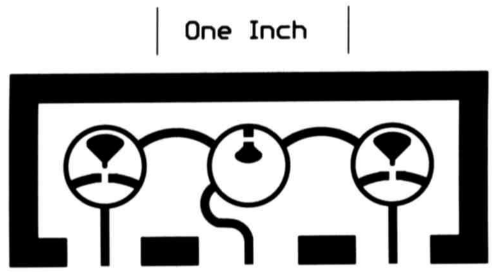

Fig 6—Full-size etching template for the microstripline mixer.

clear the ground conductors. I also tap holes on the two opposing strips without connectors to hold the aluminum cover. Fortunately, I have access to a metal brake to bend the cover. The best way to locate the holes is to install the cover and drill #50 holes through the cover and the frame. Then the cover holes are enlarged with a #44 drill to pass the 2-56 screws.

Note

[1]Bailey, Kirk, N7CCB, Larkin, Robert, W7PUA, Oliver, Gary, WA7SHI, "TNT for 10 GHz," *Proceedings of the Microwave Update 1988.*

2 W 10 GHz Amplifier

By Ken Schofield, W1RIL

(From *23rd Eastern VHF/UHF Conference Proceedings*)

The availability of the Microelectronics Technology, Inc Ku band power amplifiers on the surplus market has resulted in a gold mine of parts for 10 GHz operators. This white box not only contains 1 or 2 W (sometimes 1 and 2) IMFET devices but also the V_{dd} and V_g power supplies needed to run them. The 2 W device makes an ideal "afterburner" for the 1 W Qualcom strip available from the West Coast.[2] The box is massive and makes an ideal heat sink for the unit.

Two internal amplifier boxes, complete with covers, contain ceramic substrate gold microstrip. This stuff is so small it requires a jeweler's loupe to even see it. Modifying this gold microstrip is not a task to be undertaken on the amateur workbench! I decided to strip the box, save the screws and other useful parts and make a new board to fit the box—one that I could see, at least partially, without the aid of the jeweler's loupe. See Figure 3.

The new board is mounted in a cut down section of the original amplifier box the cut end of which has been fitted with a new brass plate made from 0.060 stock. A brass carrier cut to the size and shape of the board is fashioned from 0.032 sheet stock. The board and carrier are screwed down to the box bottom using 0-80 ss cap screws previously removed from the box. Before this is done, however, the original device landings in the bottom of the box are removed by milling flush with the rest of the box bottom. You will find that many of the 0-80 screw holes are already in the box bottom. It might be a good idea to leave the hole locations off the board and carrier and custom fit the holes to the box as some boxes may have different hole locations than others. The new hole locations can be added as required.

A schematic of the amplifier is shown in Figure 1. A few comments about the RFCs are in order. These are 0.005-0.006 mil lines. Attempting to get these on the board by etching is an almost impossible task. Strip the insulation from a short length of #20-#22 Teflon-coated stranded wire.

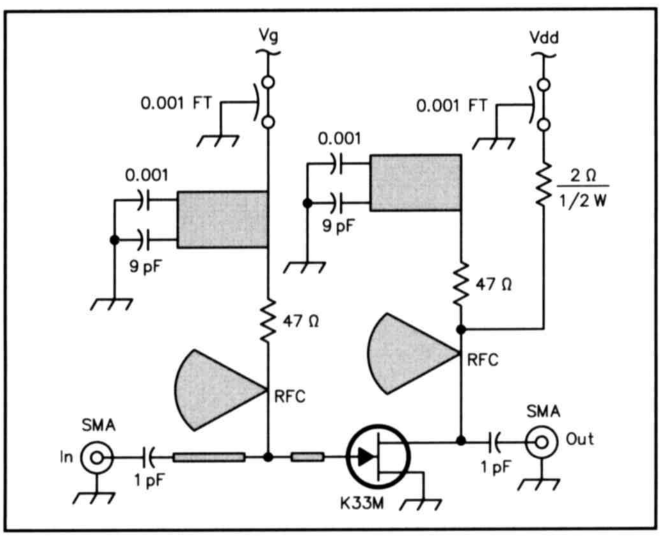

Figure 1—Schematic diagram of 2-W 10-GHz W1RIL amplifier using K33M IMFET.

