TRANSVERTERS

Building a 6-cm Transverter

By Zack Lau, W1VT

(From *QEX*, June 1996)

Here are some ideas and circuits you can use to build your high-performance 6-cm transverter. I'll describe my latest mixer and band-pass filter designs and provide references for building the entire transverter. I'll also discuss my approach to integrating everything into a useable system.

5760-MHz Mixer

The mixer for 6 cm is quite similar to the one I designed for 10 GHz, using a hybrid splitter and a pair of 6/4-wavelength ring mixers etched on Rogers 15-mil 5880 Duroid. 6 cm is perhaps the ideal band for using these mixers. The size is just about right—on 3 cm the rings are a little too tiny and on 9 cm they are getting a bit too big. More important, use of a relatively low 144-MHz IF makes it easy to design the hybrid ring for good performance at both the signal and oscillator frequencies. I published a design for a 5616-MHz LO multiplier chain in the May 1993 issue of *QEX* that will easily drive these diode mixers. You feed in the output of a 561.6-MHz no-tune LO and the chain puts out +7 or +15 dBm, depending on which output option you choose. I typically use +9 or +10 dBm to drive a pair of mixers with HSMS-8202 diodes. Using a higher IF for better image rejection often makes it necessary to compromise between good LO rejection and low conversion loss, since the 6/4-wavelength ring is frequency sensitive.

While the Hewlett-Packard HSMS-2822 diodes sort of work, I don't recommend them if you can get HSMS-8202 Ku-band diodes. I get 1.5 to 3.8 dB less loss with the Ku-band diodes. More important, it wasn't necessary to tune the mixer to get good performance. One mixer I built showed only 5.3 dB of conversion loss without tuning. The input 1-dB compression point was +1.3 dBm. If you have a spectrum analyzer, one way to evaluate the diodes is to compare the upper and lower mixing sidebands of an untuned and unfiltered mixer. With good diodes, the two should be nearly identical, while diodes with excess stray reactances may result in significant differences of a dB or two in conversion loss. Of course, if you don't mind tuning the mixer with bits of copper foil, the 2822s are useable even at 10 GHz. I'd use + 13 to +15 dBm of LO drive with the 2822s.

The radial stubs for the mixer and splitter are of different sizes. I made the one for the splitter smaller to compensate for the lead inductance of real resistors. At 6 cm, even tiny chip resistors can have a significant amount of stray inductance.

It may be worthwhile to add quarter-wave RF chokes between the splitter outputs and ground. 400 mils of 30-gauge wire should be about right for 5616 MHz. This will isolate the mixers from each other at the IF. Of course, the lack of isolation could make testing easier, since you could use the transmit IF signal to tune up the receive filter with a spectrum analyzer. On the other hand, noise from the transmit IF circuitry could prevent your receiver from obtaining the expected low noise figure. In one 10-GHz transverter I built, this noise doubled the NF from 1 to 2 dB.

Mixer Construction

After etching the board on 15-mil 5880 Duroid, I trim the board with a shear into a rectangle, leaving copper foil out to the edges. This allows me to solder 0.50×0.025-inch brass strips around the mixer board to form a frame suitable for adding a cover. These strips are drilled and tapped for five SMA connectors. SMA connectors are overkill for the IF connection, but they result in a compact and RF-tight assembly. By using 2-hole flange connectors for the IF, you can offset the center pins so they clear the ground foils. I've also used Teflon feed-throughs, but these compromise the RF integrity of the assembly. It is possible to improve performance slightly by tuning the mixers with small pieces of copper foil, but this shouldn't be necessary. My May/June 1993 *QST* 10-GHz article discusses mixer tuning.

5760-MHz Band-Pass Filter

The low IF used by amateurs makes filtering a challenge. Often, amateurs resort to waveguide filters to generate a clean signal. This works, but 6-cm waveguide is a little big for my taste. (Plus, it isn't the most commonly found

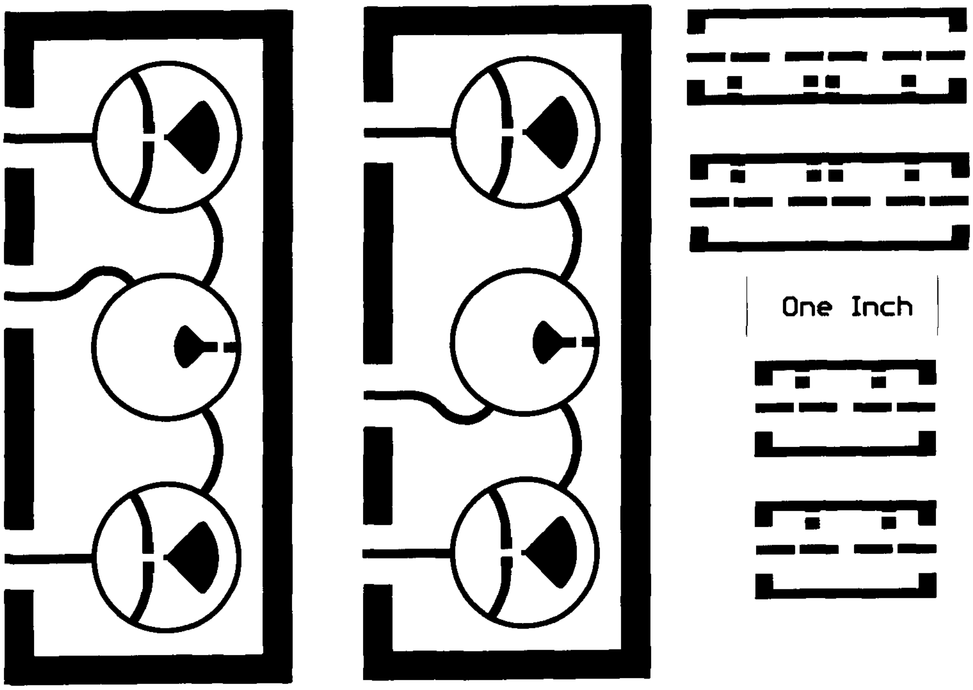


Fig 1—Etching pattern of mixer and MMIC amplifier boards. Board material is 15-mil Rogers RT/Duroid 5880 with a dielectric constant of 2.2.

stuff, particularly at New England flea markets. I think it's around, but who wants to cart around heavy metal objects that few people want?) A pipe-cap filter is pretty easy, but as some amateurs have discovered, a single pipe-cap filter gives marginal performance, especially if you are fortunate enough to find a surplus amplifier capable of generating significant power.

By carefully evaluating the plots by Kent Britain in the 1988 *Microwave Update* and making an educated guess using my knowledge of filters, I came up with a two-resonator filter that seems rather easy to duplicate—if you have the proper equipment to tune it up. I tried improving on my initial guess by varying the probes, and small changes weren't too critical, particularly when compared to trying to adjust a single resonator design for lots of spurious attenuation. The design shown in Fig 1 has a bandwidth of 47 MHz with 3 dB of loss. Using a 145-MHz IF, unwanted mixing products are at least 43 dB down. This should be adequate if you are using low-side injection with typical surplus amplifiers.

Like most of you, I don't have a network analyzer to use to sweep the filters. Instead, I used a mixer/LO to upconvert a VHF/UHF signal generator. By terminating the mixer with an isolator, I get measurements that seem to make sense. I've tried in the past to make measurements with a multiplier/filter driven by the signal generator, but this required much more work and yielded rather fuzzy results. With the mixer, I was able to cover 400 MHz around 5760 with less than a dB of variation in spectrum analyzer response; it wasn't even necessary to use the correction factor needed with the multiplier setup.

The advantage of using more resonators is the resulting steeper skirts that better attenuate the unwanted mixing products—I found the Q of the resonators I built wasn't high enough to clean up the signal with just a single resonator. Using corrosion-resistant nickel-plated brass screws is part of the reason, but I don't have a source for small quantities of silver-plated screws of the right size. Brass and nickel are significantly less conductive than copper and silver, so losses are noticeably higher if you use tuning screws made of these materials. By the time I narrowed the bandwidth of the filter enough to get lots of stop-band attenuation, the losses became excessive. In addition, with a really narrow filter you need to start worrying about effects like mechanical and temperature stability. It isn't hard to see how a filter with flexible Teflon board supporting some probe pins could be easily detuned.

Filter Construction Details

The filters use a pair of $^3/_4$-inch copper pipe caps. The one I measured actually had an inside diameter of 0.88 inches. They are tapped 8-32 at the top, though copper isn't the best material for holding threads. (You could solder brass nuts to the top, but then you would need longer screws than the 8-32 × $^5/_8$-inch screws I used.) I estimate that the screws extend about 0.35 inches into the cavity. Polishing up the inside of the caps is a good idea—a smooth surface helps raise the Q of the resonators.

I've found that tightening the lock nut pulls out the screw slightly, raising the resonant frequency. Conversely, tightening the screw against the copper threads of the pipe cap moves the screw in a little, lowering the resonant frequency. Done just right, you can end up with a precisely tuned resonator with the lock nut at just the right tension. This is how I normally tune up my waveguides and cavities. Alternately, you might consider a better tuning mechanism that doesn't require as much skill. I saw an interesting Gunn oscillator design that used dielectric rods attached to piston tuning assemblies.

The resonators are coupled together with a 2.0-inch length of 0.085-inch diameter semi-rigid coax. While I didn't experiment with different lengths, I recommend you stick with this length of coax. 250 mils of shield are removed from each end of the coax, so that the dielectric is exposed. The dielectric is left on to protect the center conductor. It probably makes sense to bend the coax first, then drill the coupling holes for the resonators through the mounting board. You can then vary the 0.75-inch spacing to suit the coax, as it isn't critical. This may be easier than precisely bending the coax to match the spacing. The 0.500-inch spacing between the probes in the resonators *should* be maintained, unless you want to experiment with a new design. I made the mounting board out of unetched 1/16-inch-thick double-sided circuit board. The poor thermal conductivity of the fiberglass is an asset—you can solder the pipe caps to the board without unsoldering the probes. Including the #33 mounting holes is a good idea even if you don't intend to use them immediately, as adding them later might take a bit more work.

5760-MHz Amplifiers

Transmit amplifier response plays a big part in the cleanliness of your microwave signal. Most surplus amplifiers useable for 5.76-GHz amateur work are designed for operation at 5.9 to 6.4 GHz. Not surprisingly, those using high-side local oscillators require more filtering, since the amplifiers are actually optimized for the LO frequency. This often isn't a problem with retuned or homebrew amplifiers, since the tuning typically results in a narrowband amplifier with rejection off the tuned frequency. But, many surplus amps do work reasonably well in the amateur band without any tuning, so many people do use them "as is."

The new Mini-Circuits ERA MMICs are just what we need to take the filtered signal and amplify it up to a level adequate for driving TWTAs or surplus amplifiers. The cascade of ERA-2/3 MMICs shown in Fig 2 has 26 dB of gain and +14 dBm of linear output. To prevent unwanted feedback, the amplifiers are only 0.50 inches wide. This results in a cutoff frequency of 11.8 GHz—high enough to offer significant attenuation over the bandwidth of the MMICs. A much wider enclosure invites waveguide propagation unless hard-to-find microwave absorber material is used. The simple MMIC circuitry makes this easy to accomplish. Amplifiers using GaAsFETs are often much wider, in order to accommodate the low-loss bias circuitry. I've also included the etching pattern for an amplifier using just a single MMIC, for applications that don't require the gain of two MMICs.

MMIC Amplifier Construction Details

After etching the board on 15-mil 5880 Duroid (ε_r=2.2), I trimmed the board to 0.50 × 1.45 inches. Next, I drilled holes for the power leads and carefully cleared away the copper ground plane around these holes with a large drill bit. Practicing with some scrap Duroid and different drill bits is highly recommended—you don't want the bit to "grab" and ruin the board. I use dial calipers for laying out the brass strips that form the frame around the board, just as with the mixer board. The frame also holds the connectors and feedthrough capacitor. I ended up mounting the capacitor on the side of the box since there was no room to mount it on the output end plate. There is a convention that the dc input connector ought to be mounted next to the output connector, but there are numerous exceptions to this rule.

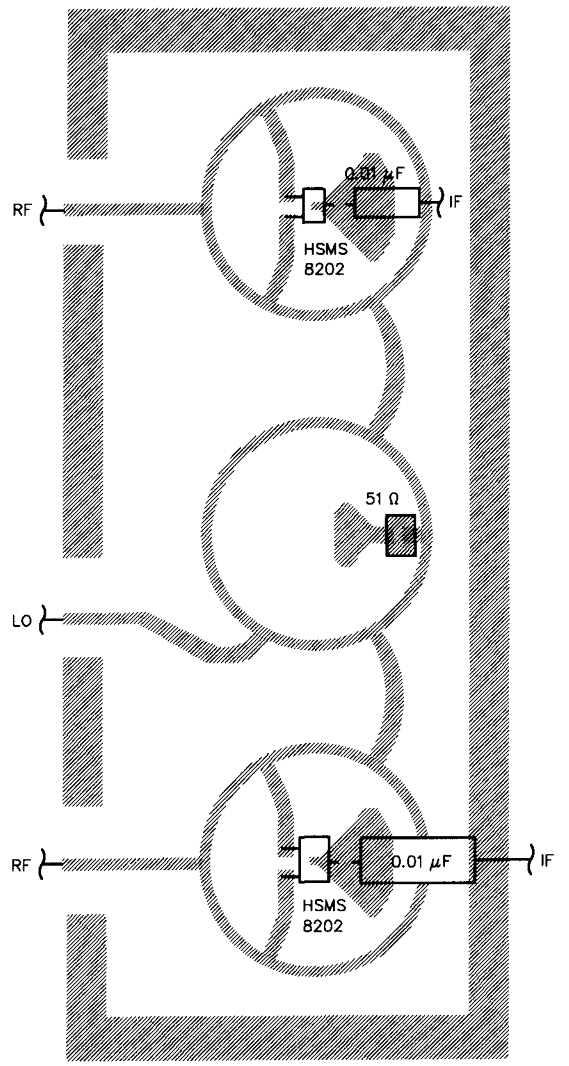


Fig 2—Parts-placement diagram for the mixer board.

I punch a 94-mil hole for each of the MMICs to sit in, then bend the grounding leads flat against the MMIC and stick them through one of the holes. After the input and output leads are flush against the board, I bend the grounding leads flush against the copper ground plane and solder them down. Then I attach the other surface-mounted parts. This differs from the usual practice of mounting the semiconductors last. I do it this way to ensure the best possible ground lead connections, which is critical for proper microwave performance. Finally, I wire up the regulator on the

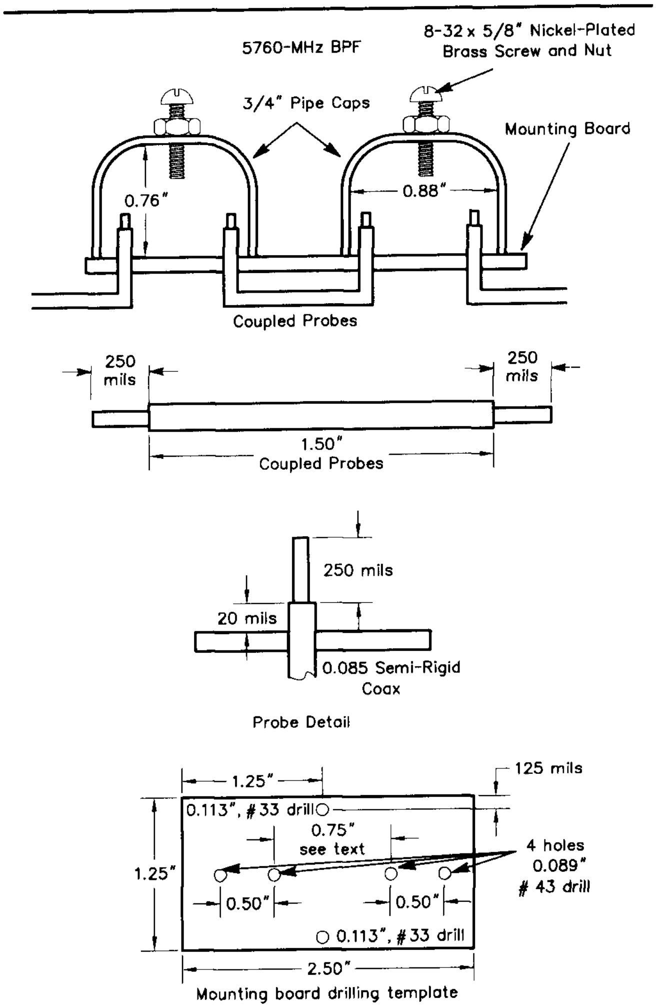


Fig 3—Coupled 5760-MHz pipe-cap band-pass filter. This filter provides a 47-MHz bandwidth with 3 dB of insertion loss.

ground plane side of the board.

There are two choices for a high-quality LNA on this band. The first is the design by Ranier, DJ9BV, in the March 1996 *QEX*. The second is the one I published in the September 1994 *QEX*. Ranier's has a slightly lower noise figure, but the 13 dB of typical gain isn't quite enough to overcome the noise generated by most 6-cm mixers, so a second stage will be required. My two-stage design with 22 dB of gain is about right for a terrestrial station. An EME station free of interference might effectively use as much as 30 dB of preamp gain ahead of the mixer to maintain a low system noise figure.

System Integration

Many of my transverters successfully use the Chip Angle sequencing circuit found in the *ARRL Handbook*. (Chapter 22, "T/R Time Delay Generator," by Chip Angle, N6CA.) As the diagram in Fig 6 shows, I avoid possible relay damage by sequencing both the RF drive to the transmit mixer and the dc power to the transmit amplifiers. While you can still end up hot-switching the relay if the relay sticks and then releases at the wrong time, I think this is so rare that I haven't bothered to design a suitable interlock. (I suppose you could sense the dc continuity of the relay and use this to indicate relay closure.) The September 1995 *QEX* has schematics that show the sequencer and IF circuitry in more detail.

I've found that this technique works just fine with semi-break-in transceivers since the transverter doesn't really care what the transceiver is doing (except for the dc control signal, obviously). Even if the transceiver is transmitting when the converter is receiving, everything is still operating acceptably—the 14-dB pad and switching-diode loss protects the receive MMIC from excessive RF. Thus, it even makes sense to use a center-off toggle switch to provide a manual

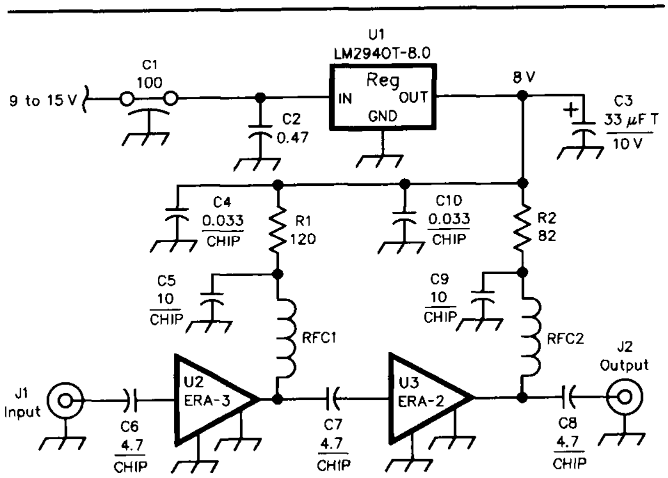


Fig 4—Schematic diagram for the MMIC amplifier.
RFC1, RFC2—3 turns no. 28 enam wire closewound, 0.062-inch inside diameter.
U1—National 2940T-8.0 low-dropout regulator.
U2—ERA-3 Mini-Circuits MMIC.
U3—ERA-2 Mini-Circuits MMIC

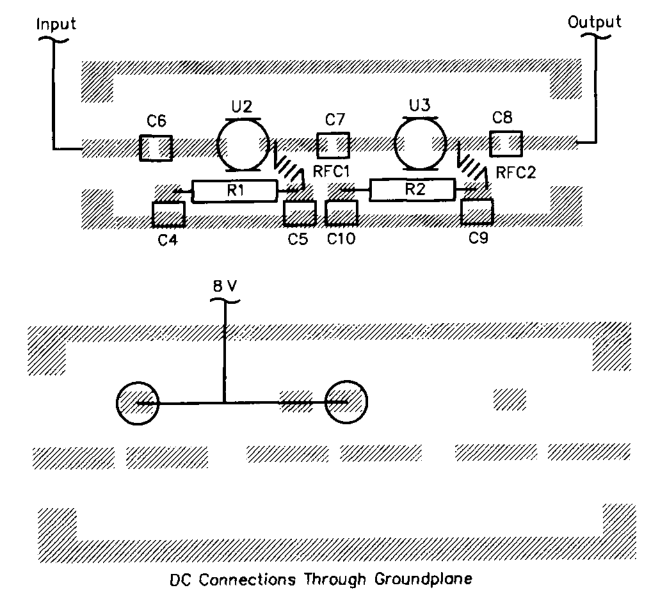


Fig 5—Parts-placement diagram for the MMIC amplifier.

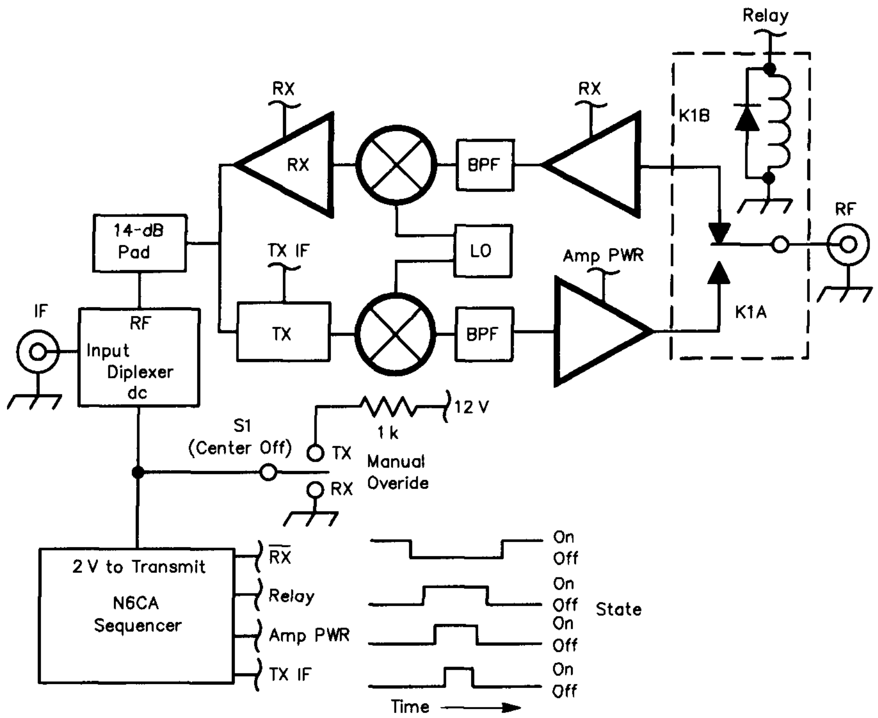


Fig 6—Using the N6CA sequencer board with a transverter and T/R relay.

override. This allows you to plug in any low-power IF radio and manually force the transverter into the proper state.

Those of us who travel 100 miles from home to operate from a tall mountain appreciate this feature. This is also handy for testing the transverter with test equipment, though there is a potential for an unexpected problem. Some signal generators cut back the RF if they see dc voltage on the output connector, so you might think there is a problem with the transmit converter when there really isn't; the transmit converter isn't putting out power simply because it isn't seeing much drive.

This scheme works well with low-power solid-state amplifiers because they are easy to turn on and off with power FETs. You can buy P-channel FETs with relatively low on-resistance—with even better devices appearing as time goes on. A simpler approach is to use the on/off control pin of the National LM2941 adjustable regulator, which offers a low drop-out voltage and delivers an amp. I'm not sure how well it will work with TWTAs, which often have warm up times and can't be turned on and off instantly, at least with inexpensive techniques.

Much less critical is the receive sequencing. Usually, the delay is so long that it doesn't matter whether the receive converter is turned on before or after the transmitter! In checking my previous work, it looks like I've done it successfully both ways. Turning the receive converter off first makes sense if you want to conserve as much current as possible, a consideration when running from batteries. I've not done any testing to see whether preamps are more resistant to high levels of RF with or without power applied, but the answer may determine the best approach with marginal relay isolation.

