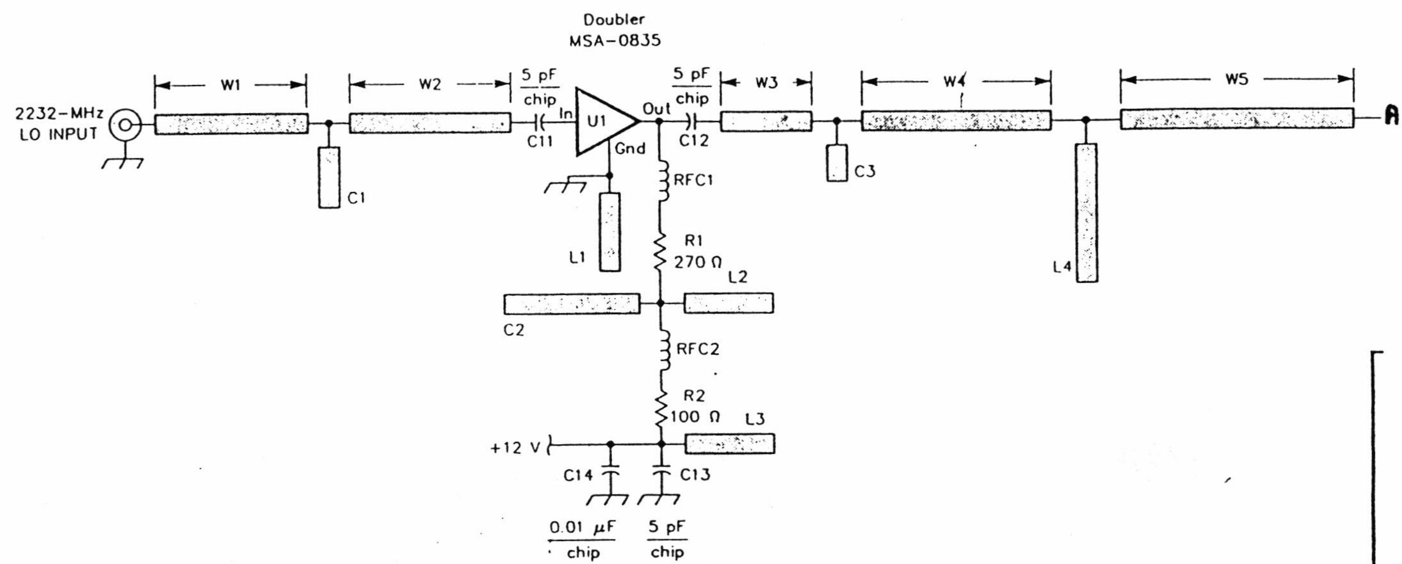
Fig 2—Schematic of the 5760-MHz transverter. R1 and R2 are ¼-W carbon-film types.

C1-C10—Capacitive stub printed on circuit board.
C11-C13—5-pF, 50-mil-square microwave chip capacitor.
C14—0.01-μF chip capacitor.
C15—18-pF microwave chip capacitor.
D1—HP-2822 diode assembly.
L1-L3—1/4-λ open-circuit stub etched on PC board.
L4—1/4-λ stub at 2232 MHz (and 1/2-λ stub at 4464 MHz) etched on PC board (see text).
L5, L6—Broad tuned circuit at 4464 MHz (and short circuit at 1296 MHz) etched on PC board (see text).
RFC1, RFC2—inductance of R1 and R2 leads (not critical).
U1—Avantek MSA-0835 MMIC.
W1-W12—Tuned lengths of 50-Ω microstrip transmission line etched on PC board.



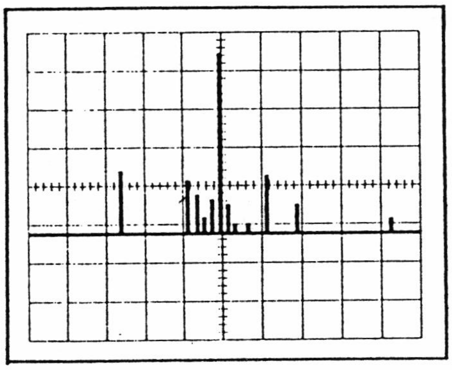


Fig 3—Output spectrum of the 5760-MHz transverter. Each vertical division is 10 dB; each horizontal division is 500 MHz. The pip at the center of the display is the 5760-MHz signal. All other outputs are more than 30 dB down.

about finding a suitable 5760-MHz TR relay; the antenna connects to a common filter for transmit and receive.

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

Design Notes

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

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

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

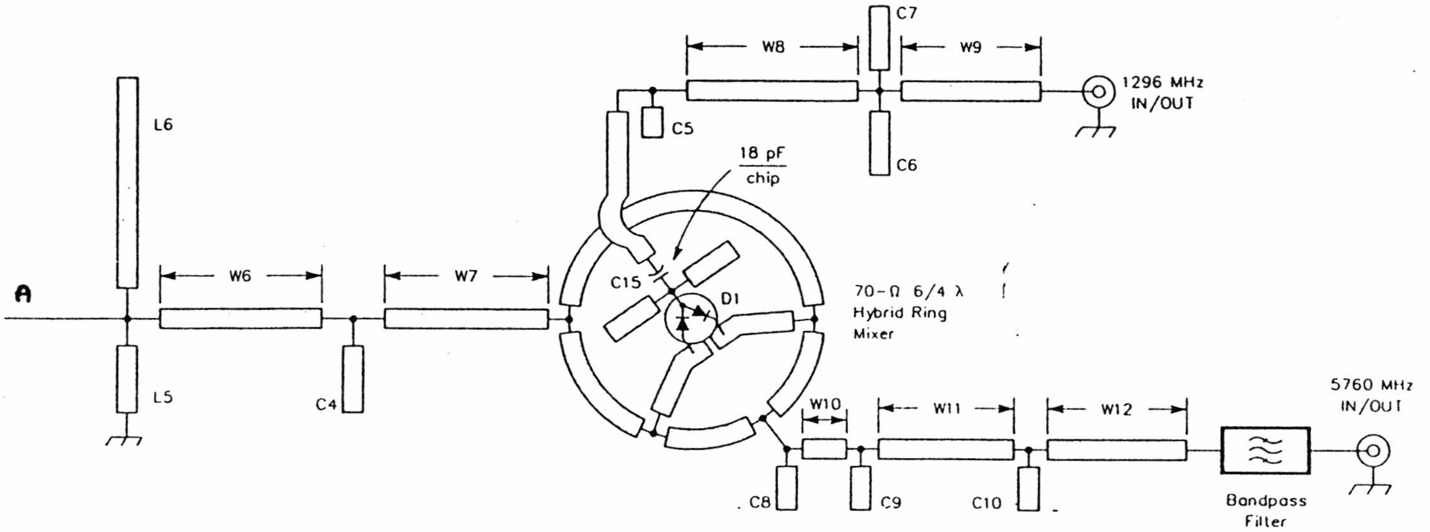
An on-board doubler using an Avantek MSA-0835 MMIC provides an output of +8.5 dBm at 4464 MHz, with all other outputs more than 30 dB down. The second harmonic of the drive signal is obtained by overdriving the MMIC amplifier. The harmonic output is increased by reducing the MMIC bias below the value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

The output of the doubler passes through a 2232-MHz notch filter (L4 on the schematic) and a single-tuned circuit at 4464 MHz (L5 and L6). L4 is an open-circuit 1/4-λ transmission line at 2232 MHz, so it appears as a short circuit at the drive frequency. L4 is an open-circuit 1/2-λ transmission line at 4464 MHz, so it appears as an open circuit at the desired second harmonic. The open circuited end of L4 is one-half wavelength away from the output of the MSA-0835 at 2232 MHz and one wavelength away at 4464 MHz, so the MSA-0835 does not see a short circuit at either frequency.