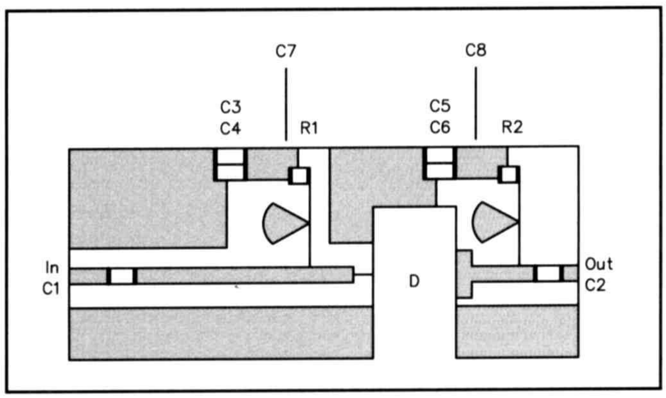

Figure 2—Parts layout of 2-W, 10-GHz W1RIL amplifier.
C1, C2—1-pF 50-mil chip capacitor.
C3, C5—0.001 chip capacitor.
C4, C6—9-pF chip capcacitor.
C7, C8—0.001 FT locations.
D—K33M IMFET.
R1, R2—47-ohm chip resistor.

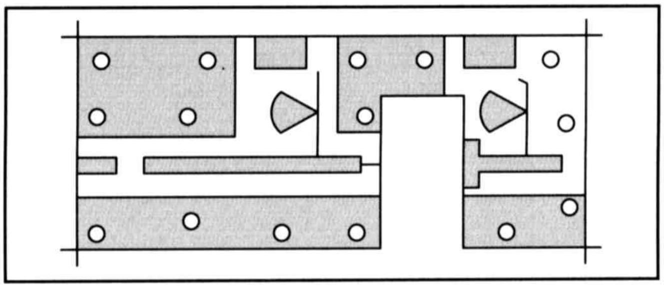

Figure 3—W1RIL circuit board to fit modified Microelectronics Tech Inc amplifier strip.
Board 1.6" × 0.75"
Mtl: Rogers 5880 Duroid, 0.015 mil
er 2.2
50-ohm line = 0.046 wide.

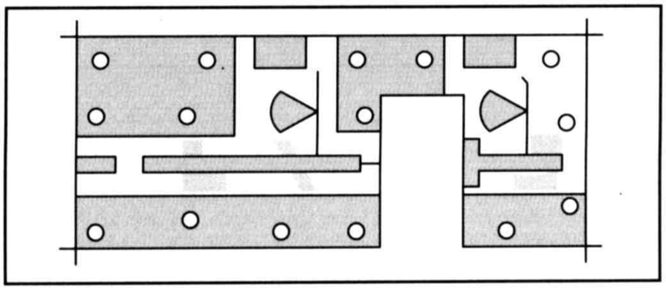

Figure 4—Circuit layout after tuning. Stubs shown in light gray.

Remove one strand of the silver plated wire from the twisted bundle and stretch it slightly to straighten. Solder to the 50 ohm lines and to resistors R1 and R2 keeping the wires straight and flush with the surface of the board. Also solder these lines to the apex end of the quarter circle bypass/decoupling capacitors. These capacitors have a radius of 0.207 and have smooth edges (despite what my computer shows!).

Tuning the amplifier is done by adding stripline pieces—see Figure 4. The gray areas on the 50 ohm line were added to tune amplifier for maximum output into a 50 ohm load. Please note the step in the line at the device input. This is actually a taper (my computer won't do that either!). The line is tapered to the width of the device input gate. Taper the line from 0.046 to 0.022 over 1/2 the distance from the RFC to the device input.

The amplifier gives the following results:

+28 dBm (640 mW) input = +33.4 dBm (2.2 W) output

V_{dd} = 7.48 @ 720 mA. V_g = −1.28

I would like to thank Bruce, N2LIV, for supplying the Tfe board for this project and Don, WB1FKF, for figuring out the power supply capabilities. Without their help I'm sure this project would not have come to fruition.

Notes

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