Connecting the Blocks

The coax and waveguide techniques that work well at 10 GHz work even better at 6 GHz, so most people don't have any problems. However, building a 10-GHz station isn't really a prerequisite to getting on 10 GHz, though almost everyone on 6 GHz also has 10 GHz. Finding 6-GHz surplus equipment is a lot easier than finding stuff suitable for narrowband 10 GHz, though sources do seem to be drying up.

At 6 GHz, you can pretty much toss out the idea of short pigtails; they just don't work with commonly found coaxial cables. You might be able to make the technique work with tiny 0.035 or 0.047-inch diameter semi-rigid coax, but I've never seen the stuff available inexpensively. By soldering the shield directly to the ground plane, you can get very short connections. The solid shield also helps, since you don't have to worry as much about stray wires shorting out the cable. Even at 2.3 GHz, pigtails seem to work only for low-power circuitry; I've had little success getting them to work above plastic MMIC power levels.

Normally, I connect all the assemblies together with SMA connectors and use N connectors for the antenna hardware. Testing cables and adapters at RF is a good idea—there is cheap hardware around that won't work well at this frequency. I've joked that one particularly cheap adapter could be used as an image stripping filter, due to its rather pronounced notch response. While I've left the BNC con-

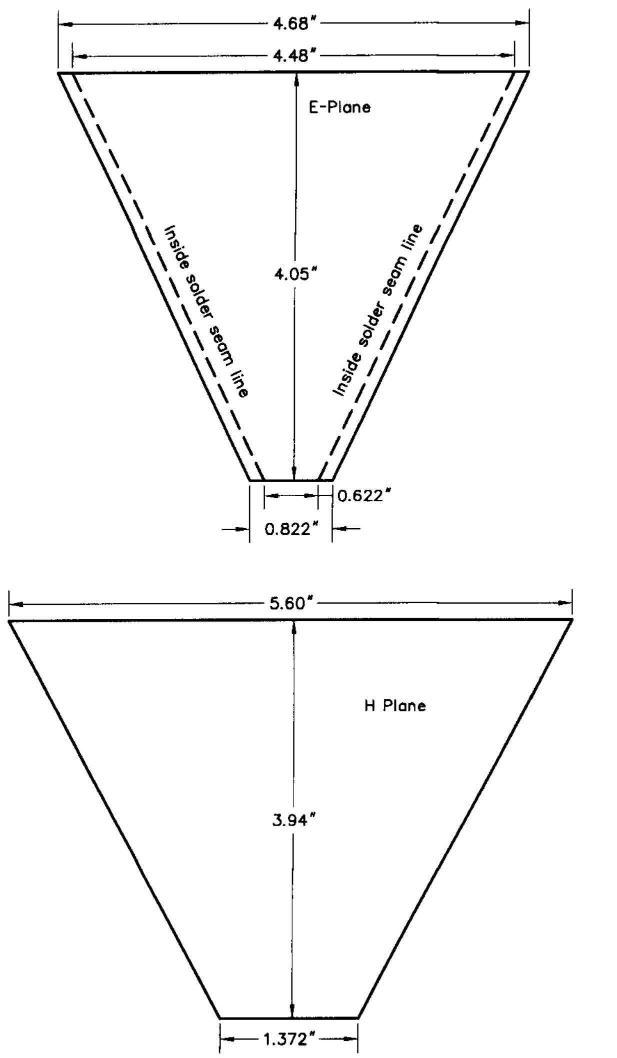


Fig 7—15.8-dBi 5760-MHz horn dimensions. The horn solders to WR-137 waveguide. The dotted line marks the inside solder seam.

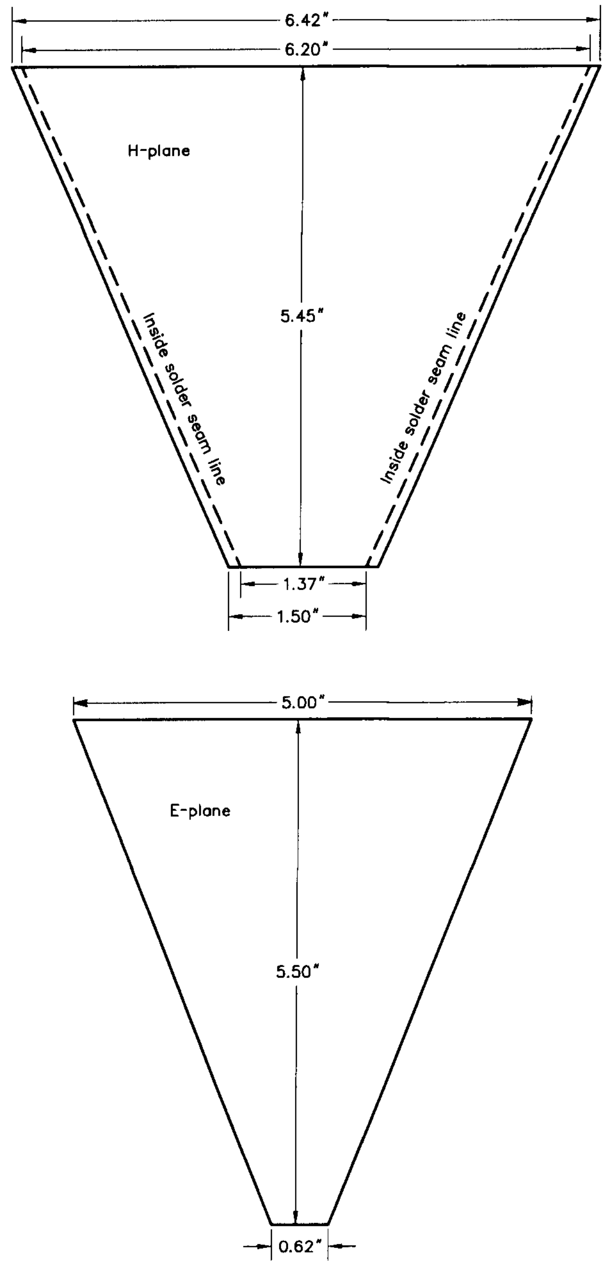


Fig 8—17-dBi 5760-MHz horn dimensions. The horn solders to WR-137 waveguide. The dotted line marks the inside solder seam.

nectors on our spectrum analyzer when doing quick checks, I don't recommend using them at 6 GHz.

At 6 GHz, your coaxial cable can actually be too big. If it supports more than one propagation mode, you can have signals propagated by different modes canceling each other out, resulting in extremely high losses. Andrew Corp suggests 5.0 GHz as the maximum frequency for LDF5-50A 7/8-inch Heliax. Of course, if you sweep the cable, you'll probably find frequencies above 5 GHz at which the cable works just fine. This is why many people are able to use 1/2-inch Heliax at 10.368 GHz: the frequency is in one of the clear windows above the cable's single-mode limit. I've had no trouble using either RG-213/U or 1/2-inch Superflex on 6 cm, although RG-213/U is rather lossy.

Horn Antennas

I included the simple antennas of Figs 7 and 8 to show an easy technique for mounting a small horn to a mast. Many textbooks and articles aren't clear how you accomplish this task. I use pieces of waveguide that come with unusual flanges. The flanges provide rugged attachment points for screws. To me, this is a lot easier than trying to fabricate some sort of bracket or welding the horn to a suitable mounting plate. This also works well for slot antennas. Besides, I have no other use for the mounting flanges. The horns have predicted gains of 15.8 and 17.0 dBi, respectively.

I made the phase centers equal by making the horn H-plane width smaller while keeping the horn length and E-plane widths constant. Having them equal is useful if you intend to feed a lens. While you could make the E-plane width wider, you actually lose a little bit of gain, even though the horn is bigger. You could say that the transition to free space is occurring too quickly for maximum gain. This is why horn gains above 23 dBi are rare—dish antennas become much more practical than a very long horn. With the 17 dBi horn, I also made the horn a little longer.

Horn Construction

I made the smaller horn out of 1/16-inch unetched double-sided fiberglass circuit board. For light weight, I made the 17-dBi horn of thinner 0.025-inch G-10 circuit board, though it's not as strong. For durability, I find it important to tape the joints with copper foil. Otherwise, people borrowing the horns return them with broken solder joints. To protect the copper from corrosion, I painted them with clear acrylic spray paint.

Since most people use coax on this band, Fig 9 shows an N-to-WR-137 transition. I used an Amphenol 82-97 UG-58A/U connector. These connectors have a center contact that press fits into the Teflon. A different connector may require a bit of experimentation, since the center contact forms part of the probe. Bare #12 copper wire is easily obtained by stripping ordinary house wire sold in hardware stores. I used a hacksaw to cut a pair of slots for the brass shorting plate. The 0.256-inch dimension is from the center of the probe to the inside surface of the shorting plate. I slid the snugly fitting 1.5 × 0.622 × 0.032-inch shorting plate through the slots and soldered it in place with a propane torch.

I soldered the horn directly to the waveguide with copper tape. A soldering iron is useful for tacking the tape into position. Then I used a propane torch to do the final soldering, since the waveguide needs quite a bit of heat to properly melt the solder. The small horn and large horns measured 14 dB and 30 dB return loss, respectively. I consider 14 dB adequate, though a purist would add tuning screws or vary the probe length for a better impedance match. People normally put screws on the centerline of the waveguide, where they have the most effect. Usually, two or three screws placed a quarter or eighth of a waveguide wavelength apart will match nearly any load. I wouldn't worry too much about not having a set-up to measure SWR—I made a couple of 200-mile 10-GHz contacts before I finally got a precision directional coupler and reduced the SWR of my dish feed below 2:1.

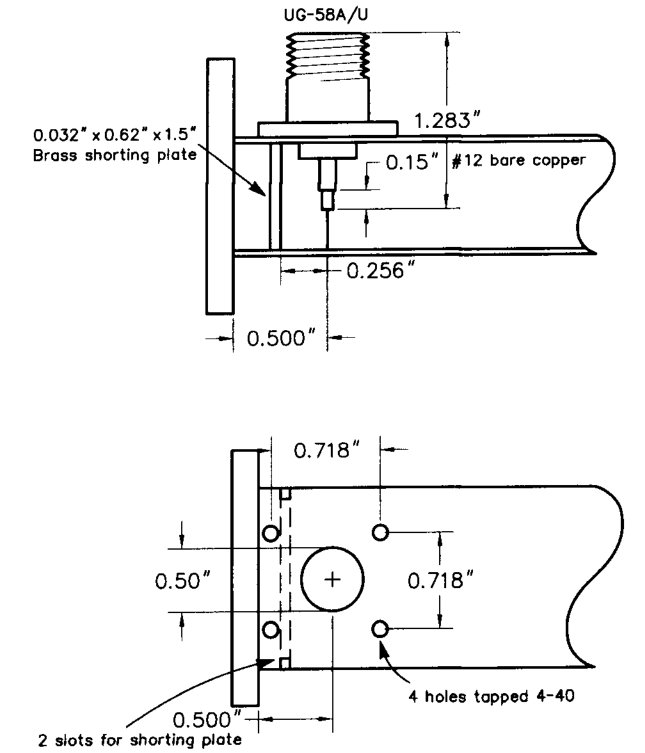


Fig 9—WR-137-to-N connector transition.

3456-MHz Transverter

By Zack Lau, W1VT

(From *QEX*, September 1996)

Jim Davey's 3456 no-tune transverter was an amazing piece of work—it was the first of the no-tunes and did a remarkable job of establishing the concept. Like many ground-breaking designs, it had a number of minor problems. Here is my approach to fixing them.

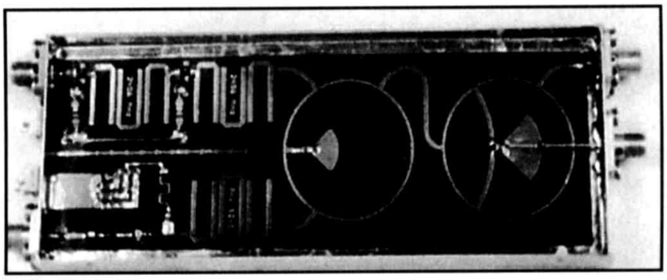
The most serious problem is board radiation that unbalances the mixers and degrades spectral purity. This is caused by the combination of a low dielectric constant and a relatively thick circuit board. Unfortunately, high dielectric constant boards have about twice as much loss, a significant drawback since the transverter requires high-Q band-pass filters. As a result, I chose a much thinner, 15-mil board. This allows an aluminum cover to be placed over the circuit with negligible effects on circuit performance. You don't need absorptive rubber to shield this circuit. Microwave absorber material can be tough to find in small quantities.

Another drawback of the original design is the lack of voltage regulators. Performance was seriously degraded as the batteries ran down. By using three-terminal regulators, the circuits work just fine between 9 and 15 V. The LM2940T-8.0 is shown in Fig 1; it not only features a low drop-out voltage, but offers reverse polarity protection in case you hook up your batteries backwards! Beware, you can turn the regulator into an oscillator by substituting an inadequate value for C9. You need a sufficiently large amount of high-quality capacitance for stability. For details, consider downloading the data sheet from National's WWW page: **http://www.national.com/pf/LMLM2940.html**.

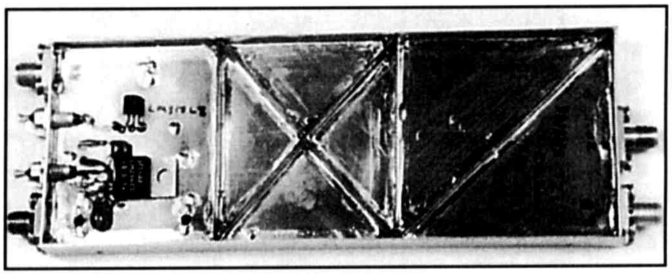
Finally, the MMICs used were a bit marginal, operating near their upper frequency limit. This version uses newer MMICs with significantly enhanced performance at 3.5 GHz. The new MMICs have so much gain that it made sense to revise the circuit topology. Instead of dual mixers, I chose to use a single mixer and a splitter. It is common practice when using a single mixer to use the same band-pass filter for transmit and receive. It is placed between the splitter and the mixer. Terminating the mixer with the splitter improves performance, while adding little extra complexity to the circuit. The situation would be different with pipe-cap or waveguide filters—then I'd have to spend nearly twice as much time fabricating the filters. In this situation, board space is saved because I don't need to place a small rectangle between two large circles.

There is also a subtle advantage for those of us using heat-transfer techniques for fabricating the boards. All the high-tolerance filters are concentrated in a relatively small area of the board. This makes it significantly easier to get an accurate reproduction. I noticed this when examining an LO board with low output (+4 dBm instead of +8 dBm). The output filter was stretched so much that the gaps between the resonators were a couple of mils wider than the design called for. This effectively raised the center frequency of the filters. As Table 1 indicates, you can still get useable performance with 3 dBm of LO drive. To assist you in fabricating the filters, the dimensions are shown in Fig 2.

The single mixer significantly reduces the circuit board area required. Only 16 square inches are needed for the LO multiplier and main transverter board, about 2/3 of the original. It also simplifies transmit/receive switching—the mixer can be hooked up directly to a +3 dBm VHF transceiver, like the Rick Campbell mini R2/T2/LM2.[1] The effect of the splitter loss on performance is negligible—the receiver still has



3456-MHz no-tune transceiver, 14 dBm output, 1.7 dB NF.



3456-MHz no-tune transceiver voltage regulators.

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

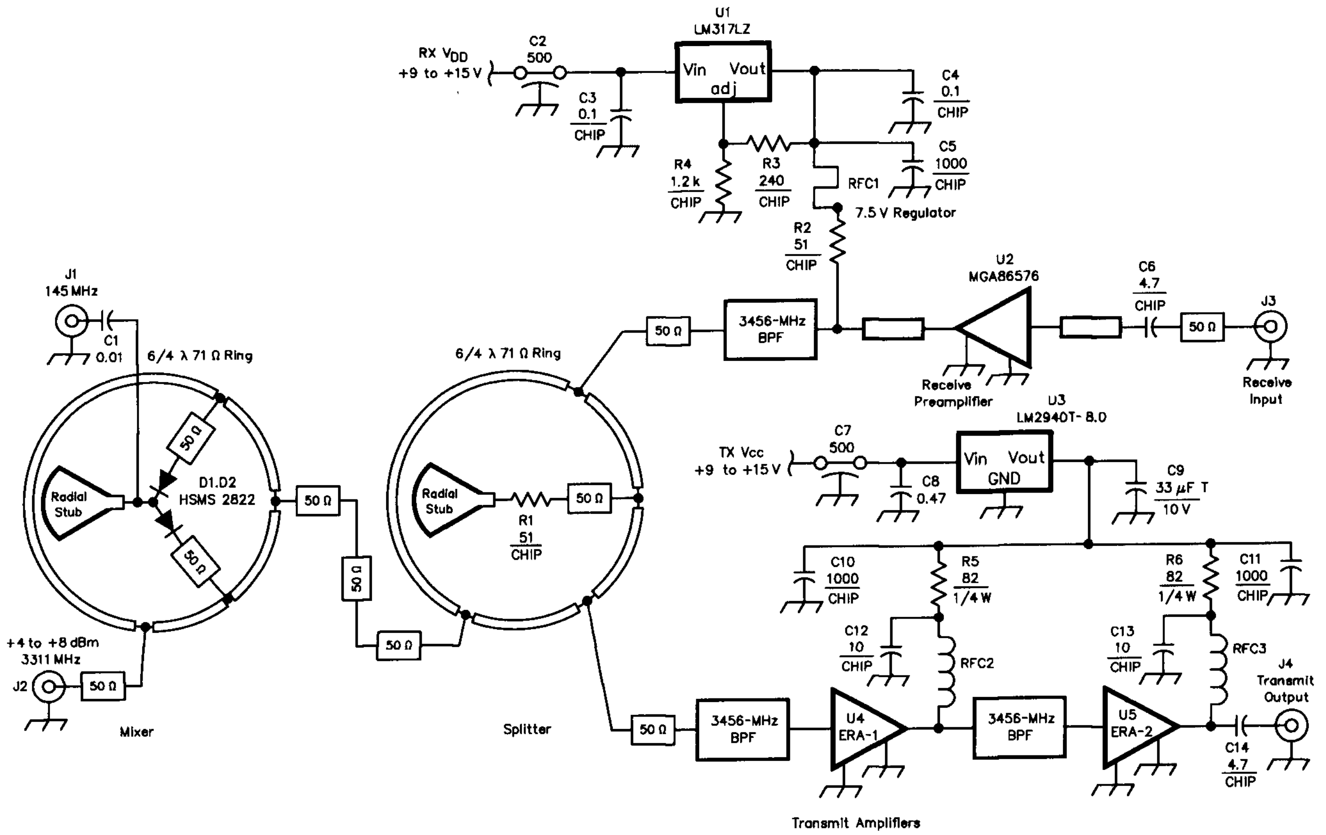


Fig 1—Schematic diagram of the 3456 transverter board.

C2, C7—Feedthrough capacitor; value not critical.
C6, C14—High-quality 4.7-pF chip capacitor like the ATC 100A. Not critical if you aren't worried about noise figure or power output.
C9—33-µF, 10-V tantalum. *The National Data Book* recommends a minimum of 22 µF.
D1, D2—HSMS 2822 packaged diode pair.
J1—Two-hole flange-mount SMA panel jack. Omni Spectra 2052-1652-02 works quite well.
J2, J3—Four-hole flange-mount SMA panel jack.
RFC1—Printed circuit board RF choke.
RFC2, RFC3—4 turns of no. 28 enameled wire close wound. 0.062-inch inside diameter.
U1—TO-92 case adjustable regulator.
U2—Hewlett-Packard MGA 86576 GaAs MMIC.
U3—National LM2940T-8.0 low-drop-out regulator.
U4—Mini-Circuits ERA-1 HBT MMIC.
U5—Mini-Circuits ERA-2 HBT MMIC.

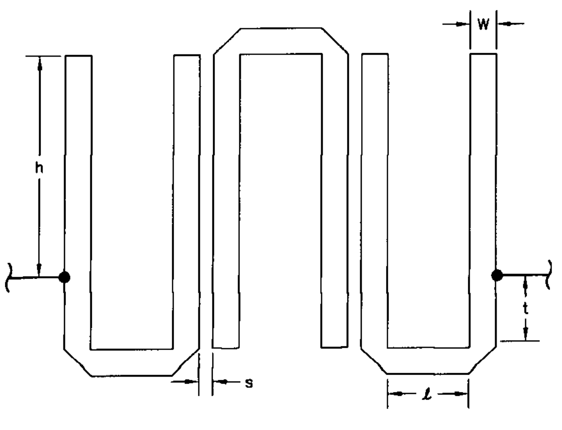


Fig 2—Dimensions of the 3312 and 3456-MHz band-pass filters on 15-mil 5880 Rogers Duroid.

		3456 MHz	*3312 MHz*
h	coupled line height (mils)	486	508
s	spacing between coupled lines (mils)	25*	25*
t	tap height (mils)	28	34
w	line width (mils)	50	50
ℓ	uncoupled length (mils)	150	150

***Modeled spacing, see text.**

a 1.7-dB NF and 14 dB of gain, while the transmitter has a 1-dB compression point of +14 dBm with 12 dB of gain. The decrease in gain on receive may actually be an advantage by reducing its susceptibility to mixer overload if a low-noise preamplifier is added. The output level is convenient for running surplus TWTAs or their solid-state replacements.

The new Mini-Circuits ERA-1 and ERA-2 have just the right amount of gain for the transmit amplifiers. On transmit, too much gain can be just as bad as too little gain. The more gain you have, the easier it is to make an oscillator. Feedback might help, but installing feedback networks deviates from the idea of a simple, reliable project with a minimum of parts. These MMICs use heterojunction-bipolar-transistor (HBT) technology. They do a good job of combining wide bandwidth with a relatively low supply voltage. At 50 mA each, they draw a fair amount of current

Table 1
Effect of local oscillator power on transverter performance

Transmitter		*Receiver*	
LO Power (dBm)	*Output Power (dBm)*	*NF (dB)*	*Gain (dB)*
3.2	13.27	1.79	15.11
3.8	13.33	1.78	15.15
4.9	13.40	1.77	15.28
5.6	13.42	1.79	15.30
6.0	13.37	1.79	15.34
6.8	13.37	1.76	15.34
7.2	13.40	1.78	15.33
7.7	13.38	1.77	15.35
7.9	13.38	1.77	15.36

to generate 20 to 40 mW of RF. I think this is a reasonable trade-off, considering the complexity of the alternatives. A discrete FET design would be more complex, requiring a lot more design work.

I haven't experimented with the new ERA-4 or ERA-5 MMICs to see if more output power can be obtained without modifying the board. I'm still waiting for the ones I ordered in late May '96. I don't recommend using the high-gain ERA-3 MMIC between the band-pass filters—it is quite likely to be unstable unless drastic measures are taken. It may be necessary to shield the printed circuit board filters from each other. You might also experiment with narrowing the "waveguide" enclosing the circuitry. You may significantly reduce the chance of unwanted waveguide propagation by installing a shield on the optional grounding strip in the center of the board. This shield would lower the cut-off frequency by a factor of two. The low-loss nature of waveguide is a significant disadvantage when attempting to build amplifiers—what better way to create an oscillator than to couple the input and output together with a low-loss transmission line?

Keep the parts close to the board to reduce their ability to launch signals into the waveguide. I laid out the board placing the RF chokes close to the edge of the waveguide, as opposed to the center. Objects in the center of the waveguide couple into the waveguide better than those close to the edges. This is why you typically put detector diodes in the center of the waveguide when you want to maximize the signal to the diodes.