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

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

The mixer is a standard design, except that the HP-2822 diode pair is used well above its specified frequency range. On the assumption that the imperfections of the diodes were reactive and relatively uniform from part to part, I built a standard 6/4-λ



Microwave Grid Hopping

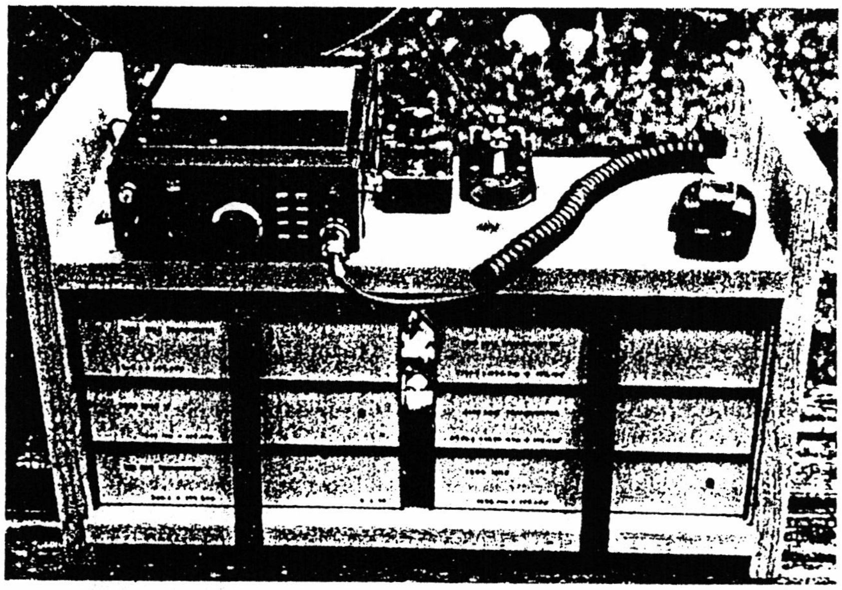
June in Michigan's Keweenaw Peninsula (grid squares EN56, EN57 and EN67) is perfect for any kind of outdoor activity. Previous experience with communicating over 10-mile paths using 19-inch dishes and 3456-MHz mixer-only rigs proved that microwave hilltopping is a great way to spend an afternoon.

With a quick introduction to the "KK7B 9EFGH† Bandswitched Rover" station (it works on 902, 1296, 2304, 3456 and 5760 MHz), Jim, K8OSF, set off in his pickup truck for the wooded hills of EN56 in search of his first microwave contacts. After locating a likely spot with only a few trees (there are trees all over upper Michigan), he connected the 12-V battery and antenna cable to the rig and began listening. For Jim, an experienced Morse operator, making a microwave SSB or CW contact was similar to operating on the HF bands, except that the complete absence of QRM made it easier to dig for weak signals. The only signal on the 5760-MHz band that afternoon was KK7B, a few decibels above the noise. After making a minor adjustment to the tripod-mounted 19-inch dish, the signals came up to 10 dB above the noise, and both K8OSF and KK7B switched to SSB.

The K8OSF hilltop location was 7 miles from my home QTH. There were a few small trees at both ends of the path. The two stations were virtually identical: K8OSF used a camera-tripod-mounted 19-inch dish, a 5760-MHz no-tune transverter board, a no-tune 1296-MHz transverter board for the first IF and an FT-290R for the 2-meter IF transceiver. Everything was the same at KK7B except for the TR-751A 2-meter IF. Neither station used 5760-MHz amplifiers for transmit or receive.

Contacts like this are a good way to get started on the higher microwave bands. As you gain experience, contacts over greater distances will be easy.—*KK7B*

†Each character of this group is an ARRL Contest Branch band designator: 9 = 902, E = 1296, and so on.



The KK7B rover station contains everything you need for low-power hilltopping on 902, 1296, 2304, 3456 and 5760 MHz.

mixer, then added tuning "confetti" empirically to improve the conversion loss. The confetti was then added to the mixer artwork, and subsequent mixers show good uniformity.

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

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

After I integrated the doubler, mixer and filter onto a single board, the interconnec-

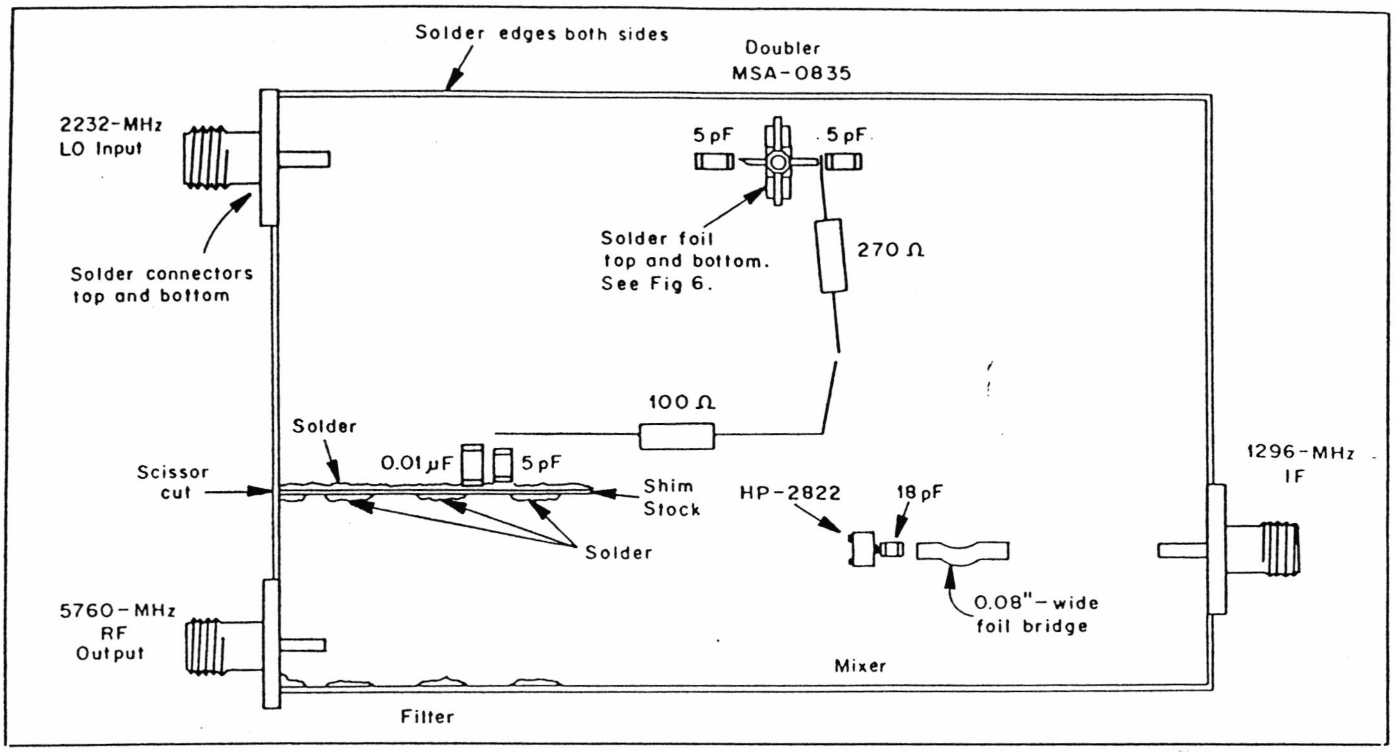


Fig 4—Part-placement diagram for the 5760-MHz transverter board (not shown actual size). All components mount on the etched side of the board. See Fig 6 for mounting details for the MSA-0835 MMIC.

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

One cautionary note: Any metal box big enough to hold the transverter board will be a resonant cavity at a series of frequencies within the gain bandwidth of the MMIC. Imagine building a 40-meter linear amplifier in a box the size of a football field! No-tune transverters and LOs that work perfectly on the bench often exhibit spurious oscillations when enclosed in a metal box. The transverter board will work fine if it is simply wrapped with plastic-bubble packaging material or nonconductive foam and mounted near the antenna feed. For permanent installations, a plastic food container works well.

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

Construction

The transverter is constructed on 0.031-inch-thick Teflon®-glass substrate with a dielectric constant of 2.5. The board is clad with ½-ounce copper on both sides.