Keeping to the idea of simplicity, the receive preamplifier is a single Hewlett-Packard MGA 86576 GaAs MMIC. It has about 24 dB of gain and a 1.6-dB noise figure. The NF

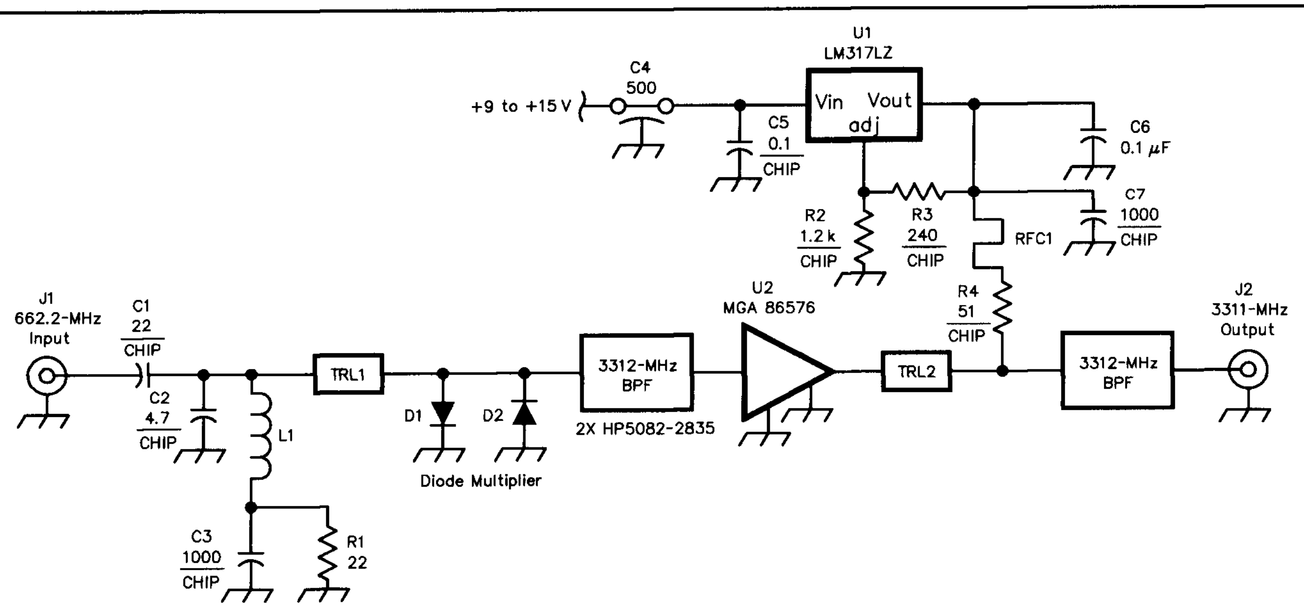


Fig 3—Schematic diagram of the 3311-MHz local oscillator multiplier.

C4—Feedthrough capacitor, value not critical.
D1, D2—Hewlett Packard 5082-2835 Schottky diodes.
L1—3 turns no. 28 enameled wire spaced 2 wire diameters. 0.089-inch inside diameter.
RFC1—Printed circuit board inductor.
U1—LM317LZ adjustable regulator.
U2—Hewlett-Packard MGA 86576 GaAs MMIC.

is degraded an additional 0.1 dB by the 10 dB of converter losses. The LM317L could be replaced by a 78L07 or 78L06 regulator, but these are harder to find and require a larger input bypass capacitor for stability. With the three MMICs I used, the highest supply voltage resulted in the best gain and noise-figure performance. I didn't do any testing past the recommended 7-V device voltage. Unit 2 still had a 1.9-dB NF and 13.65 dB of conversion gain with a device voltage of 4.82 V, so the device voltage isn't terribly critical.

The mixer and splitter both use a 180° hybrid. A good reference on these may be found in Chapter 6 of the *ARRL UHF/Microwave Experimenter's Manual.* The narrow bandwidth of the hybrid isn't a problem in this mixer application, due to the relatively low IF of 145 MHz. This is only 4% of the center frequency. The radial stub for the splitter is a little small, in an attempt to compensate for the stray inductance of the chip resistor.

The LO multiplier uses a diode multiplier, a pair of band-pass filters, and an MGA 86576 GaAs MMIC. FETs can be more efficient than diodes at frequency multiplication, but they tend to be more critical with regard to drive and tuning. The GaAs MMIC is a bit more expensive than a pair of the new ERA-3 MMICs, however, they produce just the right power level for driving a single mixer. An ERA-3 is more appropriate for driving a pair of diode mixers. Another advantage of the MGA 86576 is that it draws only 16 mA, compared to 35 mA for a single ERA-3. Actually, a pair of ERA-3s has too much gain, and running an ERA-1 and ERA-2 in cascade ups the current draw to 100 mA. This is six times as much as the GaAs MMIC draws.

The LO multiplier is designed to be used with a 662.2-MHz source. It can be easily modified to work with a more standard 552-MHz source by removing either one of the multiplier diodes, D1 or D2. As a 6× multiplier, +15 to +20 dBm of drive is needed. As a 5× multiplier, +13 to +20 dBm of drive can be used. This multiplier works much better with the correct number of diodes. If the stability of the LO amplifiers is marginal, it may be wise to add a 100-Ω resistor or resistive pad to the input of the multiplier. It isn't too difficult to envision cases where the constantly changing impedance of the diodes could cause problems.

I know that a lot of people are looking for practical ways to design the microstrip hairpin filters. The simple answer is there isn't any, at least for amateurs with little time or money. The programs that accurately simulate the discontinuities, such as the bends in the microstrip, still cost quite a bit of money. Trial and error, particularly with return loss measurements, is an effective way of designing complex filters, if you have the time. If you have a spectrum analyzer, I've found that upconverting a low-frequency signal generator with a mixer makes a good signal generator—if you add an isolator. The isolator significantly reduces the interaction between the filter and the mixer. If you aren't careful, some mixer/filter combinations can actually indicate that the filters have gain. Reflected signals can actually enhance the desired signal. As Jim Davey found out, return loss is a much more sensitive indicator of circuit performance than insertion loss.

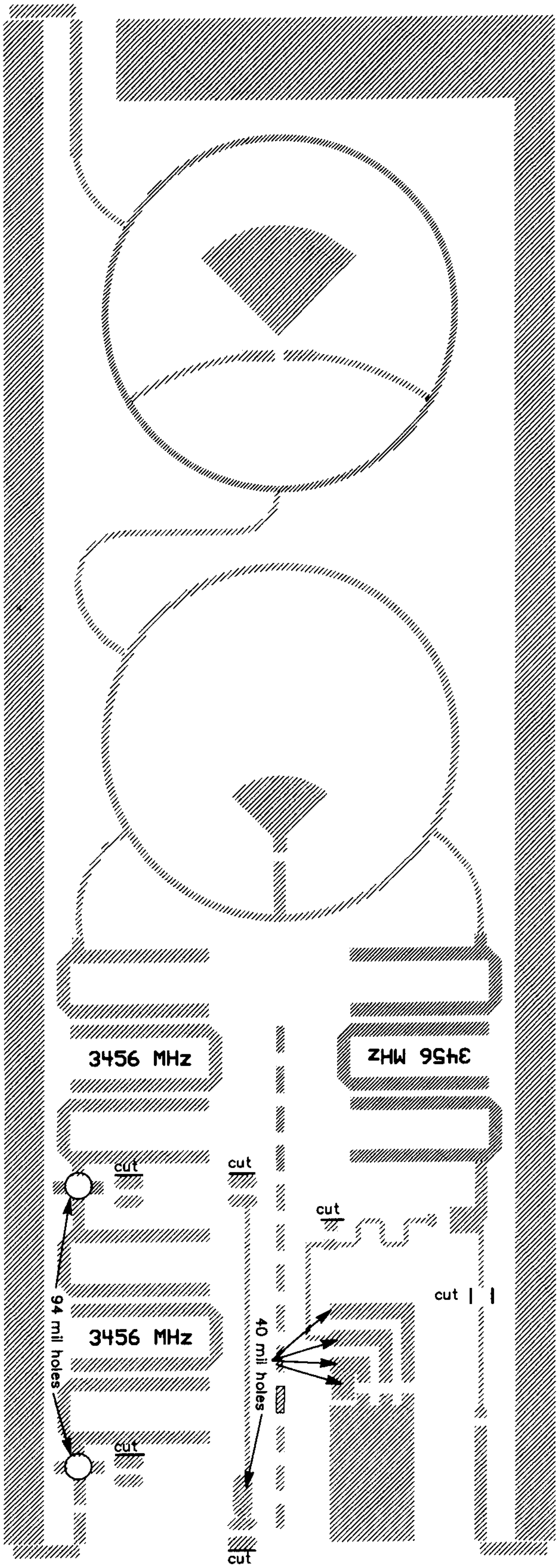


Fig 4—Cutting and drilling diagram for the transverter board.

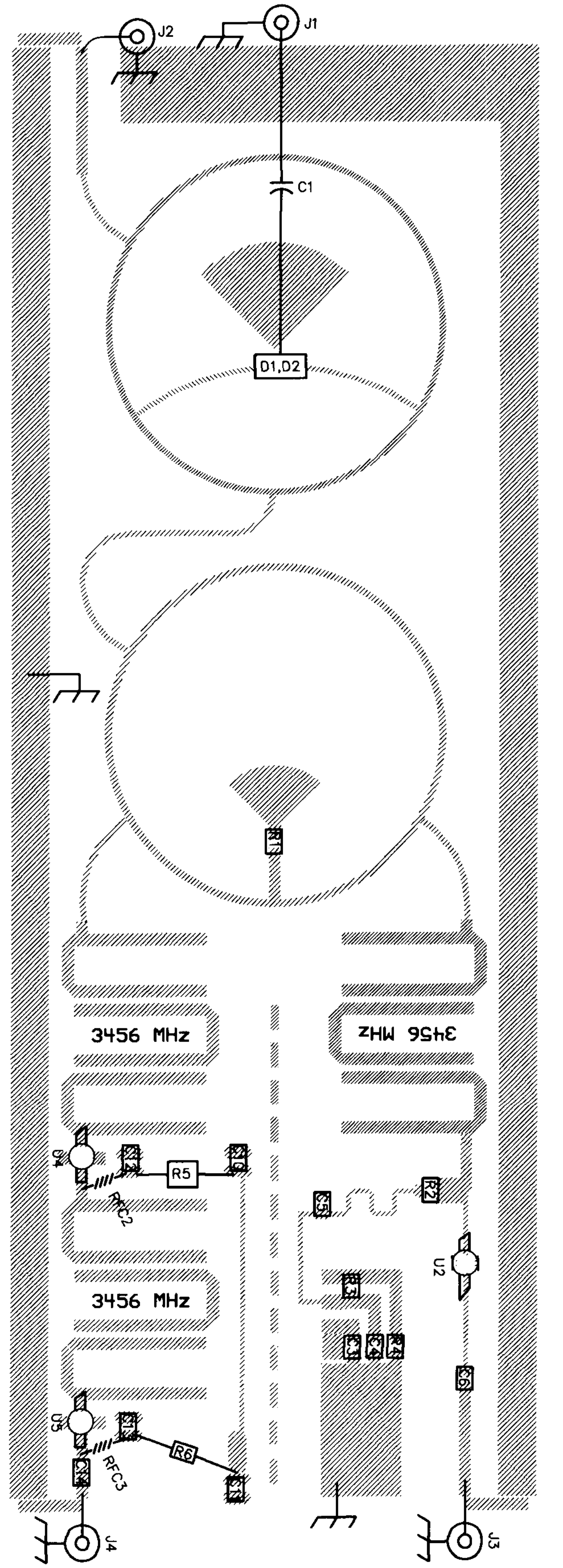


Fig 5—Trace side parts placement diagram for the transverter board.

Construction

Making these circuit boards is a stiff challenge—but some of us like challenges. Jim specified a tolerance of ±0.001 inches for the original boards. I doubt that I come that close with my etching techniques, but it's a target to shoot for. I was able to build three transverter boards, all with acceptable performance (Table 2). Mixing and LO spurs were at least 40 dB down. The second and third harmonics were only down 39 and 32 dB on one unit. This is no surprise, since there is no low-pass filtering of the output stage.

When comparing the dimensions of Fig 2 against the artwork, the careful examiner may notice that the spacing between coupled lines is actually 24 mils, as opposed to the 25 mils determined with the assistance of computer modeling. Since this was noted after several units were built and tested, I didn't bother revising the artwork. Due to the tight tolerances, I'd recommend you work with the Postscript files. If you can't download them off the 'Net from **http:// www. arrl.org/qexfiles**, I can supply the file for noncommercial purposes if you enclose a 3.5-inch disk and addressed return envelope with postage.

The boards are etched on Rogers RT/Duroid 5880 with a dielectric constant of 2.20, clad with 1 ounce rolled copper on two sides. The dielectric thickness is 0.015 inches.

When trimming the circuit board, don't forget to leave enough room for the SMA connectors. Normal square-flange connectors are 0.5 inches wide—thus the board needs to be at least 0.25 inches from the center of the 50-Ω microstrip traces to the edge. On the other hand, excessively widening the board increases the possibility of waveguide propagation, so you don't want to err too far in the other direction either. You can also use smaller SMA connectors—Digi-Key now advertises a line by Johnson Components (formerly EF Johnson Components). While a bit expensive, they supposedly work up to 26.5 GHz, as opposed to 18 GHz for standard connectors. They should be useful for 24-GHz work.

I used a hobby knife with a new no. 11 blade to cut the slits in the board for the grounding straps and MGA MMIC ground leads. The blade of the knife should just touch the outer edges of the pads marking where to cut the slits. I don't trim away any Teflon from the slits. Instead, I use a flat-bladed screwdriver to carefully close up the holes after the leads are passed through by reworking the remaining copper foil. The holes for the ERA MMICs are punched with a

Table 2
Test data for three converters

	Transmit		*Receive*	
Unit	*Power (dBm) (1 dB compression)*	*Gain (dB)*	*NF (dB)*	*Gain(dB) (compressed)*
1	14.40	12.9	1.78	15.34
2	14.10	12.6	1.75	14.95
3	14.83	12.1	1.64	12.94

Power was measured with an HP 8563E spectrum analyzer and confirmed with an HP 8481A/435B power meter. An HP 8970/346A was used to measure receive converter performance. Both an HP 8640B and a Marconi 2041 were used to generate the IF drive.

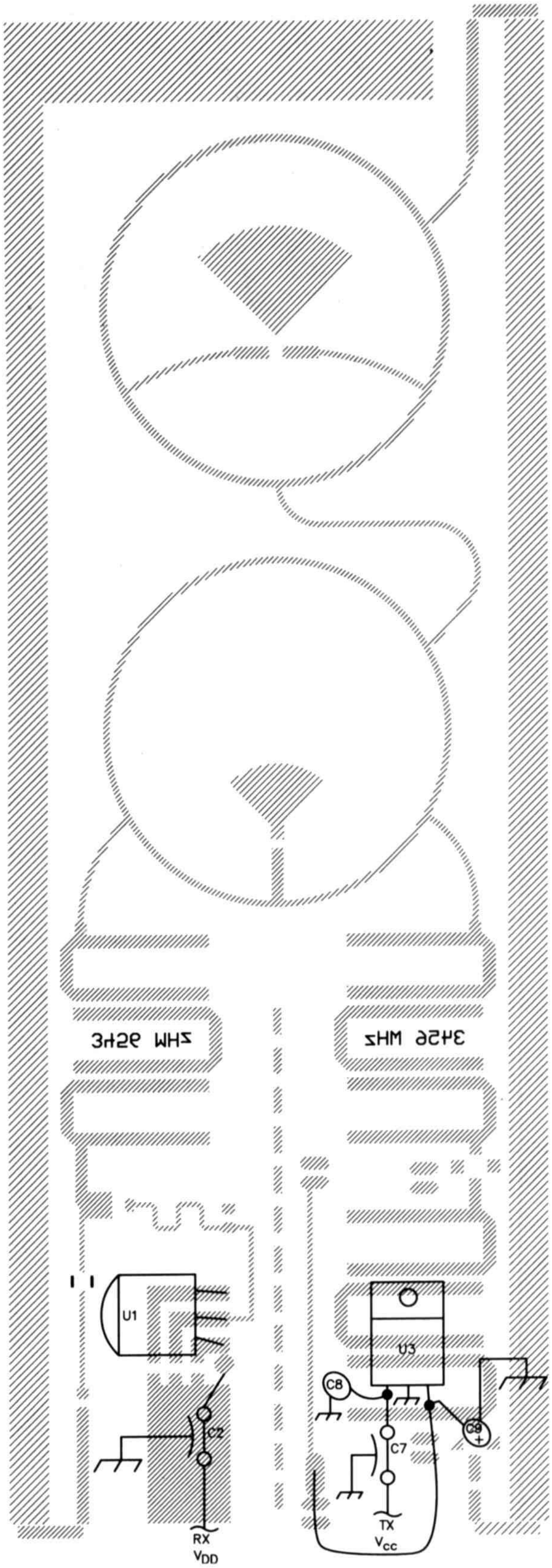
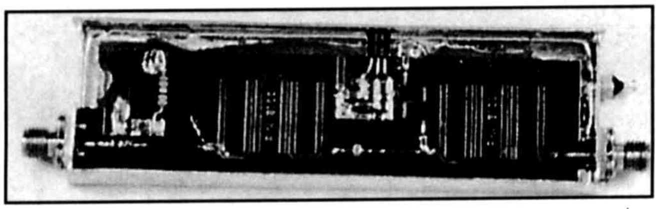


Fig 6—Ground-plane side parts placement diagram for the transverter board.



552 to 3312-MHz LO multiplier on 15-mil 5880 Duroid.

94-mil hole punch and then touched up with the hobby knife. I've found that drilling large clean holes in thin Teflon board can be difficult. After drilling the 40-mil holes for the power supply parts, I countersink the ground plane side by hand with a 1/4-inch drill bit.

Some drill bits bite too deeply into the board, so you might practice your techniques on a few scraps of Teflon board. If you work slowly and realize there is a problem, you can "save" the board by carving away the excess copper with a sharp knife.

Even with a frame made of 25 × 500-mil brass sheet stock around the edges, I felt the board could use a bit of stiffening. To stiffen the board, I soldered some thin strips of unetched double sided circuit board to the ground-plane side. The assembly was stiffest when the boards were slid against the metal frame and soldered it. Forming an X or two seems to markedly stiffen the board. Brass could be used, but the circuit board is easier to solder and has less flexibility. A stiff board is important because most chip parts can't flex along with the board—too much flexing and something will break.

A two-hole flange mount works well for input connector J1. The thinner connector can be raised slightly compared to the other three connectors. This makes it easy to connect the input capacitor, C1. Alternately, higher walls could be used to enclose the transverter board.

While there are pads for grounding the ERA MMICs with copper foil, I found it easier just to bend and solder the leads. The leads are inserted into the holes and then bent against the body of the MMIC, flattening them against the ground plane. I used copper foil with Unit 2—there wasn't a significant difference in performance compared to the other two units.

I first test the voltage regulators to make sure they are putting out the proper voltages before installing the resistors that supply power to the MMICs. The actual voltage at the MMICs can vary a bit—the GaAs MMIC can draw between 9 and 22 mA, so the voltage drop across the 51-Ω chip resistor can be anywhere from 0.46 to 1.1 V. I'm not surprised at this range—manufacturing repeatable bias points has always been a weak point of FET technology compared to bipolar. The device voltage for the ERA MMICs is supposed to be between 3.2 and 4.4 V, nominally 3.8 V.

The shared mixer ought to make troubleshooting easier. If nothing works, there is a problem in the LO/mixer, or in both the transmit and receive amplifiers. While the conversion gain isn't excessive, it ought to be enough to hear the increase in noise when you turn on the receive amplifier.

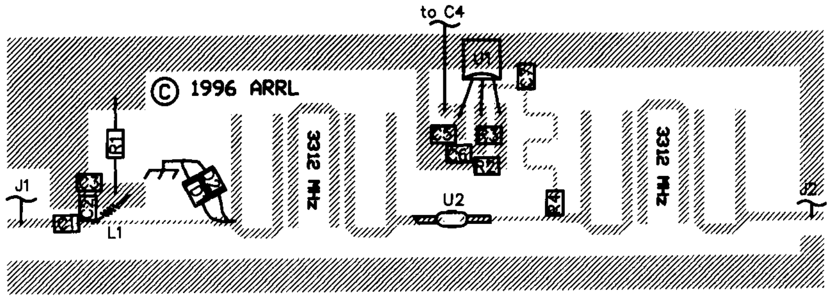


Fig 7—Parts placement diagram for the multiplier board.

Notes

[1]Jim Davey, WA8NLC "A No-Tune Transverter for 3456 MHz," *The ARRL UHF/Microwave Projects Manual*, Vol 1, ARRL, 1994, pp 3-28 to 3-34.

[2]Campbell, Rick "A VHF SSB-CW Transceiver with VXO," *Proceedings of the 29th Conference of the Central States VHF Society*, ARRL, 1995, pp 94-106. Boards and kits are available from Kanga Products, Bill Kelsey, N8ET, 3521 Spring Lake Drive, Findlay, OH 45840, 419-432-4604. e-mail: kanga@bright.net or **http://qrp.cc.nd.edu/kanga/**.

2304 and 3456 No-Tune Transverter Updates

By Steve Kostro, N2CEI,
Down East Microwave, Inc.

(From *The Proceedings of 1996 Microwave Update*)

Since the introduction of the "NO-TUNE" transverters in the late 1980s[1] Monolithic Microwave Integrated Circuits, MMICs, have greatly improved in performance and yet remained economical enough for amateur use. This article will discuss various options and some results in improving the performance of the two transverters originally designed by Jim Davey, WA8LNC. The three basic issues that will be covered will be Receiver Performance and Options, Transmit Power and Purity, and Local Oscillator Multiplication. Both transverters will be discussed simultaneously allowing comparisons to be made in construction and end results with some words on construction technique. Since detailed drawings of component placement are included, it was not found necessary to include schematics.

Receiver Performance and Options

When the receiver sections of the two transverters were originally designed, MMICs with noise figures of less than 4 dB were not economically available to most amateurs. Although low noise GaAsFET preamplifier designs[2] that were being used by hams were considered, they required special bias supplies and a large amount of circuit board space. With the scope of the project being simplicity and repeatability, and driven by economics, GaAsFETs were deemed not practical to implement into a simple transverter design. Today, with the advent of affordable GaAsFET MMICs, we now have stable drop-in, or with a little circuit modification, designs that will have less than a 2 dB NF and over 20 dB gain. The HP/Avantek MGA85676 became the MMIC of choice for both units (Figures 1 and 2). It replaces the first MAR 6 MMIC in both transverters and will produce a 2-3 dB NF improvement. (See Table 1 for approximate specifications of the MGA85676.) The transverter's noise figure could be optimized for <2 dB by performing the input circuit modification shown. With the increased gain that the MGA85676 offers, the proceeding MMICs should be com-

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

Table 1

MGA86576

Frequency	*Gain*	*50 ohm NF*	*Optimized NF*	*I drain @ Op Volts*
2 GHz	>22 dB	2.1 dB	1.5 dB	16 mA @ 5 V
4 GHz	>22 dB	2.0 dB	1.6 dB	16 mA @ 5 V

Table 2

Model #	*Gain @ 2 GHz*	*Gain @ 4 GHz*	*Pout @ 1 dB comp*	*I drain @ Op Volts*
ERA-1	>11 dB	>10 dB	+13 dBm @ 2 GHz	50 mA @ 3.8 V
ERA-2	>14 dB	>13 dB	+14 dBm @ 2 GHz	50 mA @ 3.8 V
ERA-3	>19 dB	>16 dB	+11 dBm @ 2 GHz	35 mA @ 3.8 V
ERA-4	>14 dB	>13 dB	+19 dBm @ 1 GHz	80 mA @ 5 V
ERA-5	>18 dB	>15 dB	+19 dBm @ 1 GHz	80 mA @ 5 V

The tables above are to be used as a guideline for Amateur Radio design work. They do not completely reflect the manufacturers' specifications. Please consult the manufacturer for complete data.