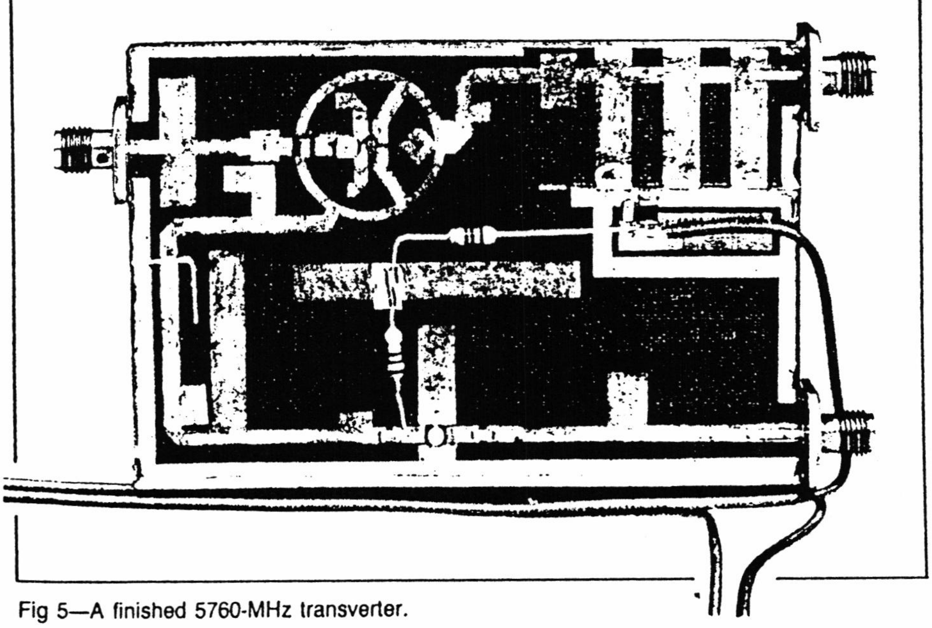


Fig 5—A finished 5760-MHz transverter.

Table 1

Transverter Specifications

General	
Frequency Range	5650-5925 MHz.
IF range	1240-1300 MHz.
LO required	2.18-2.32 GHz at +7 dBm nominal; +3 to +10 dBm acceptable.
Power requirements	12 V dc at 10 mA; 8-15 V acceptable.
Transmitter	
Output power at 1 dB compression	−6 dBm.
Saturated RF output	−4 dBm.
LO signal at RF port	−36 dBm, max.
IF drive level	approx 0 dBm.
Receiver	
Noise figure	9 dB.
Conversion loss	7 dB.
Image rejection	>30 dB.

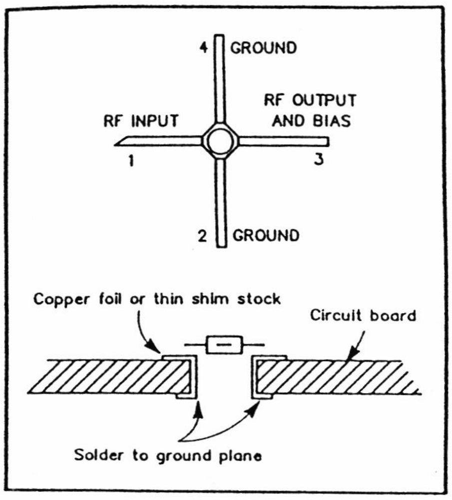


Fig 6—Mounting details for the MSA-0835.

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

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

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

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

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

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

Completing Your 5760-MHz Station

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

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

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

Selecting an antenna for 5760 MHz is easy: Nothing matches the performance of a small dish. I've had good success with a 19-inch dish that mounts easily on a small camera tripod. It shows a 10-degree beamwidth and about 25 dB gain with a simple feed. Don Hilliard, W0PW,[15] and Tom Hill, WA3RMX,[16] have presented some excellent antenna ideas for 5760 MHz.

Conclusion

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

Acknowledgments

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

Notes

[1] R. Campbell and D. Hilliard, "A Single Board 900 MHz Transverter with Printed Bandpass Filters," *Proceedings of Microwave Update '89*, pp 1-8. This book is available from ARRL for $12 (plus $2.50 postage and handling, or $3.50 for insured parcel post or UPS), or from your local dealer.

[2] R. Campbell, "A Single Board No-Tuning 23 cm Transverter," *Proceedings of the 23rd Conference of the Central States VHF Society*, pp 44-52. This book is available from ARRL for $12 (plus $2.50 postage and handling, or $3.50 for insured parcel post or UPS) or from your local dealer.

[3] J. Davey, "A No-Tune Transverter for 3456 MHz," *QST*, Jun 1989, pp 21-26.

[4] J. Davey, "No-Tune Transverter for 2304 MHz," *Proceedings of Microwave Update '89* (ARRL, 1989), pp 30-34. See note 1 for ordering information.

[5] See note 3.

[6] Copies of this article are available from the ARRL Technical Department Secretary for $2 and an SASE.

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

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

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

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

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

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

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

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

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

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

Rick Campbell, KK7B, earned a BS in physics from Seattle Pacific University in 1975, an MSEE from the University of Washington in 1981 and a PhD from the University of Washington in 1984. He is a faculty member at Michigan Technological University in Houghton, Michigan. His research specialty is microwave propagation and scattering in random media, and he teaches electromagnetics and wave propagation courses. Rick was first licensed as WN8VAZ in 1966 and is currently active on SSB and CW on the bands from 144 through 5760 MHz from grid square EN57 in Michigan's upper peninsula. Between VHF contests, he enjoys windsurfing, skiing and playing bluegrass fiddle.

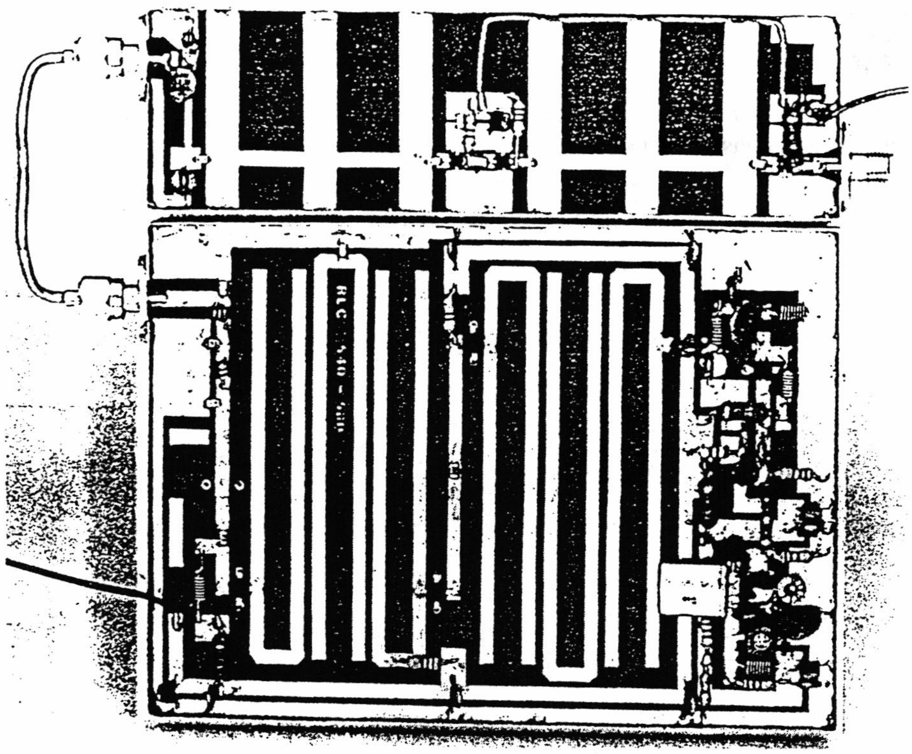
A Clean, Low-Cost Microwave Local Oscillator

By Rick Campbell, KK7B

(From *QST*, July 1989)

NDLR : les copies ne sont pas terribles mais ce sont les seules que j'ai...

Obtaining a suitable local oscillator (LO) has traditionally been a major obstacle in amateur microwave work. This article describes a straightforward, inexpensive, easily constructed microwave LO. The oscillator may be combined with a simple low-noise preamp[1], an image filter[2] and an off-the-shelf doubly balanced mixer to build a complete high-performance receiving converter for OSCAR Mode S. This LO can also be used as a building block in a transverter for the 2304- or 3456-MHz bands.



The complete microwave LO is built on two PC boards. The larger (bottom) board provides a signal anywhere from 540 to 580 MHz, depending on the crystal frequency. The smaller board is a ×4 multiplier that provides an output from 2160 to 2320 MHz, depending on input frequency. Used separately or together, these boards have a wide variety of UHF and microwave applications.

All of the critical microwave circuitry in this LO is readily taken care of by a pair of fiberglass-epoxy (G-10) PC boards. The remaining parts include noncritical chip capacitors for interstage coupling and bypassing, standard ¼- and ½-W bias resistors, inexpensive, plastic-cased monolithic-microwave integrated-circuit (MMIC) amplifiers, a pair of 99-cent diodes, a few hand-wound inductors, disc-ceramic capacitors, and a 90-MHz, 5th-overtone crystal oscillator. PC board manufacturing tolerances, component variations, and construction tolerances have all been allowed for in the design. There are no RF tuning adjustments except for the 90-MHz oscillator tank circuit.

Design Goals

This project began with a list of design goals:

1) No tuning adjustments should be required.
2) All frequency-sensitive elements are printed on G-10 board.
3) Use inexpensive, readily available components.
4) Offer sufficient output to drive a standard-level mixer.
5) Use a single 12-V power supply.
6) All spurious outputs are more than 40 dB down.
7) Have electrical, mechanical and thermal stability consistent with portable CW operation on mountaintops in bad weather.

These goals have been met, with one minor exception: The 90-MHz crystal oscillator tank circuit must be tuned to make the oscillator start reliably. This adjustment can be made by listening for the crystal-oscillator output on an FM-broadcast radio. The electrical, mechanical and thermal stability are impressive. One of these LOs was still operating after an airline baggage-handling event left the aluminum transverter case so badly bent that the top had to be removed with a hammer!

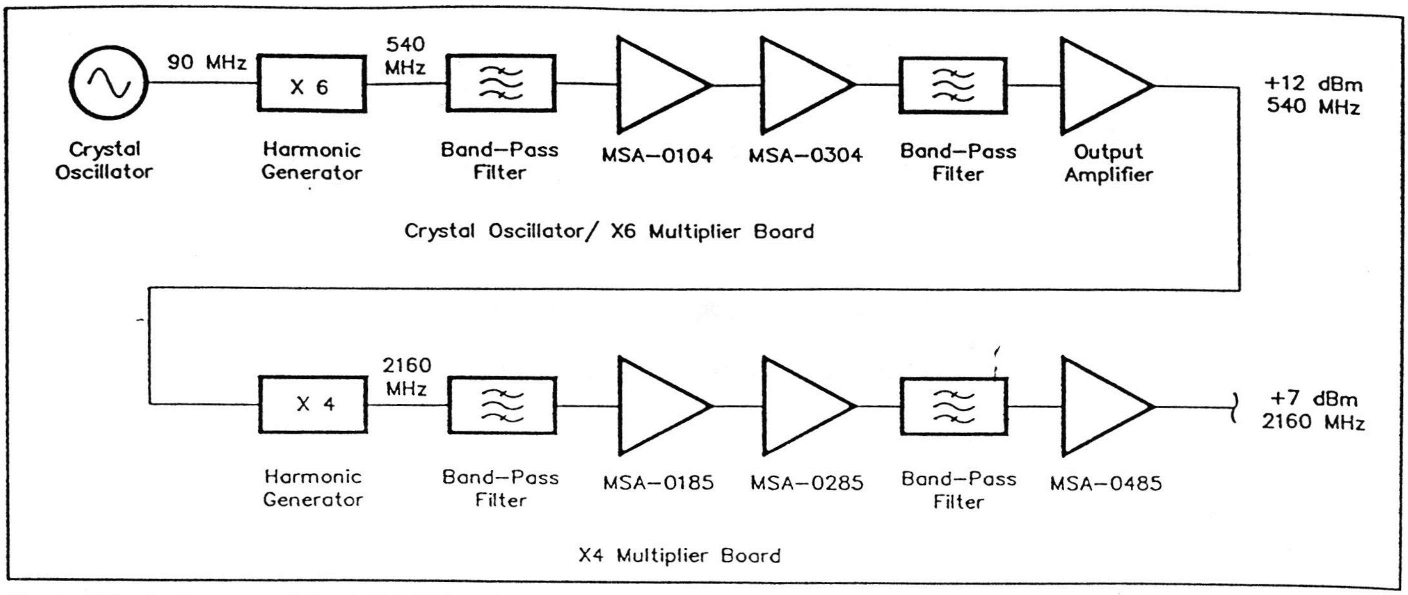


Fig 1—Block diagram of the 2160-MHz LO.

System Description

The complete microwave LO, shown in block-diagram form in Fig 1, consists of two PC boards: a crystal oscillator and times 6 (×6) multiplier board; and a ×4 multiplier board. The crystal oscillator/×6 multiplier board can generate any frequency between 540 and 580 MHz; simply choose the appropriate crystal. The output level depends on the device chosen for the output amplifier. An Avantek MSA-0404 is used for the output amplifier in the version described here. (See this article's Amplifiers section for more details.)

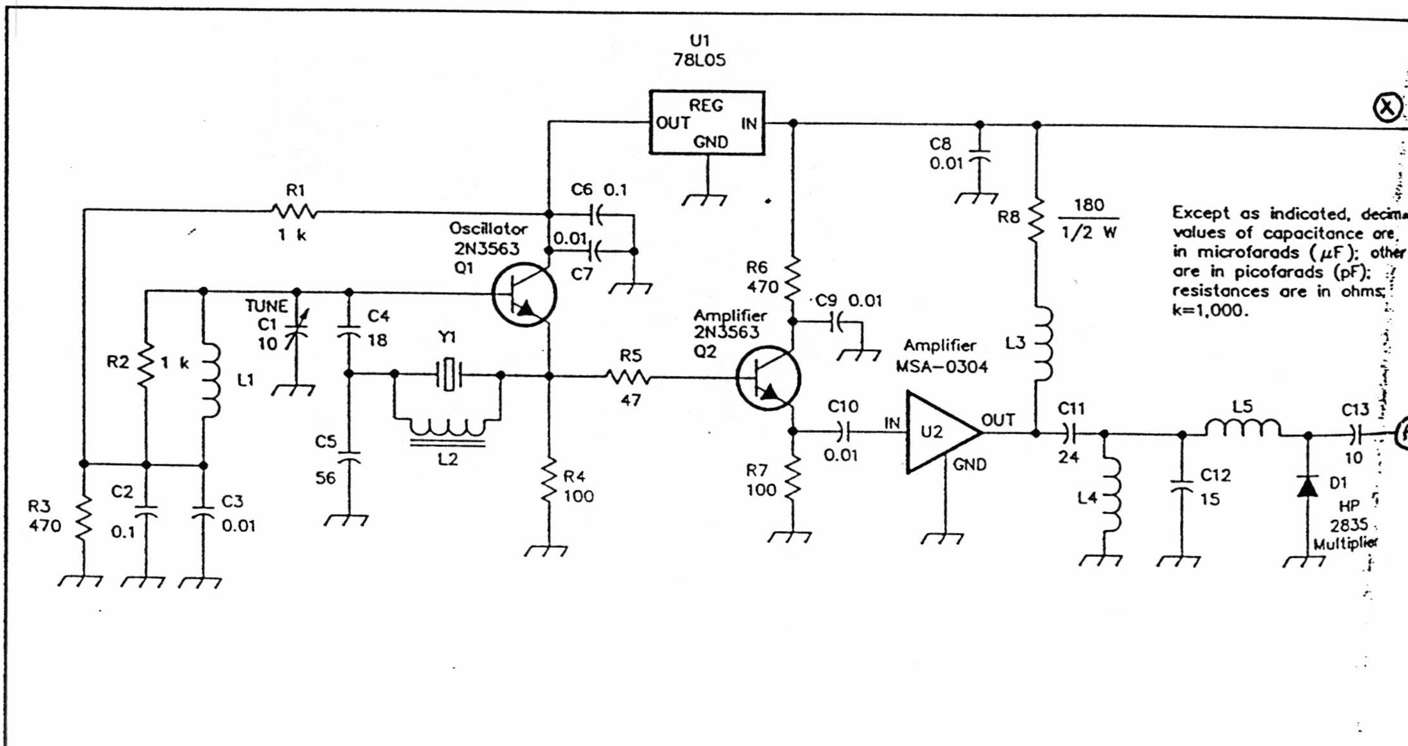


Fig 2—Schematic diagram of the crystal oscillator/×6 multiplier board. Resistors are ¼-W carbon-film types unless otherwise indicated. Capacitors are 50- or 100-V disc-ceramic types unless otherwise noted.