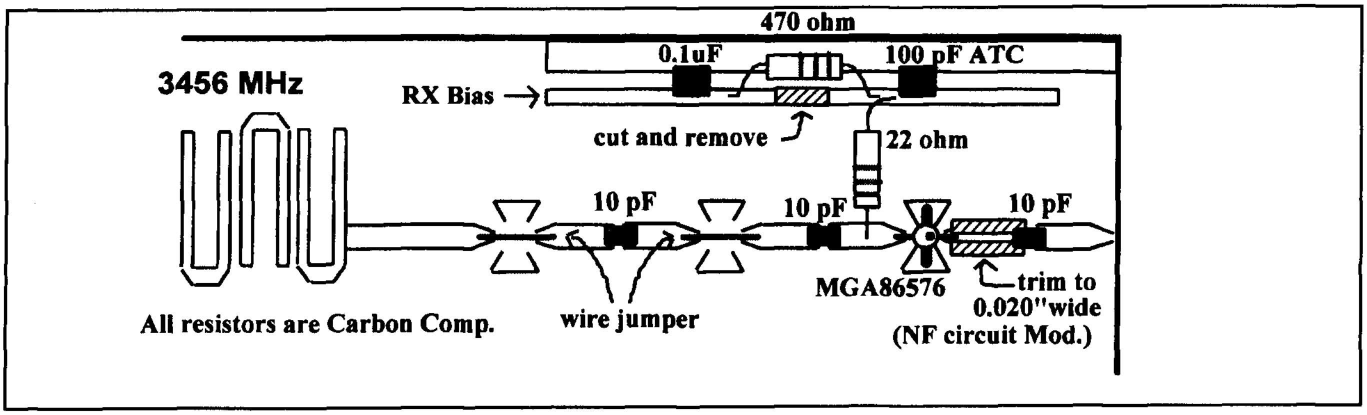


Figure 1

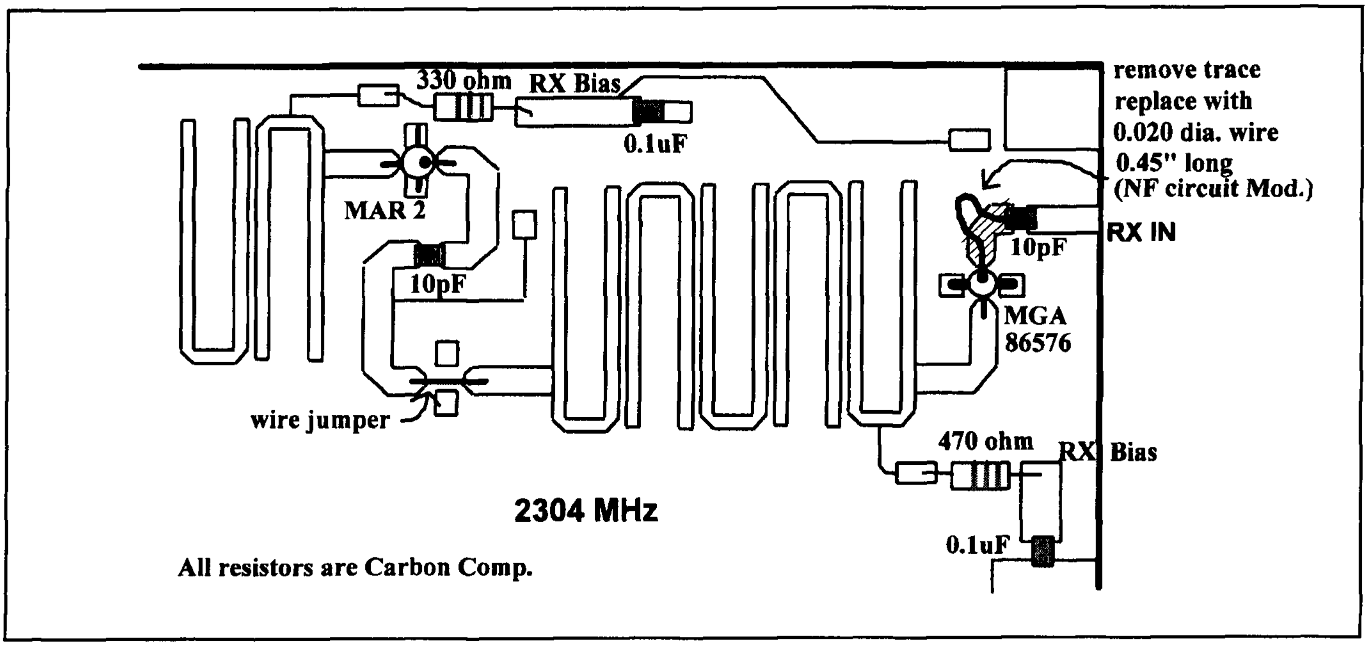


Figure 2

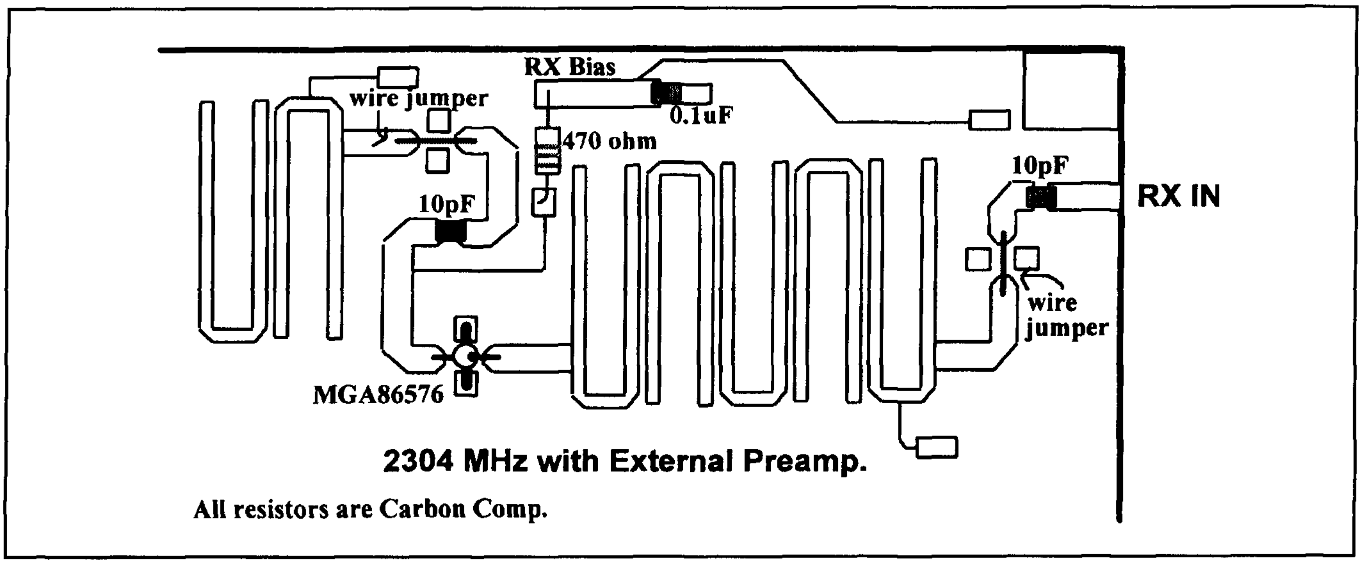


Figure 3

pletely removed on the 3456 transverter. On the 2304 unit, remove the second gain stage (between the filters) and replace the third stage with a MAR 2. If you plan to use an external GaAsFET LNA with the 2304 transverter (Figure 3), remove all three gain stages and install the MGA85676 MMIC in the second stage position, between the two bandpass filters, to ensure that the out of band products amplified by the preamp are filtered out. If an external LNA is to be used with the 3456 transverter (Figure 4), a filter should be installed (pipe-cap[3]) ahead of the transverter. It will not be necessary to perform the low noise figure modification on the MGA86576 if it is used as a second stage.

When implementing these changes, check all ground wraps and/or ground rivets. Be sure they are as flat as possible and are clean connections. The MGA86576 will oscillate if there is an excessive amount of inductance on its ground leads. Also verify that there is a good ground connection at every by-pass capacitor! If you are modifying an old board, oxidized, missing, or broken ground connections could have been the original problem keeping the transverter from working in the first place. Take your time in checking them out this time (It's a thought!).

Transmit Power and Purity

With the introduction of Mini Circuit Laboratory's ERA series of MMICs (Table 2), it is now possible to have as much as 100 mW of output power on both bands. Both transverters used a MAR-8 as a final power amplifier that at times because of its low frequency gain, became very unstable. If you managed to tame it down, 5 to 10 mW output was realistic. There were a few mods that incorporated a single gate GaAsFET as a replacement, but they were not a drop-in solution. The other problem with the TX circuit was that the radiated LO signals, fundamental and multiplied, would conduct into the final gain stage and became amplified to levels as much as –10 dBc. The radi-

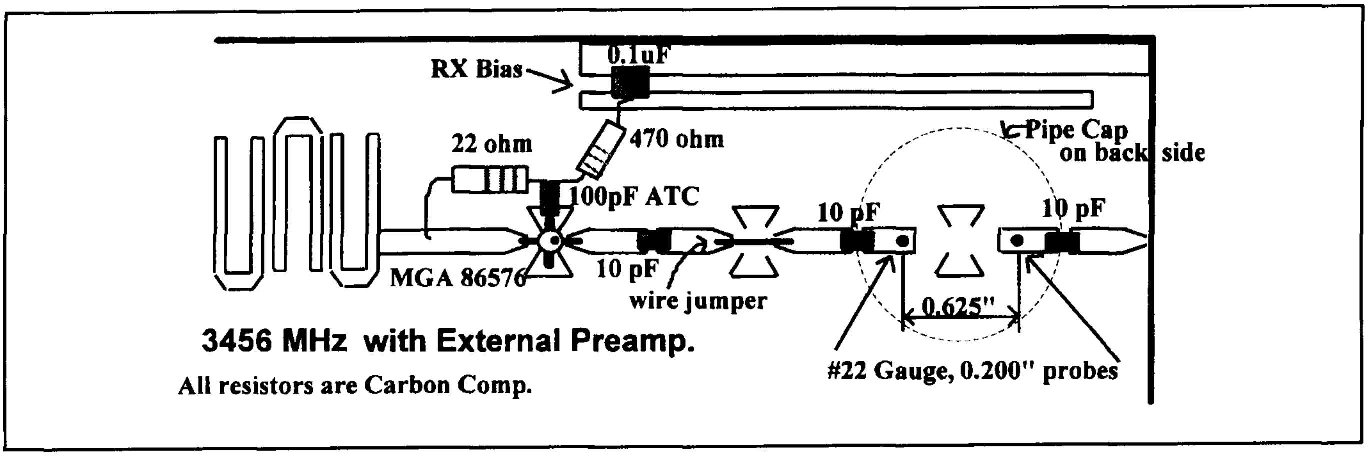


Figure 4

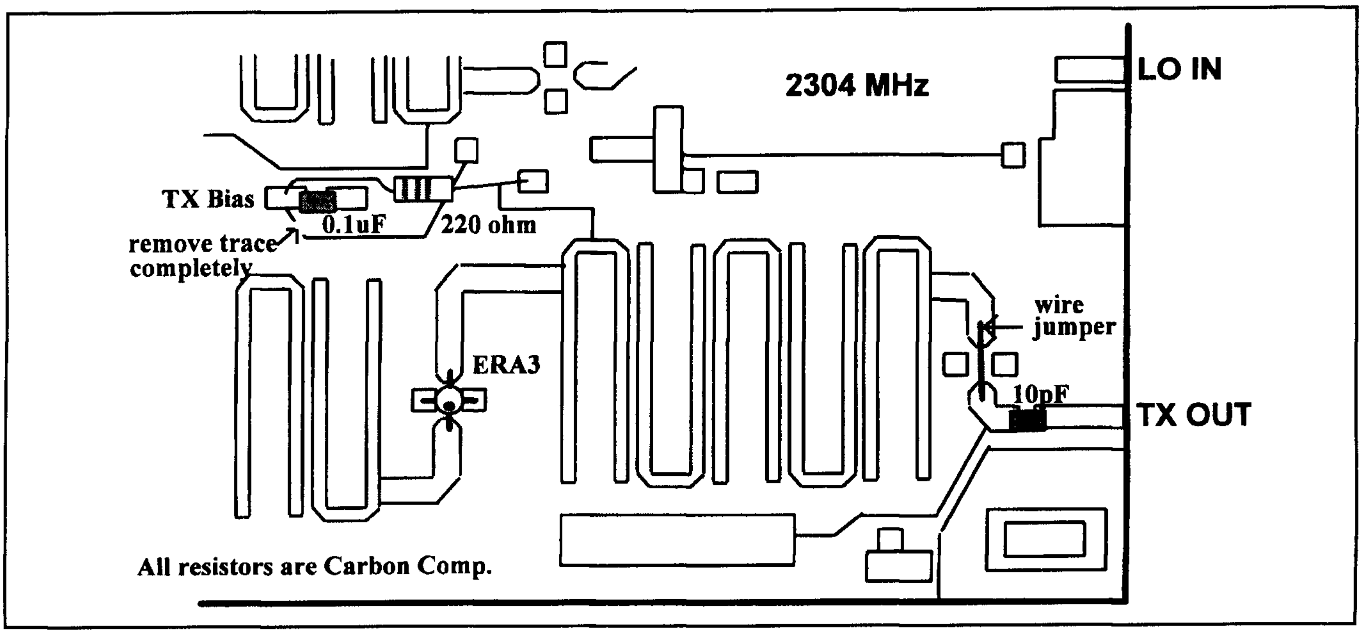


Figure 5

ated LO is a separate problem and will be addressed later in this article, but with the new MMICs and a brief explanation, both problems will be improved on.

The 2304 transverter uses two gain stages on the TX side, a MAR 3 and a MAR 8. Remove both of them. The 2304 PCB design (REV. D) only cycles the bias on the output MMIC. This needs to be changed so the TX bias is only on during TX for both stages. Perform the suggested modification to the circuit board in Figure 5. (Cut and remove the trace!) For a cleaner than original signal, install a ERA-3 in the position between the two filters . This is all that is needed for 10 to 15 mW of output. If higher output is desired, a ERA-4 could be installed in the final output position (Figure 6). If the LO is not modified, expect to see close to 100 mW output at 2304 and as much as 10 mW of output at 2160. This might be okay if you plan to use an extra band-pass filter on the output before amplifying it by any other gain stages. If other output levels are required, select any other MMIC that meets your specification.

The 3456 transverter is very similar to the 2304 unit but it used three gain stages in the TX chain. Remove all of them. In this design, you have a little more flexibility. If you figure a clean –15 dBm out of the mixer/filter combination, decide what you would want for an output, refer to Table 2 and do the math. Only recommendations are not to cascade any other MMIC with an ERA-3 between the two band-pass filters. A good combination (Figure 7) would be to use a ERA-2 driving a ERA-1 between the filters. This alone will produce 10 to 15 mW of output. If higher power is desired, an ERA-3 between the two filters driving an ERA-4 final (Figure 8) will deliver up to 100 mW output but will require a external band-pass filter if additional gain stages are to be used. Again it is recommended that the LO chain be modified to improve the spectral purity.

Like in the receive section, care must be taken in the assembly and to be sure about the grounding. If there are any doubts, fix it. Also if using the ERA-4s or 5s, care must be taken because they will dissipate a lot more power than

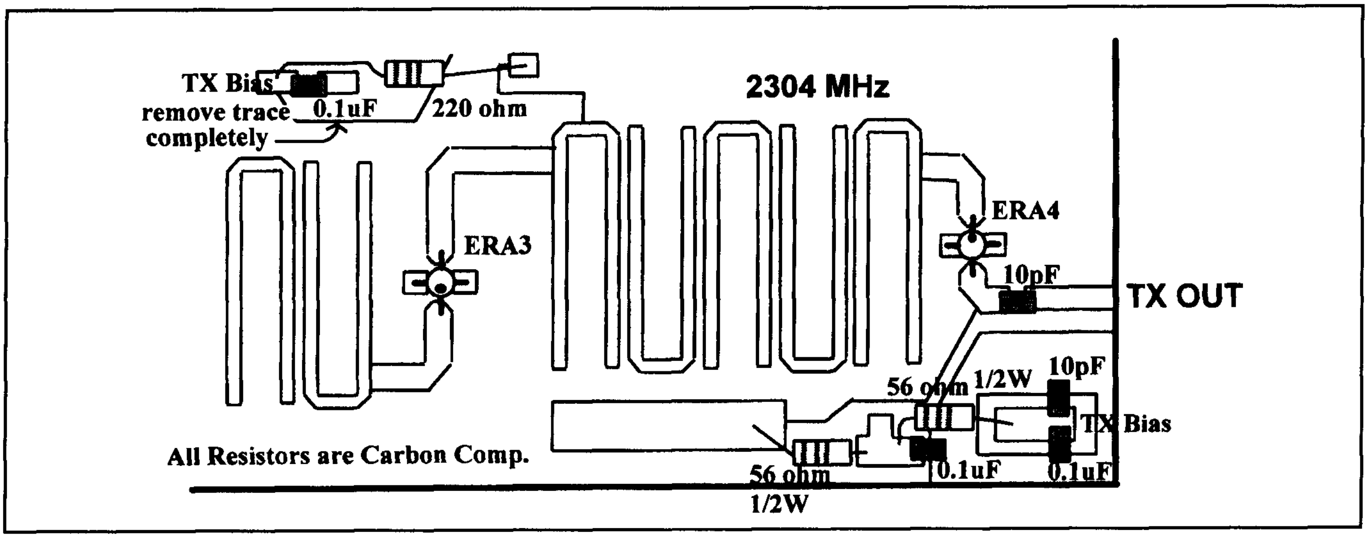


Figure 6

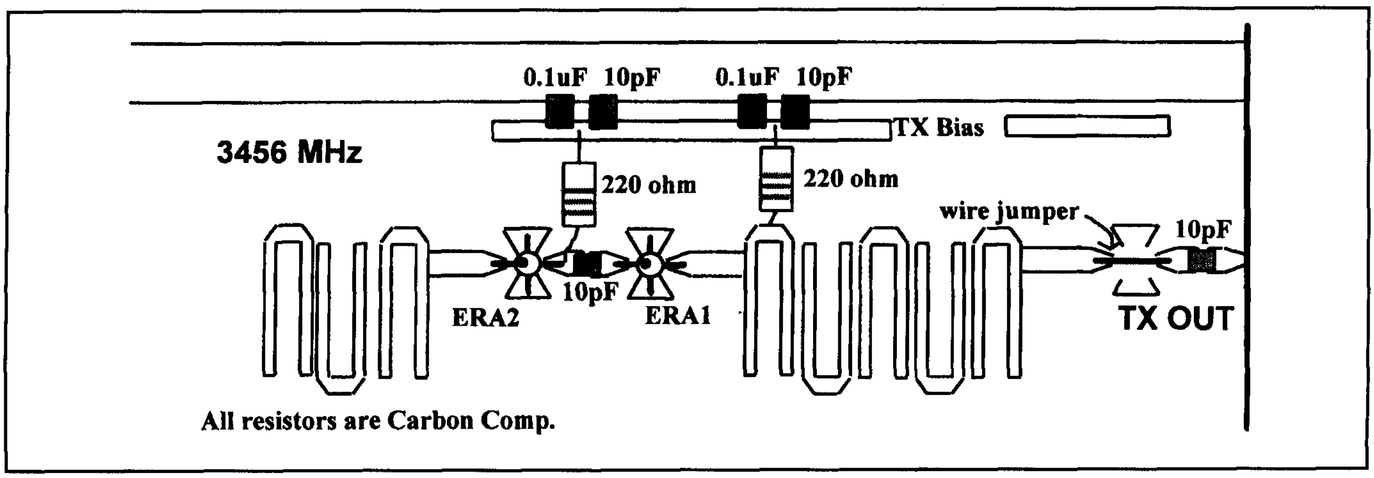


Figure 7

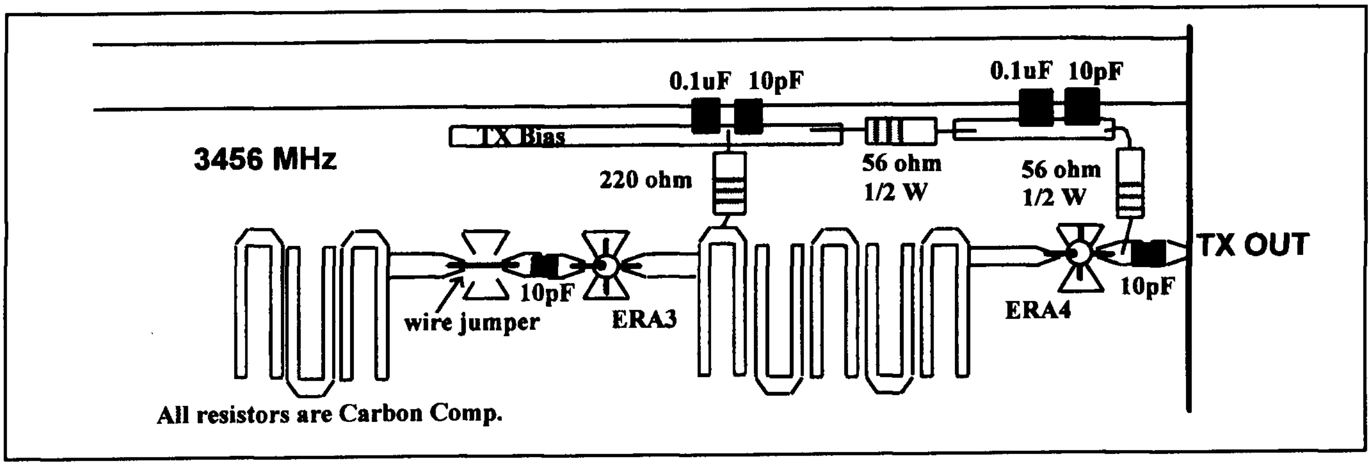


Figure 8

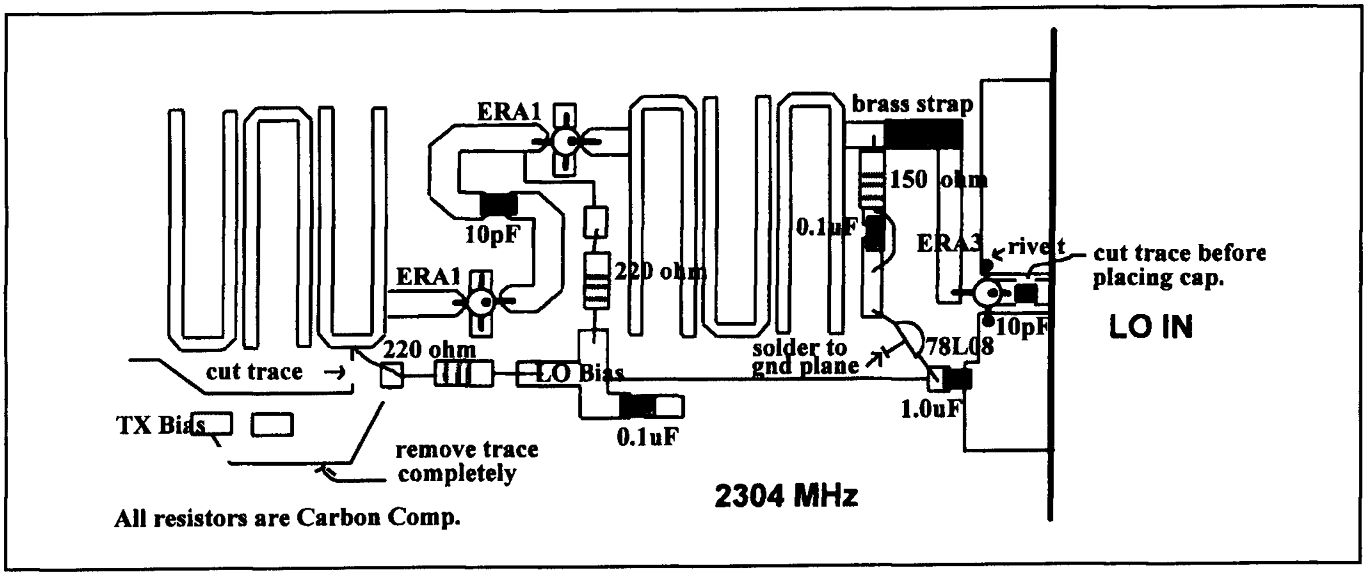


Figure 9

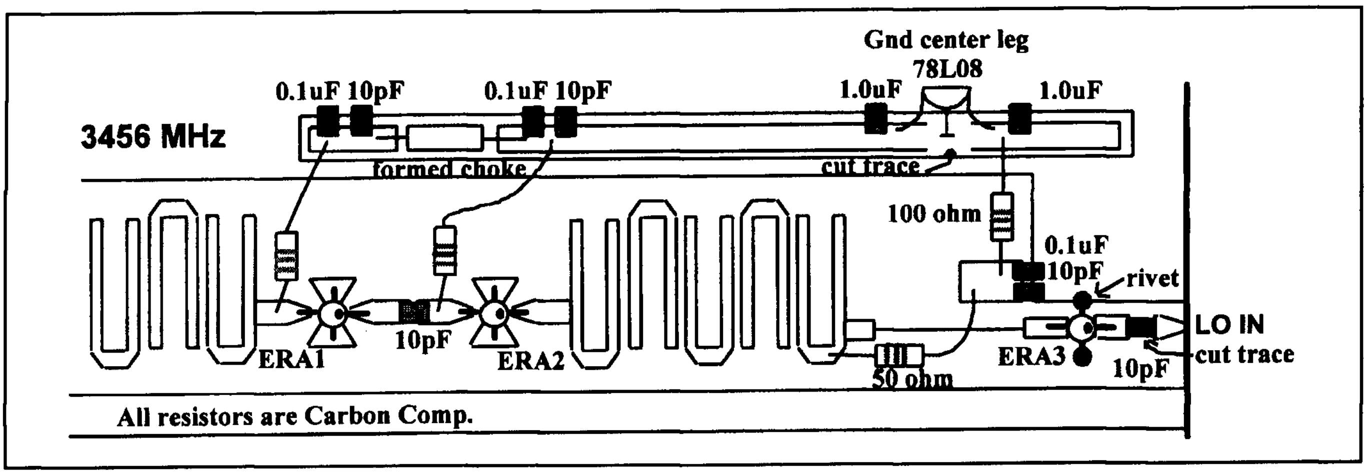


Figure 10

the older MMICs. Half-watt resistors get real warm! If you make a resistor change, do the math!

Local Oscillator Multiplication

The most difficulty encountered in using both transverters has always been caused by the 4X and 6X multiplier circuits. There have been many fixes and cures published and talked about, but the problems still remained. The major problem being the high level of 540 and 552 MHz. Energy injected into the harmonic generator circuit would be reflected back or radiated resulting in a starved amplifier chain, which would allow it to oscillate at its own desired frequency. With two fundamental oscillators now being introduced to the mixers, (the main and the self oscillator) it is easy to understand why the transverters worked sometimes really well and not so good all of the other times. Receiver noise is a good indication of operation but at what frequency would it be at if the LO chain had its own frequency being generated? Both transverters have fooled many noise figure meters. If any of the mods are performed on the two transverters, the LO chain modification would put you on the right track in fixing all of the transverter's other problems.

After reading an article published by Jim Davey on Frequency Multipliers[4] and discussing it with him, I found that it was possible to implement this idea into both transverters with little difficulty and produce huge results. Using the ERA-3 as the MMIC multiplier, it was found that a "sweet spot" existed as far as drive level versus operating voltage. A voltage was derived to allow a LO drive of +0 dBm with a plus or minus delta of 3 dB to be used. Remove all of the components and reassemble per Figures 9 and 10. When installing the ERA-3 multiplier, care must be taken in proper placement and grounding. New ground pads need to be formed with rivets. As for the amplifier section, all of the instability of the MAR-8 MMICs are eliminated along with the radiation effect. The new multiplier chain will deliver 10 mW of LO drive to the power dividers without even winding a coil and having nothing to tune.

The standard LO designed by Rick Campbell, KK7B,[5] will need the final MMIC removed from the circuit and replaced with a wire jumper. A +3 dBm signal should be the maximum obtainable level and will work correctly with the new multiplier circuits. Future discussion of the KK7B LO concerning frequency netting and stability will be discussed at a later date.

Final Notes

The modifications have been used repeatedly for over a year with excellent results. Local oscillators have been packaged together with the transverters that produce < 2 dB NF, 25 mW Pout and spurious responses of –45 dBc. Please remember that all of the new MMICs in the world don't make up for poor construction techniques. Use nothing but carbon composition resistors and keep the leads to the active circuits as short as possible. If you have a doubt about a ground wrap or rivet, repair it! Make sure that every by-pass capacitor is directly connected to a ground to the back side of the board. With the modifications, enclosures are less critical, but still could be a problem. Keep away from waveguide sizes boxes.

Contact the author at Down East Microwave (Phone: 908-996-3584, Fax: 908-946-3072, Web site: **http://downeastmicrowave.com/**) for availability of the latest PC boards. Future designs will be available that use all of the suggestions above. Some new features of the new designs will include plated through holes and a machined enclosure to accommodate both new transverters. As newer MMICs appear on the marketplace, updates will be made. Feel free to experiment with them and most of all, Have Fun!

Notes

[1] *QST*, June 1989 for the 3456 transverter, *Microwave Update 1989* & *QST*, Dec 1992 for 2304 transverter

[2] "Simple Low-Noise Microwave Preamplifiers for 2.3 Through 10 GHz," *QST*, May 1989

[3] "Cheap Microwave Filters From Copper Plumbing Caps," *Microwave Update, 1988*

[4] *ARRL UHF/Microwave Projects Manual,* pages 5-13 through 5-15

[5] "A Clean, Low-Cost Microwave Local Oscillator," *QST*, July 1989

Modernizing The 3456 MHz No-Tune Transverter

By Jim Davey, WA8NLC

(From *The Proceedings of 1994 Microwave Update)*

Introduction

Back in 1988 I presented a single board no-tune transverter for 3456 MHz at the Microwave Update '88 Conference in Estes Park, CO. Initial interest was high in this design and it was subsequently published in the June, 1989 issue of *QST*. Since that time over 190 of the transverters have been sold all over the world.

While the 3456 MHz unit has enabled many to easily get a basic station on the band, it has not been without a few problems. Stability of the active devices has been marginal under certain conditions. Some people have also had a problem packaging the finished board in a larger enclosure with the other stages needed to make a high performance transverter.

Since the time of the initial design a couple of things have happened: (1) new and better MMICs have become available and (2) I have become more enlightened on the art of microwave circuit construction. Also, many people have had a chance to work out some of the problems, most notably WB5LUA, N1BWT and WA3JUF. Their willingness to take the time to publish their improvements is greatly appreciated. This paper will gather together the suggested modifications made by others and add some new ideas of my own.

Before getting into a discussion of the shortcomings and improvements of the transverter, I should reiterate the basic design goals of the unit:

1) All filtering is done by non-critical band-pass filters. This remains the dominant feature. No metalwork or filter tuning is required to achieve a reasonably clean transmit signal (unwanted outputs down >40 dBc) and an image-free receiver.

2) Low cost. If anything, silicon MMICs have gotten cheaper since the design was first published. And where can you get a decent 3456 MHz mixer for under a buck?

3) Integrated approach to multistage circuitry. Having all the essentials on one board reduces the amount of packaging, connectors and cabling required to get the unit together.

I believe that (1) and (2) above are still important features. The more hamfests I go to, the less I am inclined to emphasize (3), especially when I see boxfuls of SMA jumper cables at 50 cents apiece.

Packaging the Transverter

Let's face it, microstrip circuits radiate[1] and this transverter is no exception. The amount of radiation is high enough to cause some problems and has made it difficult to box up the transverter in a practical matter. Also, when an outboard preamp and/or power amp is desired, the isolation between those accessories and the board may not be high enough to allow them to be packaged together without feedback. Several people have tried to cover the suggested "brass strip box" in order to isolate the low level stages from the rest of the station. Generally this has caused problems ranging from high level spurs in the transmitter to oscillation of the gain stages.

The effect of an RF-tight lid on the transverter can be almost totally eliminated by using some true RF absorber material on the inside of the lid. I tried this experiment with some of the 1/8-inch thick material available through Tom Hill[2], WA3RMX/7, and found it to do the trick. Black carbon foam does not work! Without the RF absorber, one of the most noticeable effects is that the metal cover will unbalance the mixer sending lots of local oscillator energy through the transmitter. The onboard filtering is not sufficient to eliminate the extra LO.

Another observable effect of board radiation is the coupling across the board from one circuit to another (LO multiplier to TX amp for instance) and along the length of the board (TX mixer to TX output). This can be eliminated by boxing the TX, LO and RX in separate enclosures of the proper size. I cut a transverter board into TX, LO and RX sub-modules using a sharp knife and connectorized all the ports. Each board ends up about 1.3 inches wide by 6 inches long. When boxed up using the suggested 3/4-inch brass strip and a cover made of thin aluminum, the resulting enclosure looks like a waveguide-below-cutoff. Each cover still needs to be treated with RF absorber material as described above.

Unless the transverter is totally redesigned on much thinner board to reduce the radiation effects, the above suggestions, combined with those for stability described in the next section, can produce a stable and clean transverter.

Stability Improvements

Instability has been a sporadic problem in three areas:

1) The cascaded MSA 0685/0685/0185 in the receiver can oscillate depending on source SWR, lead dress of the biasing components and power supply voltage.

2) The cascaded MSA 0885s in the LO amplifier can oscillate at VHF frequencies due to excessive gain below 1 GHz.

3) The MSA 0885 transmitter output MMIC may be unstable into certain loads.

The first problem above was largely eliminated by replacing the original 5 pF series coupling capacitors in the receiver with 1.0 pF. This reduced the low frequency gain enough to keep things under control most of the time. All the Down East Microwave units use the smaller value capacitor. Varying power supply voltage (as when using batteries) can still change the device S-parameters enough to get an oscillating condition. Paul Wade, N1BWE, described using an LM 2941 CT low dropout regulator to eliminate this problem[3]. Down East Microwave sells these regulator ICs as well as a board level kit.

Another solution is now possible for only a few extra dollars by using one of the new Hewlett-Packard MGA 86576 GaAs MMICs. This device has a flat gain curve down to low VHF frequencies and a much lower intrinsic noise figure. It requires a +5 volt supply, thus a regulator from the normal 12 volt supply is suggested by HP. I put one in stage #3 of the receiver and cut off the first two stages before enclosing the board in a brass box (Figure 1). Not only did the system noise figure drop from 5 to 2.0, but the receiver now only requires 16 mA of current instead of the 45 mA with the silicon MMICs. Rus Healy, NJ2L, has done this modification to his rover station with similar results.

The potential solutions for eliminating oscillation in the LO multiplier amplifier are somewhat limited because the LO needs to generate enough power to drive two mixers. A possible alternative MMIC lineup that would meet the gain and output power requirements would be an MSA 0385/0385/0986. Gain at VHF would be reduced considerably over the 0885 pair, but there is not a lot of room on the existing board for three stages and their bias resistors, and the three stage amp would draw over 100 mA. I did not try this modification. Instead, I took a look again at what the MGA 86576 could do. The specs say it cannot produce more than about +7 dBm at 3312, so the mixers would have a little less injection than before after filtering and splitting. To see what could be done I tried a single MGA 86576 at 5 volts. The circuit is the same as Figure 1 except the 5 pF input capacitor is not needed. The power to each mixer measured +4 dBm, which was about 2 dB more than expected. The output spectrum was exceptionally clean with

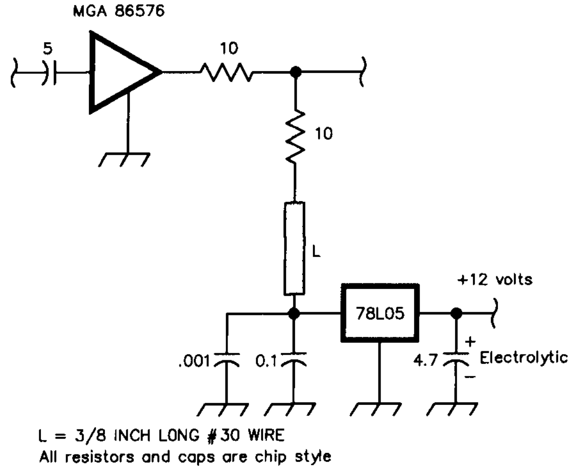


Figure 1

all unwanted products down over 70 dB. As a bonus, I found that the diode multiplier only required +8.5 dBm at 552 to saturate the MGA 86576, about a 7 dB reduction. I currently do not have an explanation for this. LO chain current drops considerably from about 75 mA for the pair of 0885s to 16 mA. I am beginning to really like the MGA 86576!

Dave Mascaro, WA3JUF, recorded his problems with the transmitter section in his club's newsletter, *Cheese Bits.*[4] It includes a detailed description of how to modify the board. I have tried this modification and it works well as described below. A discrete GaAsFET here is a better solution than the GaAs MMIC since much more output power can be obtained. Much of the feedback I have received through the years on both the 2304 and 3456 no-tune transverters has been on how one might get more power out of the transmitter. This is understandable since when these transverters are used *barefoot* as a rover rig and the other station is a fixed home or mountaintop contest station, the limiting factor in making the contact is usually going to be the ability of the fixed station to hear the low power transmitter. Although the noise figure of the original no-tune transverter is a few dB higher than a *performance station*, it is not likely going to be an issue since the fixed station usually runs higher power.

I decided to give an ATF 10135 a try as it looked like a good choice for gain and power output capability. For this device I had to make a few changes to Dave's circuit to bias the FET properly and to ensure stability. The drain resistor was dropped to 5.6 ohms to allow a Vd of 3.6 volts. The source bias resistor needed to achieve 60 mA of drain current is 18 ohms. I had to lower the Q of the gate choke with a series 5.6 ohm chip resistor to stop an oscillation at 11 GHz. With these changes I got an insertion gain of 10.5 dB. When the driver MMICs are replaced with another MGA 86576 as described below, a power output of +17 dBm (50 mW) was obtained. I have not done any extensive testing of stability with different termination SWRs like Dave did on his unit. This issue may best be investigated with a computer modeling program.

No problems have been reported to my knowledge with the two low level driver stages in the transmitter. But having just replaced the balance of the transverter with gallium arsenide, why finish the job with an MGA 86576 in place of the 0185/0285 and then just run the whole transverter off +5 volts? I couldn't resist. The results were excellent for the driver replacement. The MGA 86576 has about 6 dB more gain than the 0185/0285 pair, giving about +6.5 dBm out of the second transmit filter. This higher level is welcome as the desired 3456 MHz signal out of the mixer is reduced slightly due to the lower LO power and the ATF 10135 can achieve greater output with more drive.

Summary

Having done all the above modifications to a 3456 no-tune transverter, I have a stable unit with characteristics listed in the table below. The specs for the original unboxed unit are included for reference.

The total cost of the MMICs for the original unit is about $15. The GaAs replacements will run you only about $28 not counting any onboard or external voltage regulators. A small price for such an increase in performance.

Parameter	*Modified**	*Original*
Noise Figure	2.0 dB	5.0 dB
TX Output	50 mW	8-10 mW
LO Power @ 552	+9 dBm	+16 dBm
Total Current	110 mA	186 mA

**Three separate enclosures, absorptive material on cover, MMIC replacements as described in this article.*

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Home-Brewing a 10 GHz SSB/CW Transverter

Part 1–Narrowband 10-GHz operation—without exotic or surplus parts—has finally arrived for the microwave builder!

By Zack Lau, W1VT

(From *QST*, May 1993)

Looking for some challenging microwave equipment to build? How about a complete 10-GHz transverter with stability good enough for weak-signal CW work? You don't have to find any exotic pieces to build this project—all of the parts are fairly common. In fact, everything has been available for years.

Despite the transverter's compact package, however, it consists of several modules that you must build. And although the VHF/UHF circuitry follows the no-tune concept developed by Jim Davey, WA8NLC; Rick Campbell, KK7B; and others, the X-band (10-GHz) parts need to be tuned up—preferably with a spectrum analyzer that works through 10.4 GHz.[1]

Design Philosophy

Unlike the no-tune transverters,[2] I decided to develop the transverter as a set of building blocks with stainless-steel or gold-plated SMA connectors. Although this con-

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

Table 1

10-GHz Transverter Performance*

Transmit Converter

144-MHz Drive (dBm)	10-GHz Output† (dBm)
–10.0	3.8
–3.0	8.5
0.0	10.3
1.0	10.8
3.0	11.6
5.0	12.2
10.0	12.8

Power Output versus Supply Voltage†

(Drive signal: –0.8 dBm at 144.06 MHz)

Supply (V)	Output Power (dBm)
10.34	8.5
10.51	10.0
10.75	10.0
12.34	10.0
14.02	10.0

Receive Converter

IF (MHz)	Gain (dB)	Noise Figure (dB)
144	8.82	2.73
146	8.79	2.70
148	8.79	2.71

Noise Figure and Insertion Gain versus Supply Voltage

(IF = 144 MHz)

Supply (V)	Gain (dB)	Noise Figure (dB)
10.3	6.59	3.03
10.7	8.81	2.92
10.9	8.64	2.89
11.3	8.59	2.79
12.4	8.68	2.75
13.4	8.70	2.74
13.5	8.73	2.76
14.9	8.75	2.74

*The data in this table comes from the most recently completed prototype, which consists of the modules described in Part 1 and Part 2 of this article.
†Power output was measured with an uncalibrated HP 435B/8481A.

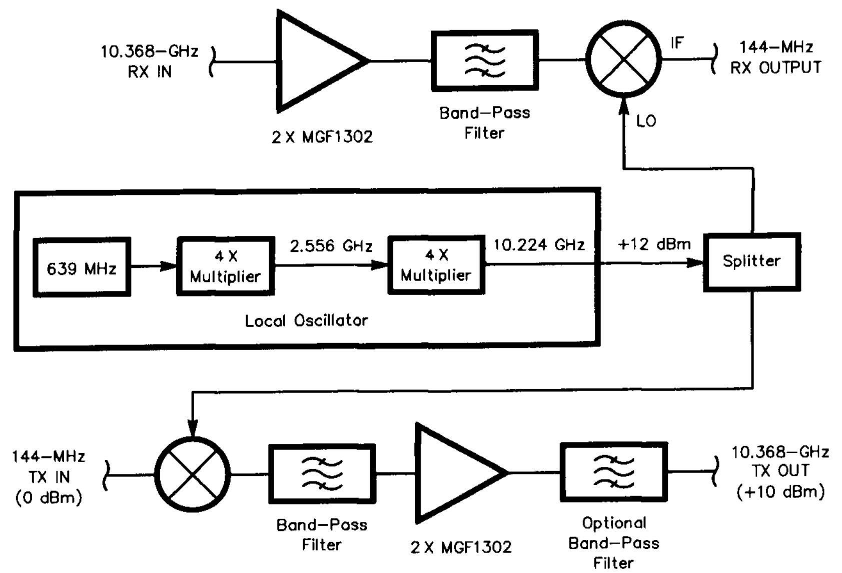


Fig 1—Block diagram of the 10-GHz transverter.

struction method is more expensive and time-consuming than a more integrated approach, it offers several advantages. Most importantly, it allows you to check small portions of the transverter for proper performance. If something doesn't work, troubleshooting is fairly straightforward. And, if you just can't get one of the modules to work, you can simply build another one. Another advantage of this construction method is the shielding that results from packaging circuits in separate boxes. This helps greatly to keep the transverter spectrally clean, with a minimum of spurious outputs and responses. Finally, the transverter is easily updated or expanded to take advantage of improving technology. Making its receiver section state-of-the-art is simply a matter of adding the 1-dB noise figure preamplifier described in December 1992 *QEX*.[3]

A Brief Overview

Fig 1 and Table 1 show the transverter's configuration and measured performance. A local oscillator (LO) feeds a power splitter that drives a pair of mixers. One mixer is used on transmit and the other on receive. The transmit mixer is followed by a filter and amplifiers. A filter following the final stage is optional. Low-noise amplifiers and an image-stripping filter precede the receive mixer. Without adequate image rejection, the receiver sensitivity can degrade by as much as 3 dB.

The Local Oscillator

The most critical part of microwave narrowband work, the LO, starts off with the circuitry developed by WA8NLC and KK7B. The 106.5-MHz oscillator (Fig 2A) is multiplied by six (Fig 2B) to produce a 10-dBm (10-mW) signal at 639 MHz. This signal is then multiplied by four and amplified to 7 dBm at 2.556 GHz (Fig 3).

This is essentially the same scheme used in KK7B's 2.16-GHz LO in July 1989 *QST*,[4] except that I modified the filters for 639 and 2556 MHz. I also added a 0.47-µF capacitor to provide a low-impedance input for the 78L05 regulator (it can oscillate if not properly bypassed). These circuits are built on fiberglass-epoxy G10 or FR4 PC-board material; the remaining circuits are built on 5880 RT/Duroid.

Choosing the Circuit-Board Material

This part was actually pretty easy: I looked around for something with low enough loss to work well, but that's also readily available to amateurs. The only stock item that meets this description is 0.015-inch-thick (15-mil) 5880 RT/Duroid. This is the same material used in the Tuesday Night Transverter published in the *Proceedings of Microwave Update '88*.[5] The thicker 30-mil 5880 RT/Duroid is definitely unacceptable, as its radiation loss is rather high. Down East Microwave is one possible source.

If availability wasn't an issue, I might have chosen a board thickness that helps to optimize stability via source inductance.[6] Another criteria for choosing board thickness is the interface with the transistors and connectors. Often, it is desirable to minimize the discontinuity between these interfaces by selecting trace widths comparable to the connector diameters and transistor-lead widths. The 15-mil board works pretty well in this area—the 46-mil trace widths fairly closely match the widths of the specified 50-mil chip capacitors.

Crystal Frequency

When choosing an LO crystal, the most important consideration is the crystal's calibration. The tolerance of the International Crystal Manufacturing high-accuracy crystal (#473590) I recommend is 10 parts per million. This means that the crystal can be as much as 1.06 kHz off the marked frequency without deviating from the specified accuracy. Because the LO is multiplied by 96, the transverter's conversion frequency could be as far as *102 kHz* from the ex-

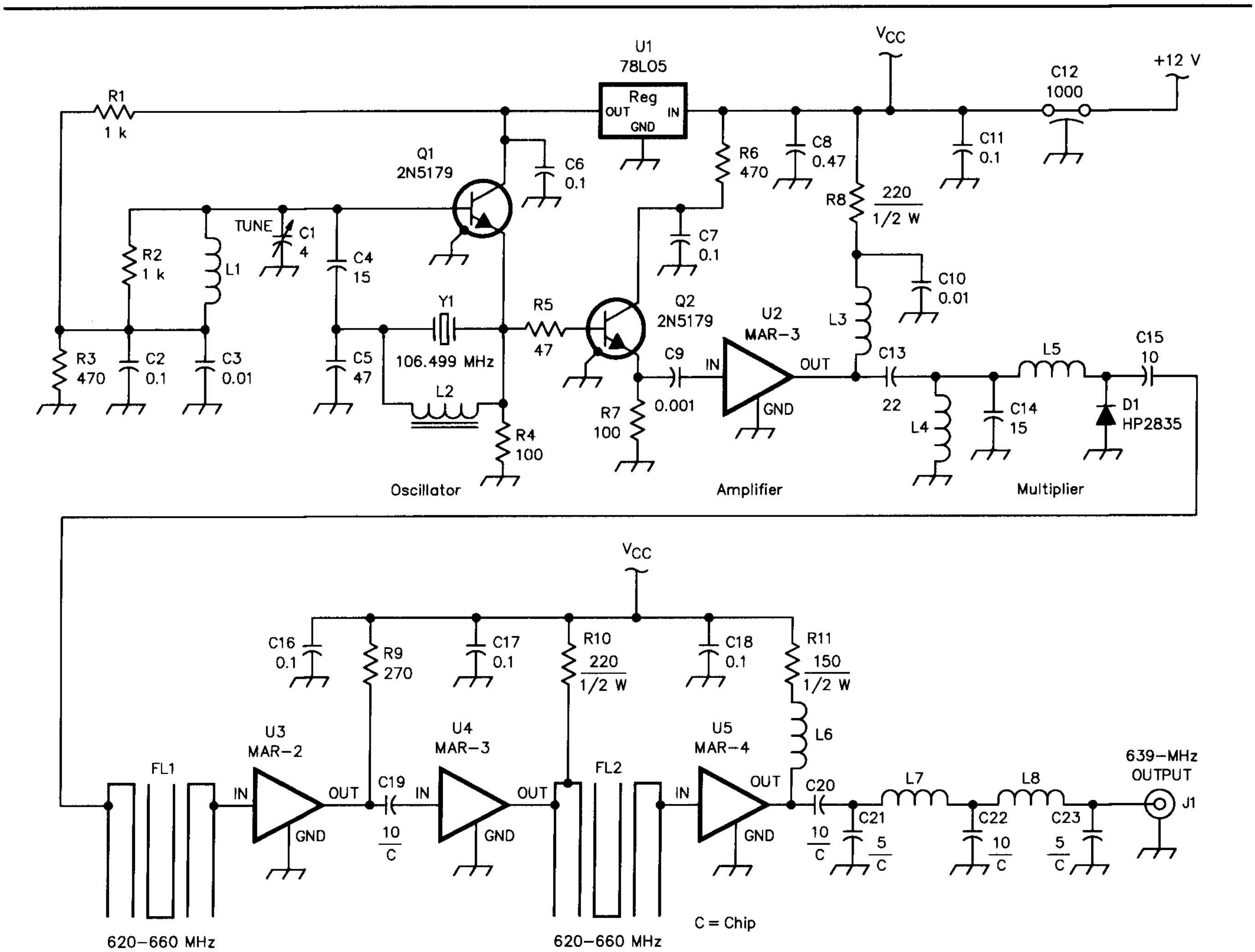


Fig 2—Schematic of the crystal oscillator and times-6 multiplier. Resistors are 1/4-W carbon-film or carbon-composition types unless otherwise indicated.