C1—8- to 10-pF trimmer capacitor. Ceramic-piston trimmer preferred; standard ceramic trimmer acceptable.
D1—Schottky diode; Hewlett-Packard 2835, 2800, 2811 or equiv. See text.
J1—SMA female chassis-mount connector preferred. See text.
FL1, FL2—Band-pass filters printed on PC board.
L1, L3, L4, L5, L6—8 turns no. 28 enam wire, 0.1 inch closely wound.
L2—10 turns no. 32 enam wire on T-25-6 toroid core, 0.33 µH miniature RF choke.
L7, L8—3 turns no. 28 enam wire, 0.0625-inch ID, spaced 1 wire diam.

The ×4 multiplier board can be used for any output frequency between 2140 and 2360 MHz. The harmonic-generator components are sufficiently broadly tuned that the board works equally well as a ×3 or ×5 multiplier. Any input level between +7 and +13 dBm is fine, and inputs as low as 0 dBm may be used, at reduced output levels.

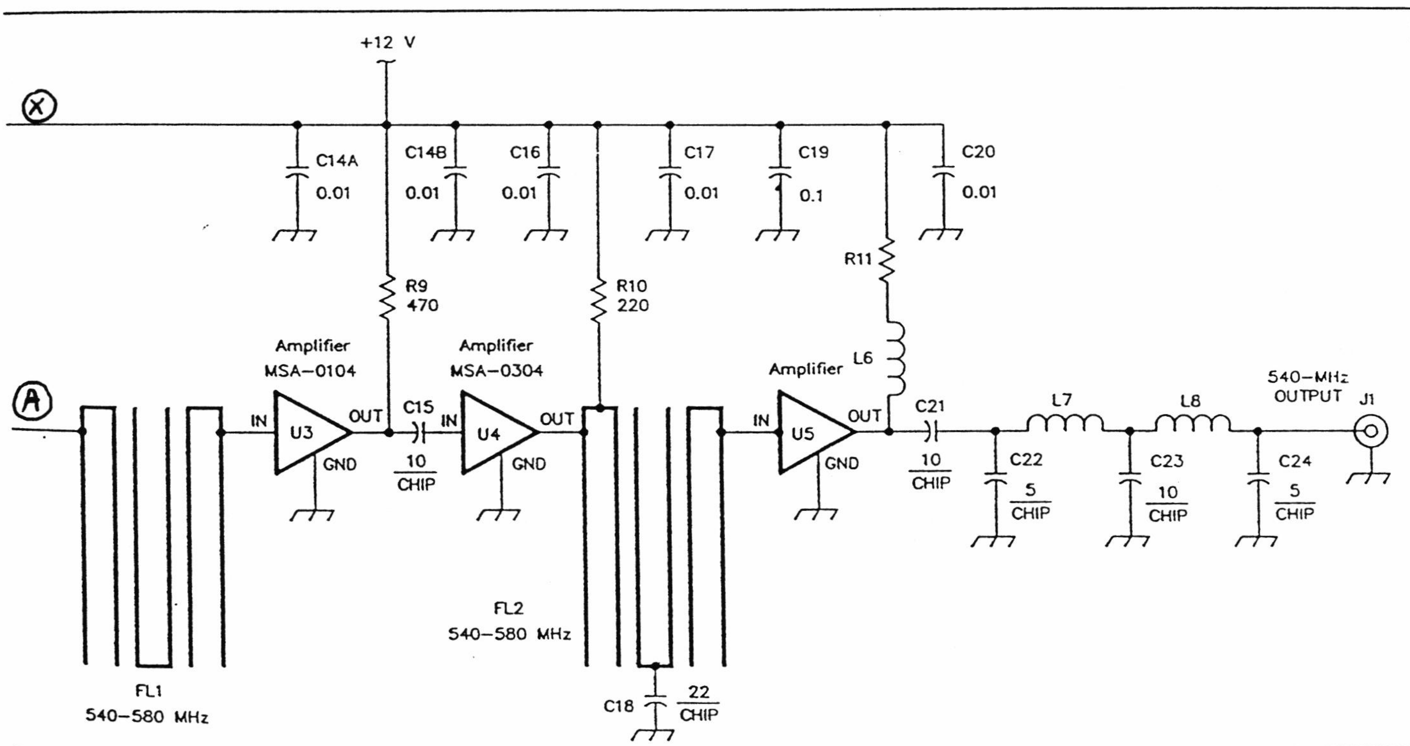
These two boards can be used independently—in fact, they were developed for two separate projects. The 540- to 580-MHz board was developed at the suggestion of Jim Davey, WA8NLC, who wanted a simple 552-MHz driver for his single-board 3456-MHz transverter.[3] The 2140- to 2360-MHz multiplier board was developed as part of a no-tune 2304-MHz transverter that was described in the *Proceedings of Microwave Update '88*.[4]

The local-oscillator system shown in Fig 1 has four functional blocks: the 5th-overtone crystal oscillator; the Schottky-diode harmonic generators; the printed band-pass filters; and the MMIC amplifiers. Each of these blocks is described in the following sections. A schematic of the crystal oscillator/×6 multiplier board is shown in Fig 2, and Fig 3 shows a schematic of the ×4 multiplier board.

Crystal Oscillator

The crystal oscillator generates the signal that is subsequently multiplied into the microwave region. In this design, the 90-MHz crystal-oscillator signal is multiplied by 24 to produce the final output signal (2160 MHz). Any long-term drift or "warblies" on the 90-MHz oscillator will be 24 times worse at the output frequency. Common crystal-oscillator circuits that work well in a 144-MHz, or even 432-MHz, receiving converter may be unacceptable when the output frequency is multiplied into the microwave region.

The Butler emitter-follower circuit shown here was originally suggested to me by Al Ward, WB5LUA, and modified to the present circuit by Jim Davey, WA8NLC. (I don't waste much time arguing with those two—when they express an opinion, they generally turn out to be right.) This oscillator will free-run on the tank-circuit frequency if the crystal and its shunt inductor (L2) are replaced with a 47-Ω resistor. This characteristic is especially useful if you want the tank circuit to operate at another frequency. After initially testing the prototype oscillator with inexpensive 2N5770 transistors, I tried replacing them with some 20-year-old pullout 2N3563s, some MPS3563s, a pair of 2N5179s that I found on the floor under the bench, and some new AT-42085 microwave transistors from Avantek. All of these devices worked in this circuit. I also discovered that the value of R1, which sets the operating points of Q1 and Q2, can be varied to change the power output. A 1-kΩ resistor was fine for all the transistors except the AT-42085s. The output power from the AT-42085s was about



Q1, Q2—2N3563, MPS3563, 2N5179 or equiv. See text.
R11—If U5 is an MSA-0404, use 120-Ω, ½-W resistor. If U5 is an MSA-1104, use a 100-Ω, ½-W resistor. See text.
U1—5-V, 100-mA, 3-terminal regulator.
U2—MSA-0304 MMIC preferred. MSA-0404, MSA-0385, MSA-0485, MAR-3 or MAR-4 also usable. See text.
U3—MSA-0104 MMIC preferred. MSA-0185, MSA-0685, MAR-1 or MAR-6 also usable. See text.
U4—MSA-0304 preferred. MSA-0285, MSA-0385 or MAR-2 also usable. See text.
U5—For +12 dBm out, use MSA-0404. For +16 dBm out, use MSA-1104. See text.
Y1—90-MHz, 5th-overtone, series-resonant crystal.

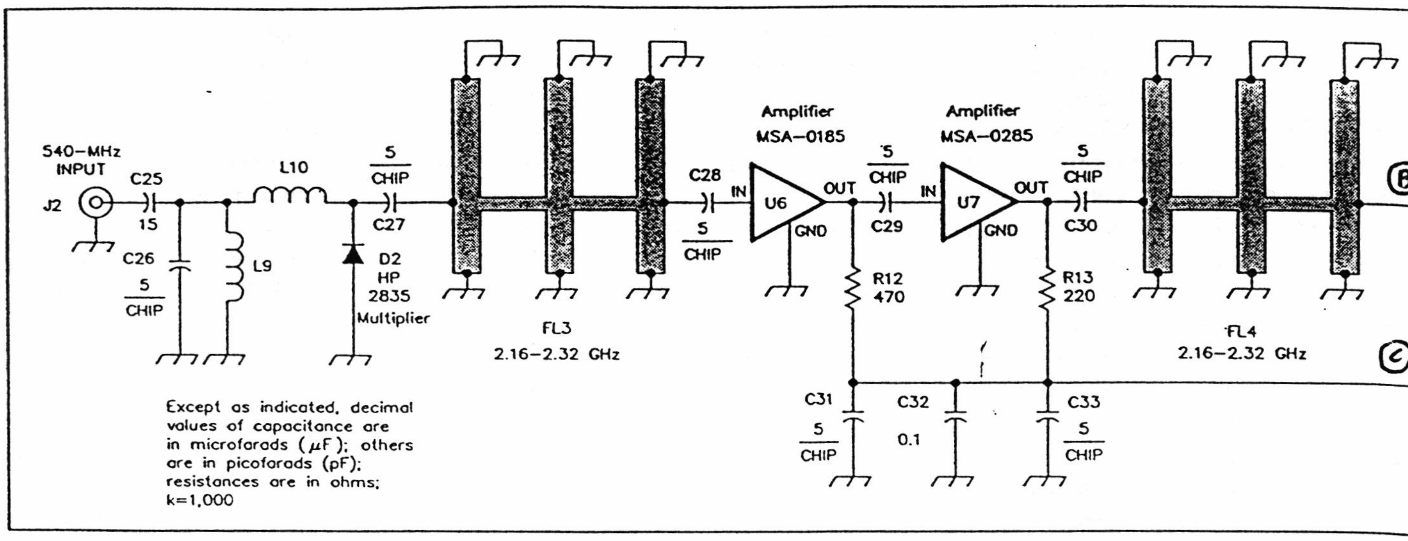


Fig 3—Schematic diagram of the ×4 multiplier board. Resistors are ¼-W carbon-film types unless otherwise indicated. Capacitors are 50- or 100-V disc-ceramic types unless otherwise noted.

D2—Schottky diode; Hewlett-Packard 2835, 2800, 2811 or equiv. See text.
J2, J3—SMA female chassis-mount connector preferred. See text.
FL3, FL4—Band-pass filters printed on PC board.
L9—3 turns no. 28 enam wire, 0.0625-inch ID, spaced 1 wire diam.
L10—Inductor printed on PC board.
U6—MSA-0185 or MAR-1 MMIC preferred. See text.
U7—MSA-0285 or MAR-2 MMIC preferred. See text.
U8—MSA-0485 or MAR-4 MMIC preferred. See text.

+6 dBm—a little too much drive for the MSA-0304 buffer (U2).

The LO shown in the photo on the first page of this article varies slightly from the schematic in Fig 2. The photo shows a Zener-diode regulator in Q1's collector circuit. When this board is used in a setup with a battery supply, the difference in voltage when switching from receive to transmit may be enough to cause an observable frequency shift. This problem is eliminated by using a 3-terminal, 5-V regulator (U1), as shown in Fig 2.

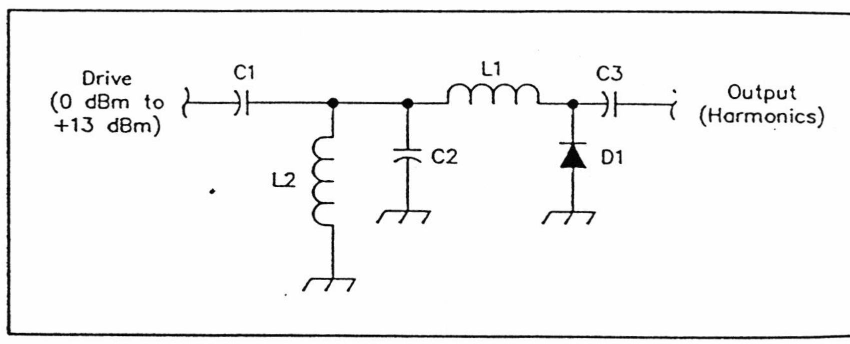


Fig 4—Schematic diagram of the basic harmonic generator used on both boards. See text for discussion.

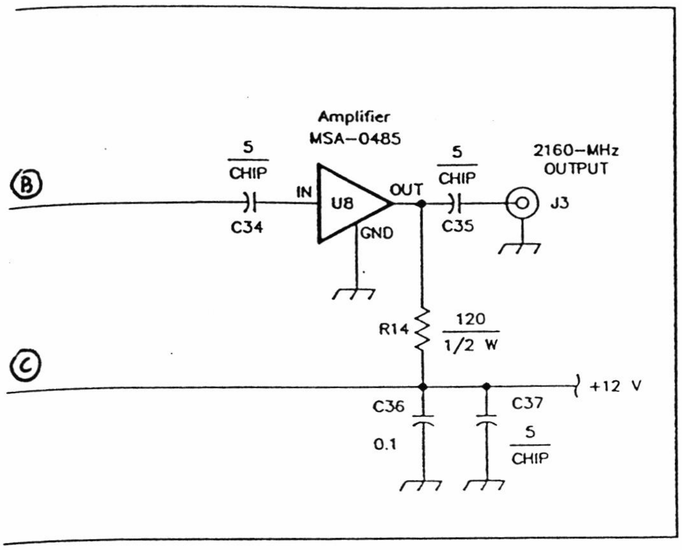
Harmonic Generator

Harmonic generation is easy—or, at least, not generating harmonics is very difficult. Solid-state power amplifiers must be low-pass filtered to get rid of the harmonics generated by the nonlinear characteristics of the transistors. In fact, if you do anything to a sine wave—clip it, drive a class-C amplifier with it, half-wave rectify it—anything that distorts its perfectly sinusoidal shape, the resultant signal will contain harmonics.

Harmonic generation is difficult only if you want to do it with high efficiency. High efficiency in microwave-LO harmonic-generator stages was important in the 1960s and 70s because amplifying a low-level microwave signal to the +7 dBm level required for many diode-ring mixers was expensive. Back then, tuning up efficient multipliers that used expensive step-recovery or varactor diodes required hours in front of a spectrum analyzer tweaking a handful of $5 piston trimmers to within a half turn of oblivion. And you had to do it all again if the drive level, load or temperature changed.

Now that unconditionally stable, broadband MMIC amplifiers are available for less than a dollar, multiplier efficiency is a minor consideration. By taking advantage of readily available modern components, we can build a broadband multiplier—with no RF-tuning adjustments—that is unconditionally stable with variations in temperature, load and drive level.