C1—Air-dielectric trimmer capacitor that can be set to approximately 4 pF. Low temperature coefficient is more important than exact value, as L1 can be adjusted to compensate.
C8—Minimum value required to stabilize U1 is 0.33 µF. An electrolytic capacitor can be substituted if proper polarity is observed.
C12—1000-pF feedthrough capacitor. Exact value not critical (100 pF to 0.1 µF should work well).
D1—Schottky diode. Hewlett-Packard 5082-2835, -2811 and -2800 work well.
FL1, FL2—Band-pass filters printed on PC board.
J1—SMA female chassis-mount connector.
L1, L3, L4, L6—8 turns #28 enameled wire, 0.1-inch ID, closewound.
L2—12 turns #30 enameled wire on T-30-6 toroid core.
L5—5 turns #28 enameled wire, 0.1-inch ID, closewound.
L7, L8—2 turns #28 enameled wire, 0.062 inch ID, turns spaced one wire diameter.
Q1, Q2—2N5179 or BFR91.
U1—78L05 5-V, 100-mA, three-terminal regulator.
U2, U4—MAR-3 or MSA-0385 MMIC.
U3—MAR-2 or MSA-0285 MMIC.
U5—MAR-4 or MSA-0485 MMIC.
Y1—106.499 MHz, fifth-overtone, series-resonant crystal (International Crystal Manufacturing #473590).

pected frequency, even without taking temperature variations into account. Although the oscillator circuit allows some adjustment to compensate for frequency error, attempting to shift the frequency seems to degrade stability.

To make sure that the conversion frequency falls inside the 2-meter band, I specify a 106.499-MHz crystal. Selecting a 106.500-MHz crystal might prove to be unwise if it was cut 10 ppm high—the usual calling frequency of 10.368100 GHz would be just below the 2-meter IF radio's 144.0-MHz band edge—a problem with some radios. You may want to choose another frequency, perhaps even lower, to move the IF to 145 or 146 MHz. If you do this, you'd be wise to investigate possible sources of interference. Keep in mind that hilltops are often pretty bad in terms of interference problems.

The stage following the 639-MHz to 2.556-GHz mul-

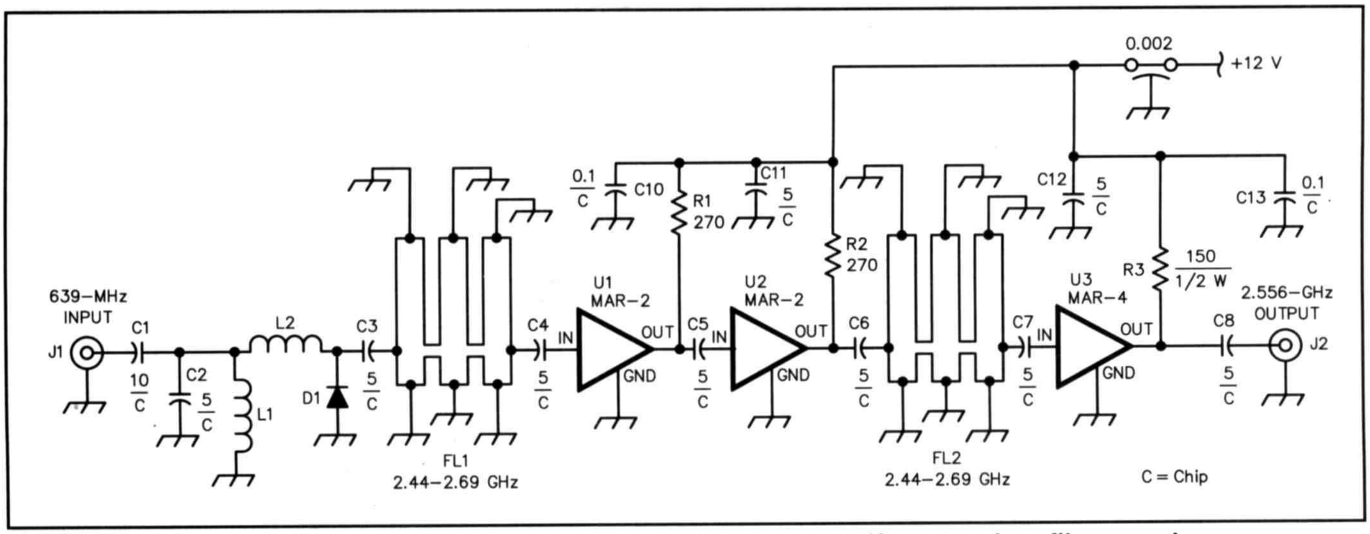


Fig 3—Schematic of the 639-MHz to 2.556-GHz multiplier. Resistors are 1/4-watt carbon-film or carbon-composition types unless otherwise indicated.

D1—Schottky diode. Hewlett-Packard 5082-2835 or equivalent.
J1, J2—SMA female chassis-mount connector.
FL1, FL2—Printed band-pass filters.
L1—3 turns #28 enameled wire, 0.062 inch ID, turns spaced one wire diameter.
L2—Printed inductor.
U1, U2—MAR-2 or MSA-0285 MMIC.
U3—MAR-4 or MSA-0485 MMIC.

tiplier is a GaAsFET multiplier, filter and amplifier (Figs 4 and 5) that takes the 2.556-GHz input and provides at least 12 dBm at 10.224 GHz to the LO splitter/mixer board (to be described in Part 2).

Bias Supplies

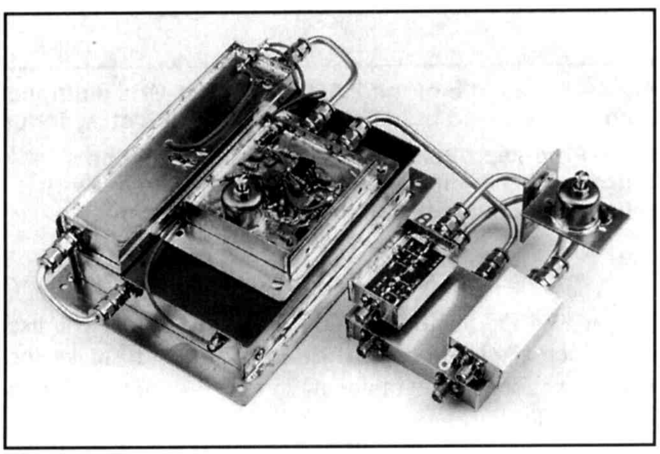
I know it's not the cheapest way to go, but I decided to build a negative bias supply into each module that requires one (all the stages that use MGF1302s). This reduces the chance of misconnecting the positive and negative supplies. I also opted for active bias supplies, as shown in Fig 5. This figure shows the two equations for calculating components for different bias conditions. For instance, to bias an FET at 3 volts and 30 mA, you first calculate the effect of any resistors used for stability. Often, a 51-Ω resistor is used to stabilize the circuit; if present, it increases the circuit bias voltage to 4.53 volts. One set of standard values that comes close to the bias conditions given above, and accounts for the 51-Ω resistors, is: R_{dn} = 16 Ω, R_{an} = 3.6 kΩ, and R_{bn} = 1.1 kΩ.

I used Intersil ICL7660s to generate the negative bias supplies because they require few external parts. A cheaper alternative is to use NE555 timer chips as oscillators driving rectifiers.[7] I published such a circuit, with a PC-board pattern, in March 1991 *QEX*.[8]

Filter Construction

The transverter's band-pass filters are made from half-inch copper pipe caps, as shown in Figs 6 and 7. These were developed by Roman Wesolowski, DJ6EP; and Kent Britain, WA5VJB.[9] They're affordable, too: You can buy half-inch plumbing caps at home-supply stores for as little as 12 cents each. (Designed to cap pipes that are 0.5 inch ID, these caps actually measure 0.62 inch ID and about 9/16 inch long.) I drill and tap the caps (at top center) with #4-40 threads and use nickel-plated brass screws; unplated brass screws should work as well. Kent Britain has forced steel screws through the caps to thread them. Don't use these screws for tuning, though, as steel is unacceptably lossy. I often polish my plumbing caps so that they look nice and solder easily.

A pipe-cap filter ahead of the mixer is adequate in terms of system noise figure, giving an image rejection around 24 dB with a 144-MHz IF. For critical applications, a waveguide filter, such as the one published by Glenn Elmore, N6GN, in July 1987 *QEX*,[10] is recommended. With such a filter, 50 dB of image rejection is easily obtained with a 144-MHz IF. However, for lightweight portable transceivers, plumbing-cap filters seem to be the best compromise. For a clean transmitted signal, you should use one at the final transmit amplifier's output as well.



The completed transverter.

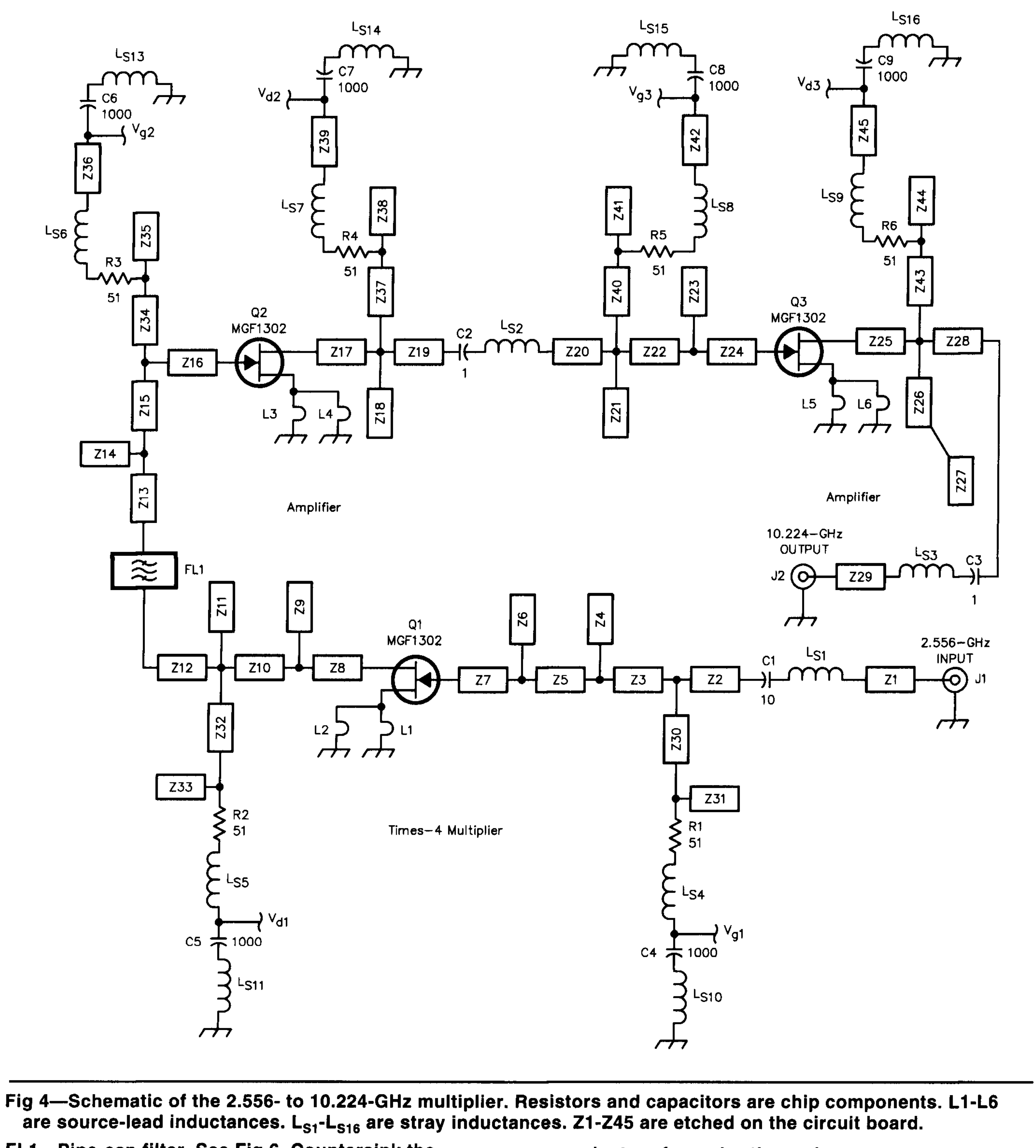


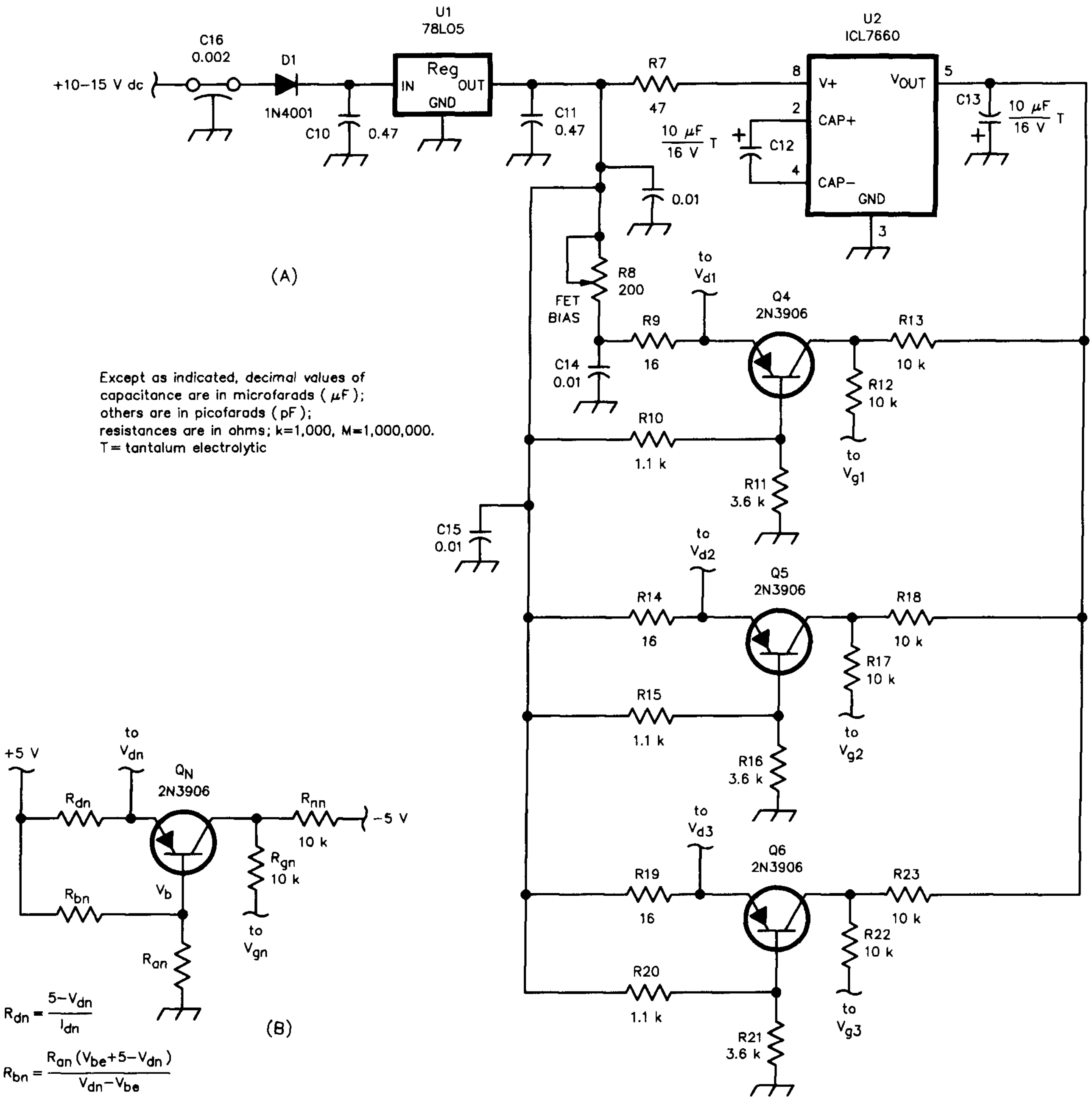
Fig 4—Schematic of the 2.556- to 10.224-GHz multiplier. Resistors and capacitors are chip components. L1-L6 are source-lead inductances. L_{S1}-L_{S16} are stray inductances. Z1-Z45 are etched on the circuit board.

FL1—Pipe-cap filter. See Fig 6. Countersink the ground-plane side of the circuit board hole (by hand) to keep the ⅛-inch UT-141 probe center conductors from shorting to it.

Q1-Q3—Mitsubishi MGF1302 GaAsFET. Substitution not recommended.

The filters are built on unetched, double-sided, 1/16-inch G10 or FR4 PC-board material. I recommend that you use 0.141-inch semirigid coaxial cable (UT-141) to make the probes. A probe length of about 75 mils is optimum. If you cut them too short—say, 50 mils—the insertion loss climbs from an acceptable 1 to 2 dB to as much as 5 or 8 dB. If the probes are cut too long—say, 100 mils—the image rejection drops to a measly 10 to 14 dB, though the insertion loss also drops (to 0.5 dB). The probes are spaced 5/16 inch center to center and the pipe cap is soldered to the ground plane so that the probes are centered within it.

How do you determine the best probe lengths and spacing for pipe cap filters? I developed the filters in this transverter using a spectrum analyzer and trial and error.



$$R_{dn} = \frac{5 - V_{dn}}{I_{dn}}$$

$$R_{bn} = \frac{R_{an}(V_{be} + 5 - V_{dn})}{V_{dn} - V_{be}}$$

Fig 5—At A, schematic of the multiplier bias supply. At B, generalized FET bias circuit and equations.

C12, C13—Tantalum electrolytics preferred, but aluminum electrolytics should work.
C16—Feedthrough capacitor. Value not critical.
D1—Reverse-polarity protection diode.
Q4-Q6—General-purpose PNP transistor; 2N3906 and MPS2907 suitable. Plastic-cased devices are easiest to use.
R8—200 Ω used in the prototypes. 100 or even 50 Ω may be suitable.
U1—78L05 5-V regulator.
U2—ICL7660 negative-voltage generator.

The signal source was an X-band mixer and the 10.224-GHz local oscillator. I could have done a lot better with a network analyzer or a scalar sweep setup, but I used what was available to me.

Filter construction can be fairly critical for optimum performance. In particular, the probes must be accurately cut to length. I estimate my error margin in measuring and cutting probe lengths to be about 10 mils. The ends of the probes are filed flat, not chamfered or rounded. Filter loss seems to be a few tenths of a decibel lower with the dielectric left on the probes, although it is easier to trim the probes exactly with the dielectric removed.

You may be tempted to use 0.085-inch semirigid cable because it's easier to handle than UT-141. A similar filter I made using this material gives 24 dB of image rejection, but has 3.4 dB of loss. The probe length for this cable is 70 mils. A filter using 100-mil probes of 0.085-inch semirigid cable has only 2 dB of loss, but the image rejection drops to a barely acceptable 17 dB. UT-141 is better for this application.

I recommend that you assemble the cable and solder it to the ground plane *before* measuring and cutting the probe length. Otherwise, the length may change as you work on the cable. With these filters, a potential problem is caused by the center conductor moving around slightly, particularly when the cable is straight and the center conductor forms the center contact at the connector end. Bending the cable helps to prevent this problem, but the best solution is to use connectors that captivate the center conductor, keeping it from being pushed inward.

Enclosures

As shown in the photo on page 2-27, I use 0.025-inch-thick, half-inch-wide brass sheet stock to make the enclosure walls. Instead of soldering SMA connectors to the walls, I attach them with #2-56 screws; either method is acceptable. The 25-mil brass stock is ideal for tapping small screw holes. Other commonly available thicknesses can also be used, although 20-mil stock is a bit flimsy and 32-mil stock is more difficult to solder.

Duplicating the Circuit Boards

Using PC-board layout software, I've developed artwork for each of the transverter's circuit boards. To make it as easy as possible for *QST* readers to build this transverter, ARRL HQ is making the circuit-board artwork available in three forms: as PostScript files downloadable from the ARRL HQ telephone BBS; as negative film for those with access to photographic methods of circuit-board production; and as laser-printed positive images that can be transferred directly to the PC-board material.[11]

Several methods are available for transferring toner from the laser engine to the circuit board. Plain paper is my favorite.[12] Start with a clean circuit board (roughed with 400-grit sandpaper) and a laser-printed reversed positive

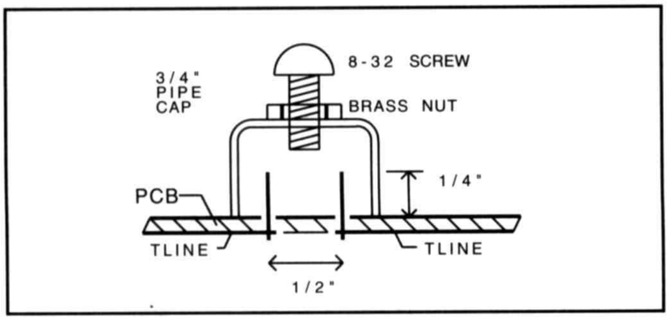


Fig 6—FL1 Pipe-cap filter. The probes are made out of the center conductor and dielectric of UT-141 coax–use the dielectric to help hold the wires in place–the length from the ground plane to the tip is 125 mils. The ground plane is carefully countersunk to avoid shorting to the probes.