The harmonic generator shown in Fig 4 is just a half-wave rectifier with a simple low-pass filter (L1) feeding in the fundamental, and a simple high-pass filter (C3) picking off the harmonics. A half-wave rectifier based on an ideal diode generates only odd harmonics. A Schottky diode (D1 of Fig 4) has an offset voltage of a few hundred millivolts, so the conduction angle is less than 180 degrees. Odd and even harmonic levels are approximately equal for drive levels up to about +10 dBm. This basic harmonic generator is used on both boards. For higher drive levels, a bias circuit w



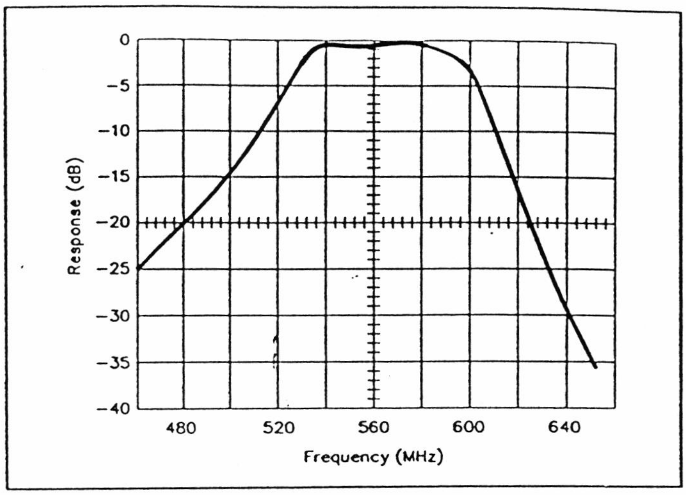


Fig 5—Frequency response of the 540- to 580-MHz hairpin filters printed on the crystal oscillator/×6 multiplier board.

a trimmer potentiometer is recommended.[5]

Other diodes will work in this circuit. Schottky diodes like the HP-2800 and HP-2811 are suitable. A silicon switching diode like the 1N4148 works fine, and even provides more output than the specified Schottky diode, but it has one major disadvantage: It may oscillate! If you don't believe this, compare the circuit of Fig 4 with that of a parametric amplifier circuit in an early VHF manual. Better yet, connect the input of a 1N4148 multiplier to a signal generator and the output to a spectrum analyzer and try it. At some combination of input frequency and drive level, the output noise floor will rise considerably, and the output spectrum will contain many discrete output signals—typically, subharmonics of the drive signal modulating the desired output. This circuit needs to operate reliably from a motorcycle battery on a mountaintop in a rainstorm, so the use of switching diodes is discouraged.

Filters

The filter selects the desired harmonic output from the harmonic generator. In the past, amateur-built frequency multipliers usually were limited to multiplication factors of 2 or 3, because of the difficulty of tuning to the correct harmonic. With fixed-tuned filters having steep skirts and flat tops, it is easy to build multipliers of much higher order. The theory behind hairpin filters (FL1 and FL2 in the 540- to 580-MHz board) and off-center-tapped half-wave filters (FL3 and FL4 in the ×4 multiplier board) is covered in the amateur and professional literature.[6-9] Only the practical aspects are mentioned here.

The primary design goals for these filters were low cost and reproducibility without tuning adjustments. To achieve the first goal, I've specified G-10 board. As a result, the filters are more lossy than equivalent designs on a more expensive substrate, such as Teflon fiberglass. The loss for each of the three-element sections used here is about 3 dB. Because a 10-dB-gain MMIC capable of compensating for this loss costs only about $1, and because better substrates may cost $100 a square foot, G-10 is an attractive trade-off.

To achieve reproducibility without tuning adjustments, the filters are made broadband (with lots of low-Q resonators), rather than narrowband (with only a few high-Q resonators). The resulting passband characteristic, shown in Fig 5, looks more like that of an SSB filter in an HF rig than something out of an LO chain! This filter characteristic is fundamentally different from a single-tuned circuit that must be tuned exactly on frequency—a signal anywhere in the 40-MHz-wide passband (540 to 580 MHz) passes through, but the undesired 5th and 7th harmonics 90 MHz above and below the passband are greatly attenuated by the steep filter skirts.

There are two major advantages to using flat-passband filters in a crystal-controlled LO. One is that the frequency may be moved anywhere in the passband by simply changing the crystal. The other is that allowance for manufacturing tolerances and variations in circuit-board material can be designed in before the circuit is built, rather than having to be tuned out afterward. For example, if the PC-board manufacturer is a little sloppy and the production boards are 1% smaller than the prototype, the passband will be 5 MHz higher. The desired signal will still get through, and the undesired signals will still be attenuated.

Even a change in circuit-board material from G-10 to its fire-retardant variant, FR-4 (which has a slightly different dielectric constant), results only in a well-behaved upward shift in the passband. The FR-4 boards provide a few decibels more output at 576 MHz, and the G-10 boards are a few decibels better at 540 MHz. The passband shape is the same for both G-10 and FR-4 materials.

In an attempt to discover how tolerant the 540- to 580-MHz hairpin filters are, I reduced the length of several hairpin resonators by one millimeter. The output dropped about 2 dB, worst case, and the spurious outputs remained acceptably low. I also sprayed a complete 2160-MHz LO with a thick coat of clear Krylon, and could not detect any change in the module's output. These filters work better than double-tuned circuits; require no tuning; are tolerant of manufacturing, device and construction variations; allow a range of frequencies to be generated from a single board layout; and cost no more than the PC-board material required to make them.

Amplifiers

The MMIC amplifiers serve two important functions: They raise the level of the desired harmonic, and they provide broadband, 50-Ω interstage terminations. Many suitable amplifiers are available, and substitutions require only a little care concerning output level. The MMICs most popular with hams are available from Avantek and Mini-Circuits Labs. The Avantek line has part numbers starting with MSA-; the MSA-0404 is an example. Mini-Circuits parts numbers start with MAR-; the MAR-4 is an example. The devices used here were chosen because they were closest to the top of the pile of parts on my workbench. Drive to D1 should be about +10 dBm—an MSA-0304 or '0404 works well for U2. The desired harmonic coming out of FL1 is at about −17 dBm, so a device with a 50-Ω input and more than 10 dB of gain is useful for U3. The MSA-0104, MSA-0185, MSA-0685 and MAR-6 are good choices.

The second interstage amplifier, U4, must deliver about +10 dBm and provide a 50-Ω load for FL2. MSA-0304, MSA-0285 and MAR-3 devices have all been used in this stage. The final stage, U5, may be omitted if +6 dBm is enough output. If U5 is not used, also omit R11, L6, C20 and C21. A good plan, however, is to use a high-power device that will not saturate, and then add a resistive attenuator to reduce the output to the desired level. An MSA-0404 will provide +12 dBm output, and an MSA-1104 works well for +16 dBm out.

It's not really necessary to operate any of these amplifiers linearly, but the output stage should be kept at or below its 1-dB compression point, or the spurious outputs will rise to an unacceptable level. MMIC selection is left up to the builder. Just about anything will work—and who knows what will be available next year for 49 cents? If you'd rather not think about which parts to use, just use those specified in the schematics. If you use devices different from those specified, be sure to use an appropriate bias resistor, as specified in the manufacturers' data sheets.

Construction

Both boards are etched on 0.062-inch-thick, double-clad, G-10 PC board material. All components are surface mounted on the etched side; the other side is unetched and acts as a ground plane. The components necessary to complete this project are available from several suppliers, and etched PC boards and complete parts kits are available as well.[10]

The first construction step for the crystal oscillator/×6 multiplier board shown in Fig 6 is to drill the component holes and file the edges flat. The filter elements need not be grounded, so the exact size of the board is not critical. The edges of the upper and lower ground foils must be connected all the way around the board, though. A plated-through hole or shorting wire every half inch or so works fine. Copper tape wrapped the whole length of each edge and soldered top and bottom also works well. The method I recommend is to solder brass or tin walls all the way around the circuit board, making sure to solder to both the top and bottom of the circuit board. This results in a nice, rigid box with solderable shield walls suitable for mounting feedthrough capacitors and the output connector.