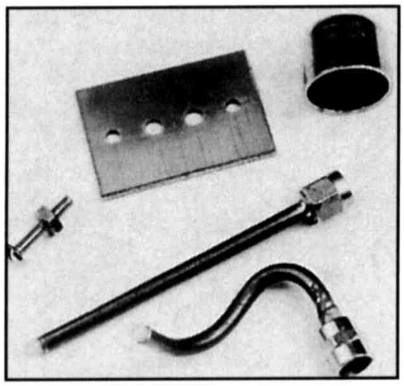


Fig 7—The pieces that make up a 10-GHz band-pass filter, before assembly. *(photos by Kirk Kleinschmidt, NTØZ)*

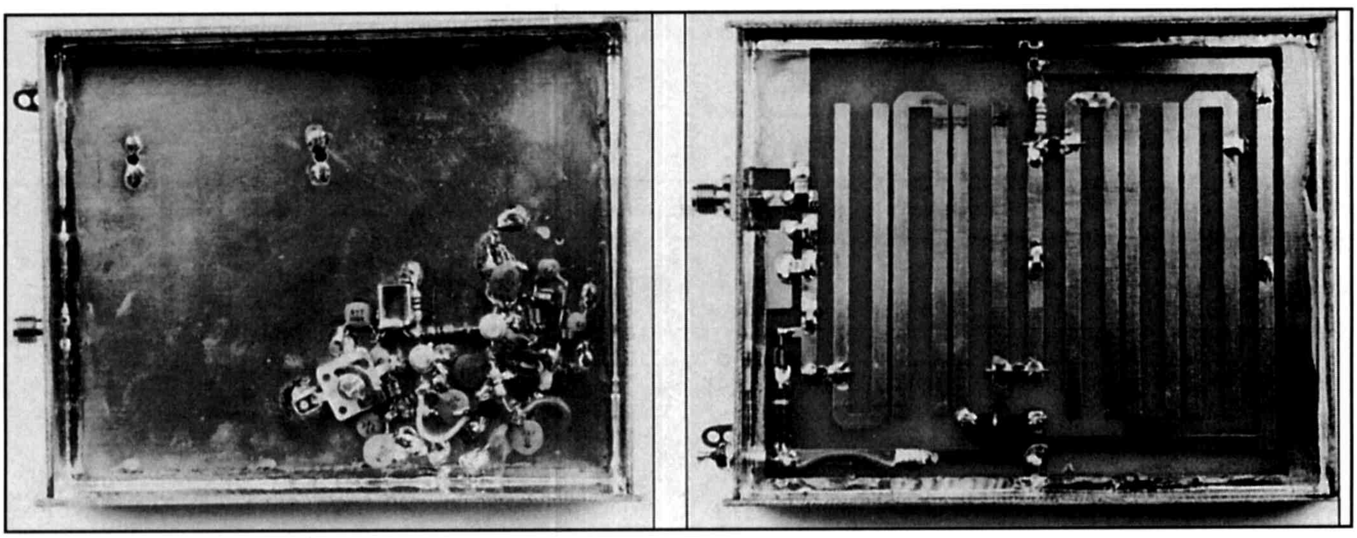


Fig 8—At left, a photo of the 106.5-MHz oscillator circuit, built ground-plane style on the back of the 639-MHz multiplier board. Oscillators built this way exhibit stability an order of magnitude better than etched PC-board versions. This is especially important for minimizing drift at the 96th harmonic of the oscillator frequency—10.224 GHz. At right, a top-side view of the 3¾- × 4¾-inch assembly shows the 639-MHz etched band-pass filters.

image of the board on plain paper. Then use an ordinary household iron at its linen setting to iron the image onto the board. Buffer the iron from the sheet of paper with the pattern on it with a second, clean sheet of paper. Run the iron over the board in a pattern that uniformly heats the material for 30 seconds or so for the 15-mil Teflon boards and at least a minute for the G10/FR4 boards. The iron's heat liquefies the plasticized toner and fuses it to the circuit board.

After ironing, place the board and paper (now fused to the board) into plain water for a few minutes, then remove it from the water and carefully rub away as much of the paper as you can. If the transfer process leaves incomplete traces, clean the board again with sandpaper and start over with a new copy of the artwork. You can correct minor imperfections with an etch-resist pen and carefully cut pieces of Scotch tape. Cover the bottom (ground-plane) side of each board with Scotch tape, then etch the boards. Peel off the tape and remove the toner with plain steel wool.

Oscillator Construction

I didn't develop circuit-board artwork for the 106.5-MHz local oscillator. If you want stability adequate for a 10-GHz SSB/CW system, a quartz-crystal-controlled system is marginal—you really can't throw away any stability to make construction easier. Remember: *The LO is multiplied by 96 before being mixed with the 144-MHz IF signal!* You could use a double-sided circuit-board layout, except that stability is 10 times worse than that of a ground-plane version. So, I opted for the ground-plane version (Fig 8A). I also used a high-stability, air-dielectric trimmer at C1, as some ceramic trimmers have a high temperature coefficient. The trimmer value isn't critical, as L1 can be adjusted to compensate.

I recommend that you build and align the oscillator as follows. Build the oscillator with a 47-Ω, ¼-watt resistor in place of the inductor/crystal combination (L2 and Y1). When you power up the circuit, tune C1 so that the oscillator operates at 106.5 MHz. After replacing the 47-Ω resistor with the crystal and its resonating inductor, verify that the oscillator starts reliably as power is applied. A minor adjustment of C1 may be necessary for reliable starting. I don't recommend trying to adjust C1 for a given oscillation frequency.

The 639-MHz to 2.556-GHz multiplier (Fig 9) has no tuning adjustments. You simply verify that its power output is between 5 and 10 dBm.

Amplifier Design

I chose to use MGF1302 GaAsFETs for all the 10-GHz circuits. These seem to be the most readily available, low-cost parts that work well at this frequency. The transverter uses seven of them, and they cost less than $7 each from several

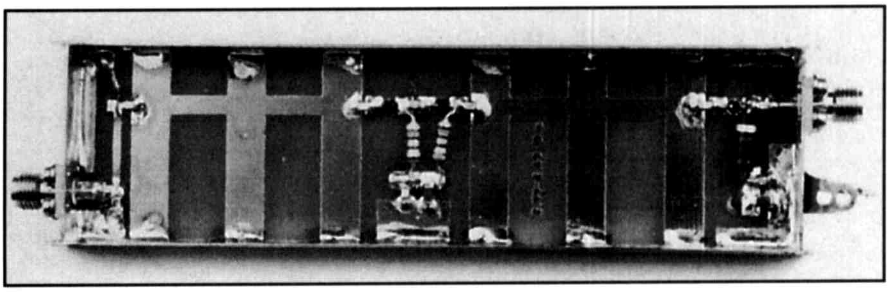


Fig 9—The 1¼ × 5-inch 639-MHz to 2.556-GHz multiplier board uses MMICs to provide a 5-mW filtered intermediate LO signal that drives the 10.224-GHz multiplier.

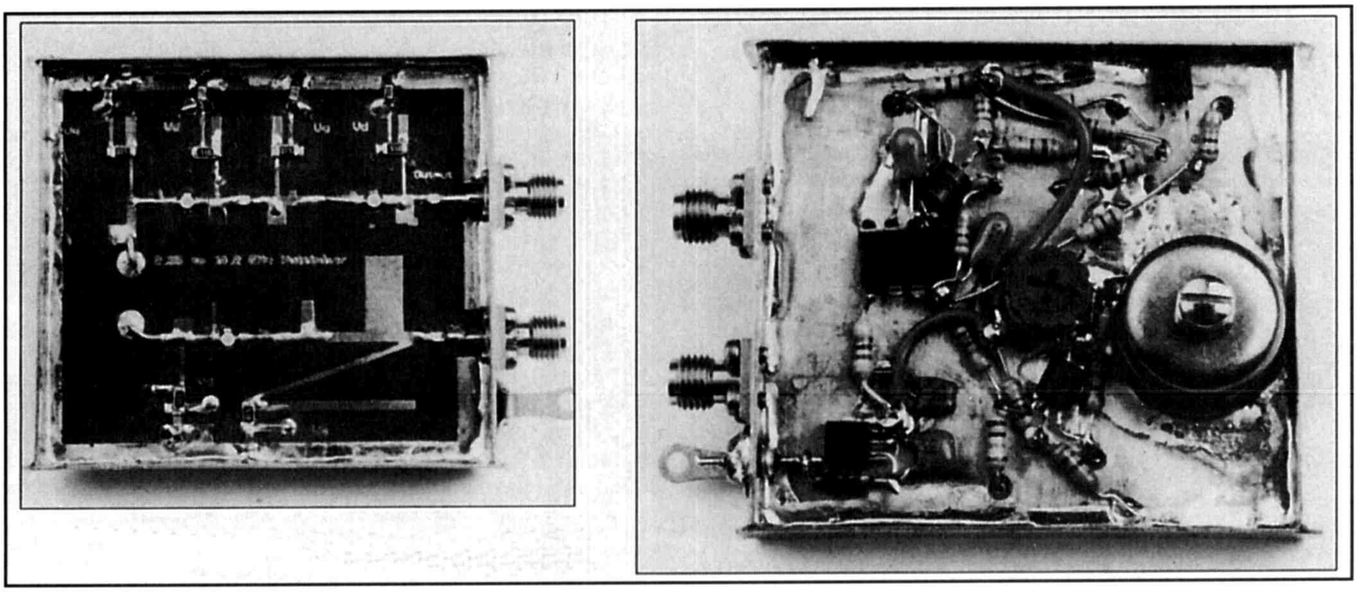


Fig 10—At left, the business end of the 2⅛- × 2⅜-inch 2.556- to 10.224-GHz multiplier module. At its far left, the UT-141 probes couple to the pipe-cap filter shown at right. The FET biasing circuits are also shown at right.

sources. Ideally, a transverter like this would use 10-GHz MMICs for gain blocks, but these weren't available during project development. Not only were the available packaged GaAs MMICs too pricey (around $40 each), but they weren't designed to work at 10 GHz. The second choice was the Avantek ATF13735, but commercial purchasers have made the standard part the short-leaded ATF13736, which is more difficult to use than the long-leaded version. I prefer to use devices with long leads since they're easier to install in circuits that use lead inductance as a circuit component.

Multiplier and Amplifier Construction

In each of the transverter's building blocks, I build the RF circuitry on one side of the ground plane and the biasing circuitry on the other. After etching the boards, I drill and countersink holes for the power leads. Also countersink the ground-plane foil around the multiplier board's filter-cable holes so that the UT-141 center conductor doesn't short to the ground plane. Countersink the holes *by hand* with a relatively large drill (3/16 to 1/4 inch). It's important to do this manually—you'll be surprised how easy it is to drill through such thin, soft material! Cut slots for the FET source leads as discussed in the next section. Then add the brass walls and install the connectors. Build the biasing circuitry after the transistors are installed.

Circuit performance may be improved slightly, as discussed in the next section, though the design is relatively broadband and should operate adequately despite minor construction variations. Computer simulations predict gain flatness within a decibel across the 10-GHz band.

I glue RF-absorptive rubber or foam to the insides of the enclosure lids. This reduces the chance of waveguide effects disrupting circuit operation.[13]

GaAsFET Installation Tips

Beware of soldering irons with significant ac leakage. People blow up lots of devices because their soldering iron tips aren't at ground potential. Measure your soldering iron's tip-to-ground potential if you have any doubts.

The circuits in this transverter use the GaAsFET source-lead inductance as a circuit component. Use the photos as guides when installing them. Bend the source leads down at the ceramic device body, then insert them into holes carefully cut in the circuit boards using a #1 X-ACTO blade or similar weapon, as is done in *The ARRL Handbook*'s GaAsFET preamplifiers.[14] Be sure to cut the holes so that the device is centered on the board traces. Once the device is installed, bend the source leads up flush with the bottom of the board and solder them to it.

Of course, take the usual precautions when handling GaAsFETs, which are static-sensitive. Chapter 24 of *The ARRL Handbook* discusses these practices.

Adjusting the 2.556- to 10.224-GHz Multiplier

First, adjust the filter-tuning screw for maximum output. Next, set the bias trimmer for maximum power output. You may then want to tune the amplifiers. Do this using a tuning tool made out of a 1/16-inch-square piece of thin copper sheet or foil stuck into the end of a piece of Teflon tubing. Slide the tool along the input and output lines, looking for hot spots—places where the presence of the foil makes the power output increase. After finding them, turn off the power. Next, solder a piece of foil at each hot spot and adjust its position with high-quality tweezers.

Coming in Part 2

When you finish building the blocks described here, you'll have a clean 10.224-GHz local oscillator. Next, I'll describe the mixer/splitter board and the preamplifier/power amplifier circuit, and some 10-GHz antenna ideas.

Notes

[1] It *may* be possible to tune up the system using a Gunnplexer unit with an S meter, but I haven't attempted it and can't guarantee that it will work. A Gunnplexer should be able to pick up a properly functioning LO even with lots of attenuation between them.

[2] *QST* has published a series of no-tune transverters developed by KK7B and WA8NLC, including versions for 903, 2304, 3456 and 5760 MHz. The most recent of these is J. Davey, "A No-Tune Transverter for the 2304-MHz Band," *QST*, Dec 1992, pp 33-39. See the notes at the end of D. Mascaro, "A High-Performance UHF and Microwave System Primer," *QST*, May 1991, pp 30-33, for details on the others.

[3] Z. Lau, "The Quest for 1 dB NF on 10 GHz," RF, *QEX*, Dec 1992, pp 16-17.

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

[5] K. Bailey, R. Larkin and G. Oliver, "TNT for 10 GHz," *Proceedings of Microwave Update '88* (Newington: ARRL, 1988), pp 80-95.

[6] Amplifier stability is affected by the inductance of the FET source leads.

[7] I've also used little surplus boards with surface-mount NE555s, although some of these are poorly constructed and had to be resoldered for reliable operation.

[8] Z. Lau, "Power Supply for GaAsFET Amplifier," *QEX*, Mar 1991, pp 10-11.

[9] K. Britain, "Cheap Microwave Filters," *Proceedings of Microwave Update '88*, pp 158-162.

[10] G. Elmore, "A Simple and Effective Filter for the 10-GHz Band," *QEX*, Jul 1987, pp 3-5, 15.

[11] The ARRL BBS can be reached at 860-594-0306 (1200/2400, N, 8, 1); one 250-kbyte file, KH6CP10G.ZIP, contains all the PC-board artwork. Send paper and film artwork requests to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Request the MAY 1993 *QST* KH6CP 10-GHz TEMPLATE and be sure to indicate whether you need paper or film artwork. The template package also includes part-placement diagrams for the transverter's circuit boards. There is a charge of $2 for ARRL members and $4 for nonmembers. Please enclose a check made out to ARRL.

[12] J. Grebenkemper, "Ironing Out Your Own Printed Circuit Board," *QST*, July 1993, pp 42-44.

[13] K. Britain, "Works Great! Until You put it in the Box?" *Proceedings of the 25th Conference of the Central States VHF Society* (Newington: ARRL, 1991), pp 33-34. See Note 5 for ordering information.

[14] L. Wolfgang, ed, *The ARRL Handbook for Radio Amateurs*, 1993 ed (Newington: ARRL, 1992), pp 32-22 through 32-38.

Home-Brewing a 10 GHz SSB/CW Transverter

Part 2—Designed to work with last month's 10.224-GHz local oscillator, this month's mixer, power amplifier and preamplifier round out your narrowband 10-GHz transverter.

By Zack Lau, W1VT

(From *QST*, June 1993)

In Part 1, I described a 10.224-GHz local oscillator (LO) designed to drive a dual-mixer board like those used in the no-tune transverters. If you've completed the modules described in the first section, you should have a working 10.224-GHz LO. The mixer board described this month contains a two-way etched power splitter that delivers equal LO signals to the transmit and receive mixers, which are also etched on the same PC board. On transmit, one of the mixers combines a 144-MHz IF signal with the LO to generate a 10.368-GHz signal; on receive, the other mixer combines the incoming 10.368-GHz signal with the LO to produce a 144-MHz IF output. An external pipe-cap filter (described last month) in each 10-GHz mixer line eliminates the image, passing only the desired signal. Two-stage GaAsFET amplifiers of the same RF design, but using different bias settings, serve as a 10-GHz preamplifier and power amplifier.

The first section also shows the transverter block diagram, and covers construction techniques and etching-pattern availability for the transverter's circuit boards, component sources, and performance data for the finished transverter.

Mixer Construction and Tweaking

If you're building transverters from surplus hardware, the most difficult module to obtain is not the LO, but the mixer. Builders have gotten widely varying results, even when copying the same design. For most people, 10 GHz is just too high a frequency to accurately build a no-tune mixer that works well. The difficulty is that a full-wavelength microstrip transmission line is only 0.6 inch long at 10 GHz. So, a typical rat-race mixer (which requires signals to be 180° out of phase for proper cancellation) really needs to built with tolerances under 0.005 inch (5 mils).

This problem has several solutions. One is to simply accept the inferior performance. Usually, the conversion loss isn't too bad if you copy a known-good layout, but the LO rejection relative to the PEP output signal can be as little as 10 dB. For receive purposes, LO rejection really doesn't make much difference.

A better solution is to tune the mixer. Once you've etched and assembled the mixer board, terminate all ports in 50-Ω loads or sources. You don't want to look at the mixer through an image-reject filter, unless it is properly tuned. Otherwise, the mixer and filter tuning will interact, making it difficult to adjust the mixer for proper operation. I normally connect the mixer to the LO, attach a 0- to −10-dBm, 50-Ω 144-MHz source at the IF port, and a spectrum analyzer at the RF port.

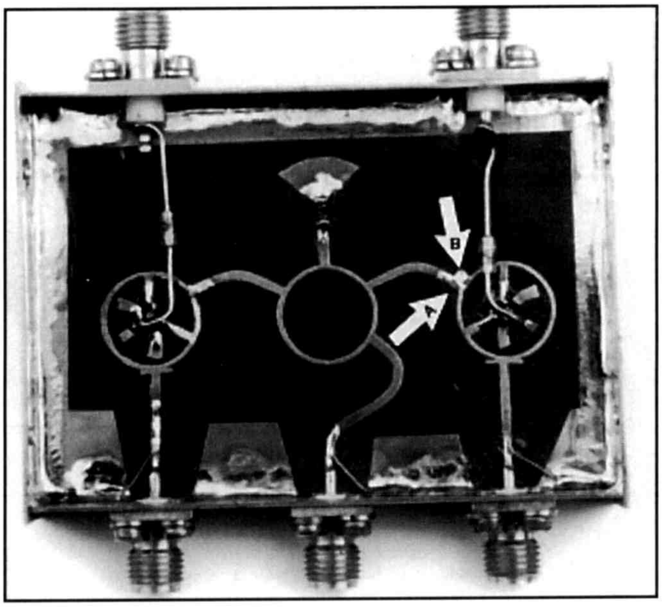


Fig 11—The splitter/mixer board. The IF ports (at top) are connected to the mixers via 0.01-μF encapsulated chip capacitors. Ceramic capacitors can be used instead. The mixers use Hewlett-Packard HSMS 8202 Ku-band diode pairs. The HP HSMS 2822 2-GHz diode pair can be substituted, but this device has more conversion loss at 10 GHz than the HSMS 8202. Mixer tuning for best LO rejection is done by adding a small piece of foil at point A or point B. *(photos by Kirk Kleinschmidt, NT0Z)*

I usually adjust the LO rejection first. This is done by placing a small piece of copper foil at point A or B indicated in Fig 11—at either side of the junction between the 70-Ω ring and the LO-input line. This shortens the transmission line slightly on one side. Usually, the LO suppression improves with the copper at one point and worsens with it at the other point. True, the copper foil mismatches the amplitude slightly, but this is better than having an improper phase shift. Usually, LO rejection is 17 or 18 dB below the saturated output (this equates to the specification-sheet figure of 27 or 28 dB of LO-to-RF port isolation). Keep in mind that even a lid covered with absorptive rubber or foam affects the tuning slightly. You don't want to tune the mixer to perfection only to have to retune it after installing a cover.

I find that the obtainable LO rejection depends on how well I made the board. Mixer rings that look almost perfect often allow 5 or 10 dB better rejection; ones that look as if they were drawn quickly with a crayon may be almost impossible to tune (though they often work just fine for receive).

Finally, tune the mixer's RF port for maximum output into a 50-Ω load (as described in Part 1 under "Adjusting the 2.556- to 10.224-GHz Multiplier"). I've been unable to etch mixers consistently, so all of my mixers are a little different.

Three short wires, 0.21 inch of #28 enameled wire, serve as 10-GHz RF chokes and 144-MHz shunts at the mixer board's RF and LO inputs (Fig 11).[15] This improves the isolation between the mixer's IF ports. Without them, there is little to stop a 2-meter signal from crossing the power divider. Adding these wires increases the isolation between the IF ports from an almost negligible 4 dB to a decent 40 dB.

It shouldn't be necessary to tune the load termination, though you may want to. As you might guess from the layout, I tacked on the radial stub to ground the 51-Ω chip resistor. Purists may want to use a 68-Ω resistor and tune out the reactance to get a really good 50-Ω load at 10 GHz (as is done in the TNT). If you have them, you can also use 50-Ω microstrip terminations for this; I've gotten them from surplus isolators.

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

10-GHz Power Measurement

Measuring RF power at 10 GHz presents a challenge; calibrated measurement devices can be very expensive. Fortunately, measuring *relative* power requires only a diode detector and a sensitive dc voltmeter. The 10-GHz power measurements required to optimize this transverter needn't be absolute; relative power measurement is acceptable. A convenient way to measure power in a 50-Ω system is to couple some RF into a low-offset Schottky diode, such as a Hewlett-Packard 5082-2835 (commonly used as a microwave frequency multiplier), filter its dc output, and measure this voltage with a high-impedance voltmeter. This measurement approach gives useful output down to the milliwatt level. Of course, you can also use a commercial diode detector rated to 10 GHz.

To build a detector, etch or cut a 50-Ω microstripline on a small piece of Rogers 5880 RT/duroid with 1-oz copper cladding (the same material used in the transverter's 10-GHz circuits). See Fig A. A 50-Ω trace is 46 mils wide (0.046 inch) on this material. Terminate the microstrip in SMA connectors and enclose the board with brass strip for rigidity, like the transverter. Mount the diode and other components as shown in Fig A. Fig B shows the equivalent circuit. The length of the diode lead that runs along the 50-Ω stripline affects the amount of RF energy coupled into the diode, as does its spacing from the microstrip trace.

This detector can be used for tuning the transverter's multiplier, filters and amplifiers. To use the detector, terminate one end in a 50-Ω load that's good to 10 GHz. (Alternatively, you can substitute a 50-Ω microstrip load for one of the SMA connectors.) Couple RF into the other port via a 3- to 10-dB attenuator, to ensure that the circuit under test is terminated with a stable 50-Ω load. Measure the voltage on the feed-through capacitor using a sensitive voltmeter or oscilloscope.—*Kent Britain, WA5VJB*

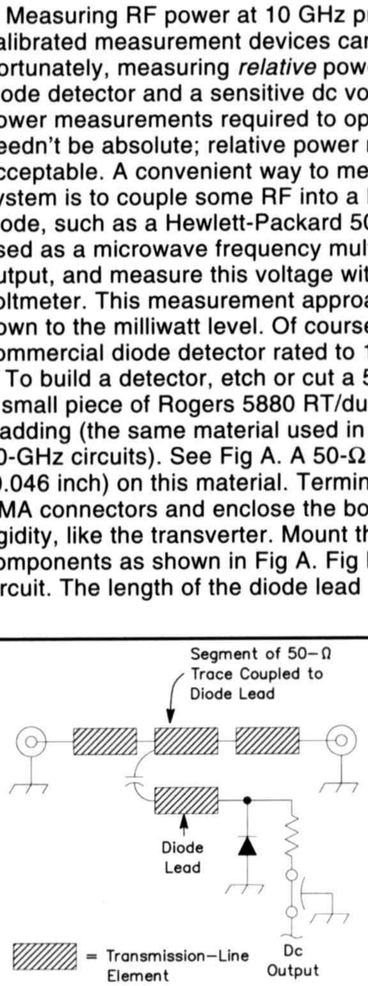


Fig B—Equivalent circuit of the diode detector.

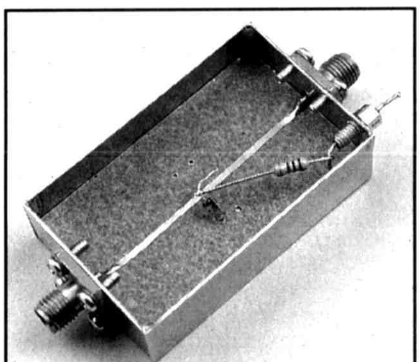


Fig A—This 2- × 1¼-inch diode detector gives useful dc output for 10-GHz power measurement down to about 1 mW. It uses a low-offset Schottky diode (such as the HP 5082-2835), with its anode lead soldered to the ground plane. Its cathode lead follows the 50-Ω microstrip trace for about ¼ inch and is spaced about ⅛ inch from the trace (neither dimension is critical; a longer lead and closer spacing increase coupling). A 1- to 10-kΩ resistor, also soldered to the cathode lead, routes rectified energy to a feedthrough capacitor.

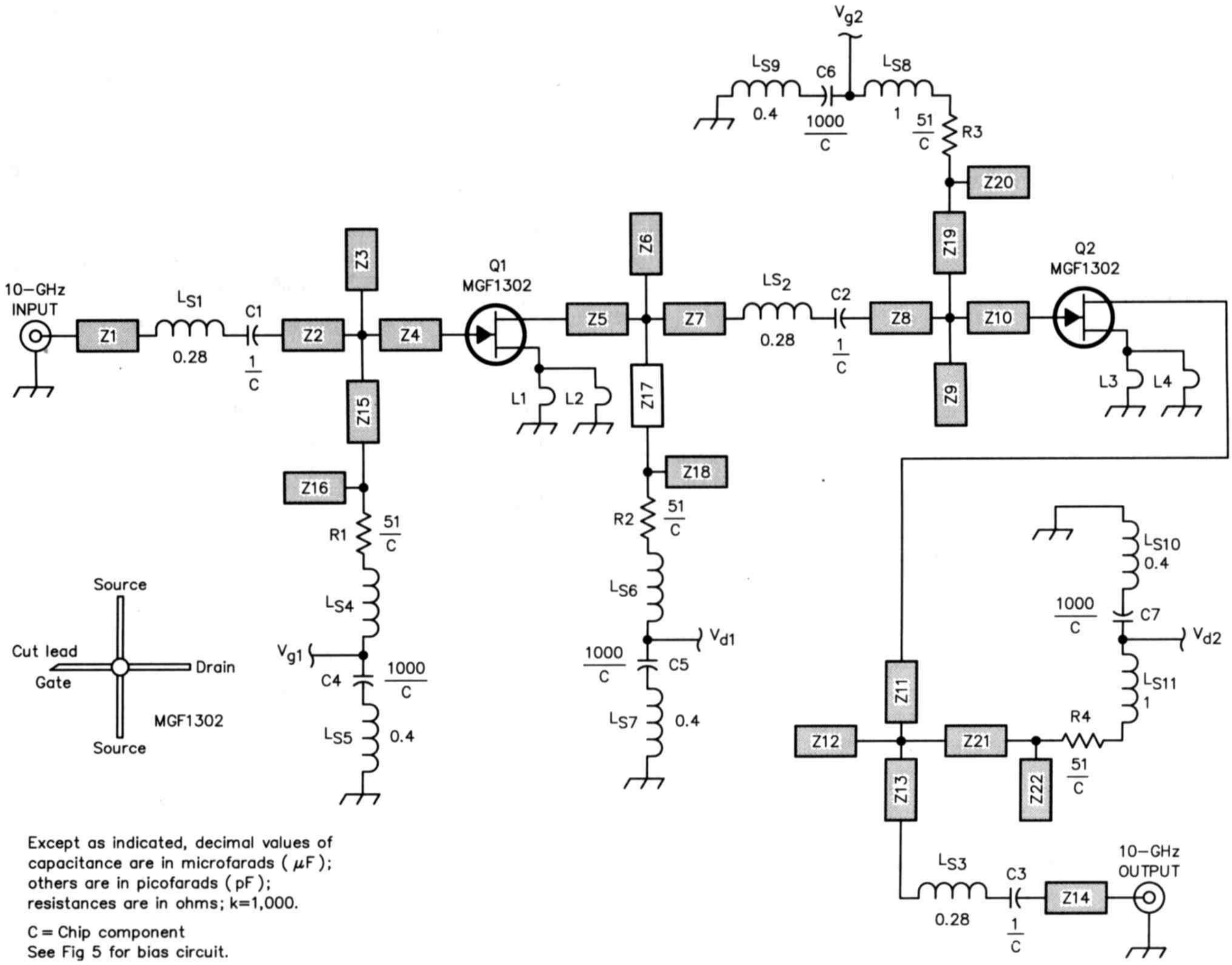


Fig 12—10-GHz low-noise amplifier. Chip components are used unless specified otherwise. L1-L4 are source lead inductances. L_{S1}-L_{S11} are stray inductances (in nanohenries). Z1-Z22 are etched on the circuit board.

C1-C3—1 pF. Use high-quality, 50-mil ceramic chip capacitors such as ATC 100As.

Q1, Q2—MGF1302. Substitution not recommended. Set bias at 10 mA and V_{ds} = 3.0 V for low-noise receive preamplifier operation. For the transmit amplifier, set bias at 30 mA and V_{ds} = 3 V. For additional biasing information, see the text and Fig 5.

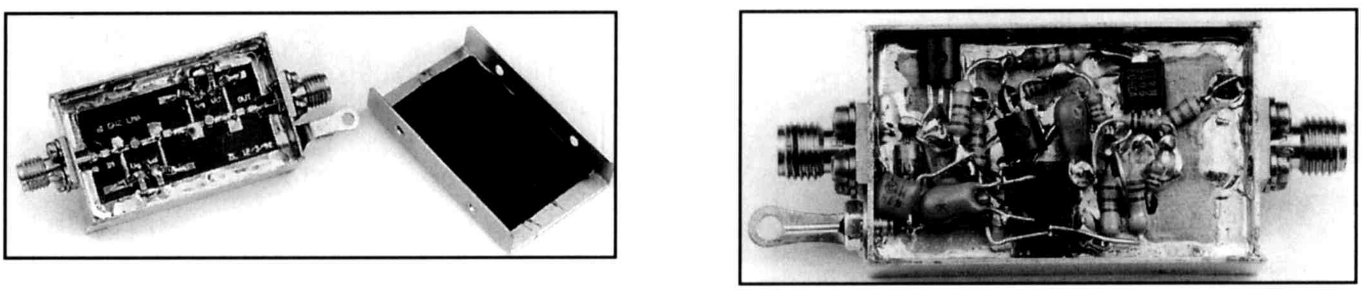


Fig 13—Top and bottom-side views of the completed 10-GHz two-stage amplifier. Microwave-absorptive foam is glued into the lid to suppress waveguide propagation modes inside the amplifier enclosure, which could provide enough feedback from output to input to cause oscillation. The amplifier's bias circuitry, like that of the 10.224-GHz multiplier (Figs 4, 5 and 10) is built on the bottom side of the amplifier board for convenience. External biasing is also acceptable and may be more convenient.

Power Amplifier and Preamplifier

The same RF design is used for the transmit and receive amplifiers (see Figs 12 and 13). Tripling the bias current from the 10 mA used in the receive-side amplifier to 30 mA in the transmit amplifier increases the circuit's 1-dB compression point from 5.7 to 10.8 dBm. Gain increases from 18 to 19 dB and the noise figure rises by about 1 dB. Computer-modeling results indicate that the Rollett stability factor, K, drops a little, but since it's still above 3 (a K greater than one denotes a stable design), this shouldn't be a problem—even if the amplifier is terminated at the input and output with a high mismatch (such as sharp filters).

It may be possible to get a bit more output by increasing V_{ds} to slightly more than the 3 volts I used, but this would require redesigning the bias circuit (Fig 5). Like many transistor amplifiers, this amplifier's saturated output, typically 14 dBm, is more than twice the recommended output for linear operation.

System Integration

To complete the transverter, build two band-pass filters as shown in the first section. You can tune them with the aid of the diode detector described in the sidebar, but a few minutes with a spectrum analyzer makes the process easier.[16] Then, following the block diagram of Fig 1, assemble the transverter's blocks. Connect a suitable IF radio, check to make sure the transmit converter and receive converter operate, and you're on the air!

Antenna Thoughts

Most people looking for a high-gain antenna end up with some sort of parabolic reflector. If you put a low-gain horn antenna in the right spot in front of a reflector that is anything close to a parabola, it will probably outperform anything of similar dimensions on this band. People have used everything from metal snow sleds to trash-can covers, in addition to more obvious choices such as light collectors and surplus military/commercial gear. Commercial sources for new dishes exist, but even small dishes are expensive when purchased new. Look for a surplus reflector.

Perhaps the simplest antenna I've seen is a quarter-wave monopole—with a piece of sheet metal as the ground plane! The most complicated is undoubtedly a loop Yagi—it works, but it is more of a curiosity than a practical way of getting 18 to 20 dBi of gain. A horn is much easier to make. *The ARRL Antenna Book*, *The ARRL UHF/Microwave Experimenter's Manual*, the RSGB *Microwave Handbook, Volume 3* and various VHF/UHF/microwave conference proceedings contain duplicable designs. Chapter 18 of the RSGB *Microwave Handbook, Volume 3*, contains all the information you need to get started.

I have yet to adjust one of my 10-GHz antennas with an SWR meter, yet I've made lots of 10-GHz contacts of more than 200 km. Usually, if I do any tweaking at all, such as adjusting the location of a dish feed, it has been for maximum received signal. Similarly, I've adjusted my coax-to-waveguide transitions this way, adjusting tuning screws for minimum loss. Of course, even SWR is no indication of how well an antenna really works. The *real* test is to compare antennas and see which does best.

Summary

Although it takes some effort to build, the transverter described in this two-part article provides useful and exciting 10-GHz SSB and CW capability. Perhaps the best part is that you don't have to hunt through flea markets to find a surplus "brick" LO and filters, or deal with any of the other traditional hassles of getting on this fun band! What hilltop will *you* operate from in this year's ARRL 10-GHz Cumulative Contest in August and September?

Notes

[15]Part-placement diagrams with component values and more detail for each of the transverter's modules are part of the template package obtainable from the ARRL Technical Department Secretary. See Note 11, page 2-32 for details.

[16]When tuning these filters, you can use the finished 10.224-GHz LO and a power meter (see the sidebar) to make sure that the filters aren't tuned to the LO or image frequency. To tune a filter, first connect it to the LO and adjust the tuning screw for maximum output at the LO frequency. Then adjust it for peak response at 10.368 GHz by connecting it to the transmit-amplifier output and backing the tuning screw *farther out of the filter cavity* while looking for maximum filter response.—*Ed.*

An Image-Phasing Transverter for 10.368 GHz

By Doug McGarrett, WA2SAY

(From *The 22nd Eastern VHF/UHF Conference*)

When all the local 10-X group started talking about using a 2-meter SSB input to their 10 GHz transverters, *and* using a separate 2-meter sideband radio for liaison, it started to sound expensive. In addition, when you put your power amplifier on a standard mixer, you waste half of your capability amplifying your image, unless you build a filter at the RF output frequency and insert it before the power amplifier. Why not use brains, instead of brawn, I thought.

The image-phasing mixer is a technology which goes back to the early days of single sideband radio.[1] Back when the Central Electronics 10-B was the way most people got on SSB, that was the technology they used. So there's nothing new here. Just a bit higher in frequency, that's all.

The magic of the image phasing mixer is that, as an up-converter, you need no filter to get rid of your image frequency. And the usual double balanced mixers suppress the local oscillator pretty well also. As a down-converter, you again need no preselector to keep out the image frequency noise. Typical image rejection is 30 dB or better for narrowband operation.

The basic device is shown in Figure 1, along with an HF transceiver, a local oscillator source, and an antenna, as well as incidental equipment. This is not going to net you any tremendous DX, but it has made a 2-way QSO over a 24 mile obstructed path! Think what it will do when it has RF amplifiers going in both directions between it and the antenna! The RF amplifiers will be a receiving preamp with about 1.5 dB noise figure and 15 dB or so gain, and a transmitting amplifier with better than 1 watt output and about 25 dB gain. Add microwave relays, a dish antenna, and stir.

Meanwhile, the mixer: In order to achieve image phasing, you need 90° phase shifters at one of the RF ports (not both) and the IF port. The other RF port must be fed either in phase or 180° out of phase. Fortunately, the devices to do this are fairly commonly available on the surplus market. (Except for the IF hybrid, which can be home-brewed.) My IPM is made up of two WJ (Watkins-Johnson) "MinPac" M80D double balanced mixers, fed with an Omni-Spectra 2032-4096-00 quadrature hybrid, and a Narda 4315-2 Wilkinson hybrid. All of this is hooked together as directly as possible. One of the hybrids has male connectors on one side, the other is connected using male-male SMA adapters.

I discovered (by Murphy's Law) that you have to have the quadrature hybrid on the antenna side. Otherwise, you generate upper sideband, but receive lower sideband. (Note that what I mean by "sideband" in this application is the frequency relationship between the local oscillator and the signal at the antenna port.) I have not done the vector analysis to determine why this happens, but I can testify that it does! (Later research indicates that the quadrature hybrid is always shown connected at the antenna end of the mixer. He who does not study history is doomed to repeat it, they say.)

The IF hybrid was made using a one-to-one transformer wound on a small ferrite toroid, with capacitor coupling across the input to output connections at top and bottom (Figure 2). If this is done right, you wind up with a 3-dB hybrid directional coupler with the classic 90° phase shift between the two output ports. This device can be found in the ham literature, as well as other places.[2]

The performance of the mixer was evaluated for use directly on an antenna, since that is how I knew I would first have it hooked up. With a local oscillator drive signal of +19 dBm, and an IF signal of +10 dBm, I could get +4 dBm out of the mixer with reasonably good linearity. If I didn't care so much about linearity, I could push the RF up to about +15 dBm, and get about +7 or +8 dBm out. That's fine for CW. The image rejection (at +10 dBm IF drive) was 34 dB. The LO rejection was 17 dB. Noise figure with a mediocre IF preamp was 11 dB.

The above was written as a quick report; since then, Bruce Wood, N2LIV, has asked me to expand on it somewhat for the Proceedings of the Eastern VHF/UHF Conference. One of the items not discussed in detail is the IF hybrid. As shown by the reference, the first article to discuss this, as far as I'm aware, was by Reed Fisher,

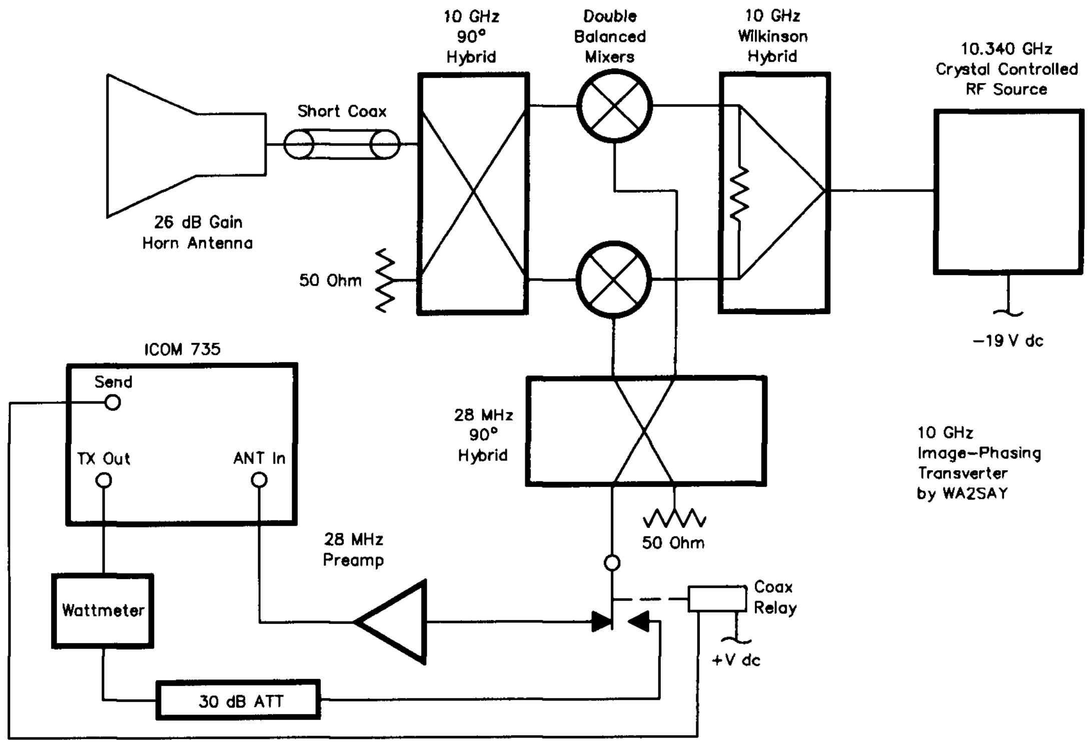


Figure 1

W2CQH. He followed the IRE paper with an article which appeared in *Ham Radio*, I believe, but my copy was loaned out to someone who needed it more than I did. (That someone also got my copies of Norgaard's and Dome's papers on single sideband and phasing networks.)

The trick is to wind a very tightly coupled twisted pair of wires on a small toroid form, in a fashion similar to that of Ruthroff's transformer.[3] In this case, the twisted pair transmission line which results will be terminated in a capacitor at each end. The design is obtained by means of these equations:

$$Z_0^2 = L/C$$

where Z_0 is the impedance at each port of the coupler, L is the inductance of one winding on the core, and C is twice the value of the terminating capacitors. (See Figure 2.)

The frequency at which the device operates with 3 dB coupling is determined by setting

$$\omega L = Z_0 \text{ or, } L = Z_0/2\pi f$$

and the terminating capacitance, C/2, is determined by

$$C/2 = 1/(4\pi f Z_0)$$

The toroidal core should be a material of relatively low permeability, such as FaiRite type 67 material. I optimized the transformer design using EEsof Libra for Windows, and then tweaked using a network analyzer. In the end, the phase shift was just shy of 90°, so I cheated and added about an inch of RG188/U cable to one port.

It is possible to buy the if hybrid, if you choose the right frequency. Both Toko and Mini-Circuits sell 90° hybrids ready made. I don't know what the cost of the Toko version is, since I can't find a source who will sell in "onesies." Pennstock (a distributor) has them, but they also have a minimum order of $100. Mini-Circuits also has 90° hybrids. The price is pretty steep, but they will sell single units to individuals who live outside New York State. (They can't handle the required in-state sales tax.) The Toko part number for a 28.35 MHz center-frequency hybrid is B4QF-1004. The Mini-Circuits part number for a 23 to 40 MHz hybrid is PSCQ-2-40. If you choose this route, all you have to do is mount the hybrid on a little piece of circuit board and add SMA connectors.

The rejection performance of the image phasing mixer is critically dependent on the equality of the power division and the accuracy of the 90° phase shift. The equation that defines the performance is

$$\text{Image rejection} = (1 + 2\cos\phi + R^2) / (1 - 2\cos\phi + R^2)$$

where ϕ is the phase error from quadrature, and R is the amplitude ratio, in volts.

A simple computer program[4] (in Microsoft or GW

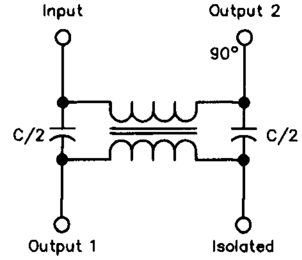


Figure 2

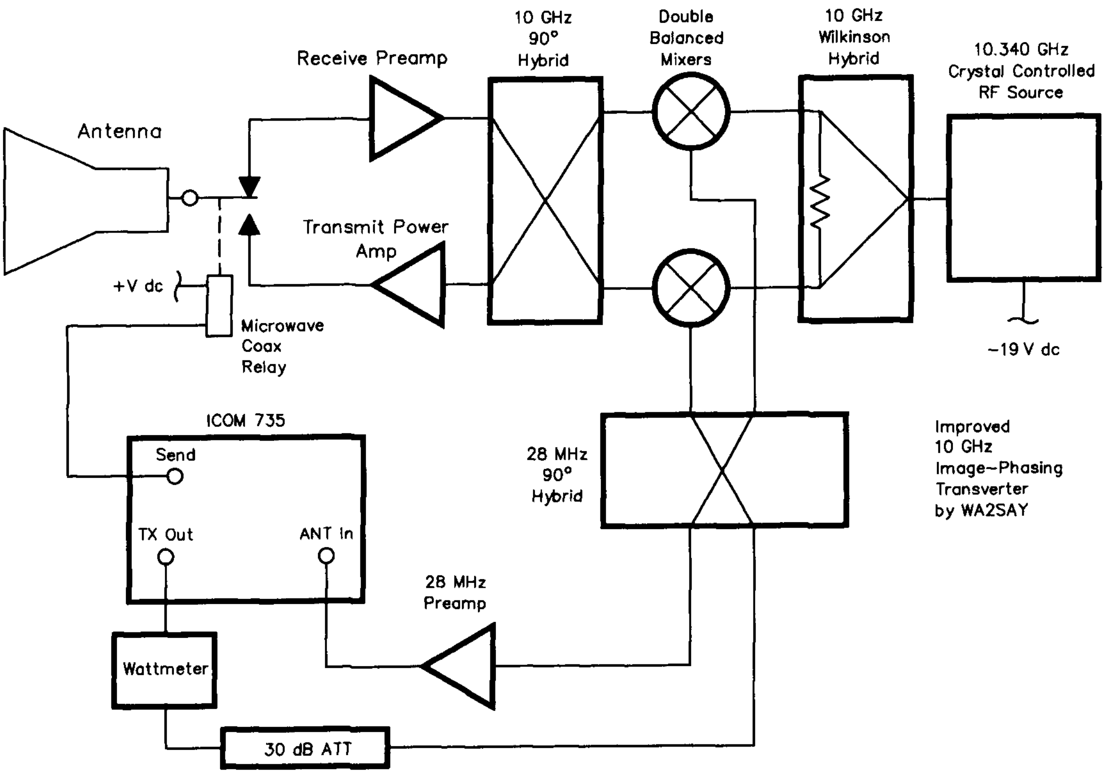


Figure 3

Appendix: IPMREJ.BAS

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100 REM IPMREJ FINDS IMAGE REJECTION
    IN AN IPM
110 REM ACCORDING TO K2RIW :
120 CLS
130 PRINT"IMAGE REJECTION PROGRAM FOR
    AN IMAGE PHASING MIXER"
140 PRINT
150 PI=3.14159265#
160 INPUT"DB DIFFERENCE";D
170 INPUT"PHASE DIFF, DEGREES";DEG
180 REM CONVERT DB TO VOLTAGE RATIO
190 K=10 ^ (-D/20)
200 REM CONVERT DEGREES TO RADIANS SO
    BASIC CAN DO COS ON IT
210 THETA=DEG*PI/180
220 PARTIAL=(2*K/(1+K^2))*COS(THETA)
230 REJ=(1+PARTIAL)/(1-PARTIAL)
240 DBREJ=10*LOG(REJ)/LOG(10)
250 DBREJ=INT(DBREJ*10+.5)/10
260 PRINT
270 PRINT"IMAGE REJECTION IS
    "DBREJ"DB"
280 PRINT
290 END
```

Basic) is provided in an appendix, which will allow analysis of the effect that any combination of amplitude and phase error will have on the performance. (The mixers, of course, must be carefully constructed to meet the necessary criteria. Fortunately, most good microwave balanced mixers designed for your frequency of interest will work well, with a minimum contribution to system error. The insertion loss, however, is directly dependent on the mixer insertion loss characteristics.)

The above discussion does not mention the need for RF gain in both the transmit and receive directions. While it is a simple matter to configure single-pole double-throw RF relays at each end of the mixer, and at the junction of the antenna feed with the RF preamp and the RF power amp a little inspection of the circuit leads me to the conclusion that you can build the whole transverter with only a single coax relay! Look at Figure 3. If you took only one hybrid and used the "other" port, you would get the wrong sideband.

But if you use the "other" port on both hybrids, you should get the same sideband. Since these ports are otherwise isolated, the scheme should work, but I confess that I have not yet tried it.

Afterword

The ICOM 735 radio and the 30 dB pad have been replaced by a modified Radio Shack Realistic all-mode HTX-100 10 Meter radio. RF is taken before the final, which has been disconnected, thru a 13 dB attenuator. The receive IF signal is fed into a new, direct input port to the receiver in the radio.

The form suggested in Figure 3 has been implemented. The RF preamplifier is a modified TVRO LNA-down-converter module, using just the 10 GHz front end portion, for a noise figure of <2 dB and a gain of about 34 dB. A short waveguide section was added at the preamp input, and tapped for 3 brass 0-80 screws, which were adjusted to bring the noise figure down from 3.5 dB.

The RF power amplifier is now an 8-W TWT, satellite surplus. This unit has about 40 dB gain, and a circulator and small pad is used at its input. Note that the RF quadrature hybrid must be terminated in good impedance matches at all ports if this scheme is to work with proper image rejection. The same is true of the IF hybrid; the 13 dB of padding on the 28 MHz transmitter output port is applied right at the hybrid.

Finally, the antenna is now a 30-inch diameter dish, fed by a 3/4-inch circular waterpipe waveguide "shepherd's crook" feed with a Chapparel launcher.

References:

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[2] Fisher, R. E., "Broad-Band Twisted-Wire Quadrature Hybrids," *IEEE Trans. on Microwave Theory and Techniques*, May 1973.

[3] Ruthroff, C. L., "Some Broadband Transformers," *Proc. IRE*, vol 47, Aug. 1959.

[4] Knadle, R., K2RIW, private communication.