The next step is to add the bridges across the unused breaks on the crystal oscillator/×6 multiplier circuit board (between C21 and C22), and a piece of copper tape to short the cold end of multiplier diode D1 to the ground plane. The MMIC ground leads are bent at right angles to the device body, passed through holes drilled in the board, and soldered directly to the

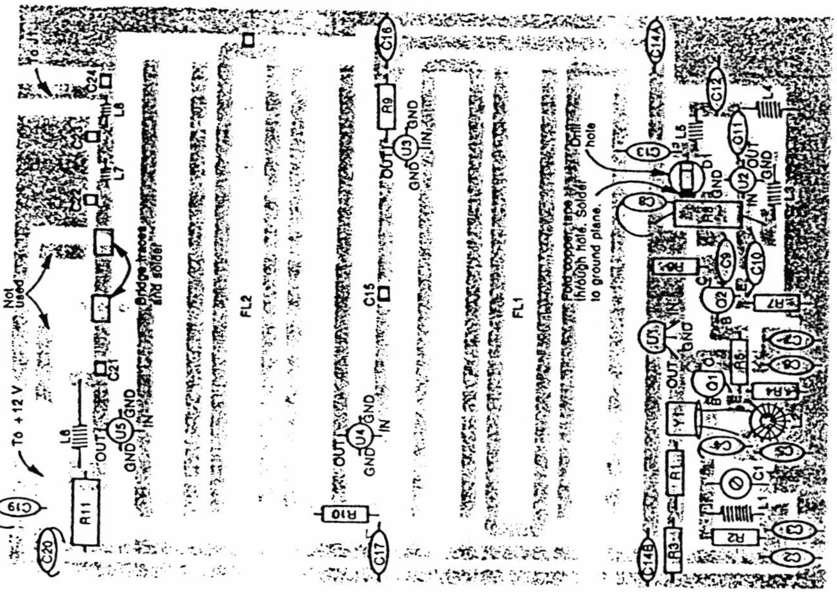


Fig 6—Part-placement diagram for the crystal oscillator/×6 multiplier board (not shown actual size). All components mount on the etched side of the board.

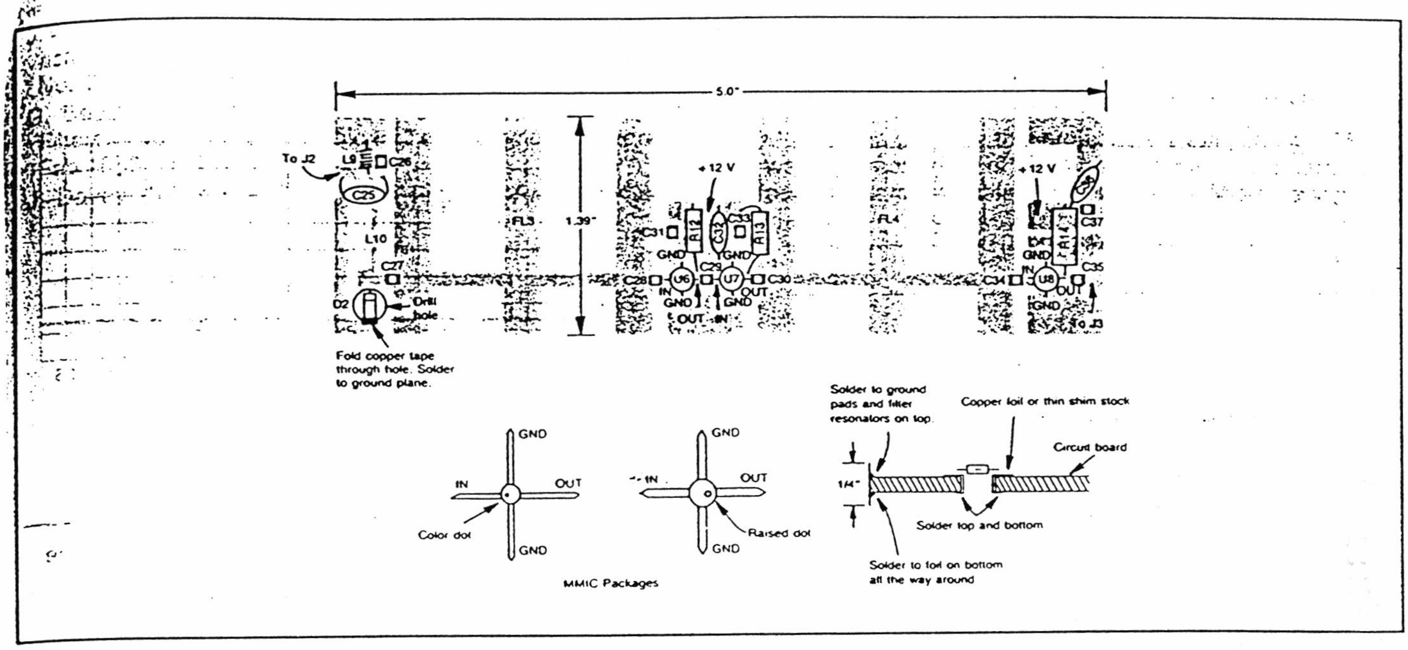


Fig 7—Part-placement diagram for the ×4 multiplier board (not shown actual size). All components mount on the etched side of the board.

ground plane. The 5-V regulator's ground lead also goes through the board and is soldered to the ground plane. I usually put the chip capacitors on first, followed by the inductors, resistors, disc-ceramic capacitors, diode, transistors, 5-V regulator IC and MMICs, in that order. The crystal can be installed as shown on the parts-placement diagram by sticking it to the board with a small piece of double-sided foam tape.

I prefer to use SMA end-launch connectors, even at the output of the crystal oscillator/×6 multiplier board, because they are small and easy to use. (I also use an SMA connector at the output of the ×4 multiplier board because it's an excellent microwave connector. By standardizing on connectors, I don't have to keep switching adapters on my power meter.)

After you've carefully checked all your mounted components against the parts-placement diagram, you can apply 12 V to the board. Tune C1 until you hear the 90-MHz signal in a nearby FM broadcast radio. Then turn the power supply on and off a few times to make sure the oscillator starts reliably. The crystal oscillator/×6 multiplier board is now complete.

The filter topology on the ×4 multiplier board (Fig 7) differs from that of the ungrounded hairpins on the crystal oscillator/×6 multiplier board. The width of the ×4 multiplier board determines the resonant frequency of the shorted half-wave filter elements. The correct length for all the half-wave resonators is obtained by cutting the circuit board precisely to the width shown in the drawing, and then soldering the board's brass wall to the ground plane on the bottom, and to the end of each resonator on the top of the circuit board. Plated-through holes, ground wires and copper tape wrapped around the board edges will not work with this layout.

I obtain the correct board dimensions by scribing a line on the top of the circuit board at exactly the correct place. Then I cut the board slightly oversize. Next, I lay a large, flat file on my workbench and work the circuit board back and forth until the board edge is filed to the scribed line. This results in a nice square edge as shown in Fig 7. Only the width of the ×4 multiplier board is critical. Because FL3 and FL4 form a bandpass filter with a flat passband response, construction errors of up to about 0.032 inch do not significantly affect the output level at 2160 MHz.

After soldering the side and end walls to the ×4 multiplier board, add copper tape to ground the MMICs and multiplier diode as shown in Fig 7. Then add the chip capacitors, disc-ceramic capacitors, inductor, diode, MMICs, bias resistors and SMA connectors. No adjustments need be made to the ×4 multiplier board

Performance

Fig 8 shows the output spectrum of the crystal oscillator/×6 multiplier on G-10 board with a 90-MHz crystal and an MSA-0404 output device. The plot is from dc to 1 GHz. The largest spur, at 450 MHz, is 70 dB below the +12-dBm, 540-MHz output. The 360-MHz and 630-MHz spurs are just barely visible at about 75 dB below the 540-MHz output. The harmonics at 1.08 GHz and 1.62 GHz (not shown) are more than 55 dB below the 540-MHz output, and are not measurable because of the limited dynamic range of this spectrum analyzer. This is a clean LO!

Different crystal frequencies result in different spurious-output levels. Worst-case spurious outputs are about –45 dB for any output between 540 and 580 MHz.

Once the board is built and tested, frequency stability can be enhanced considerably by thermally insulating the crystal oscillator. I usually tape a small piece of sponge packing foam over the oscillator and then package the entire system in a box to keep rain and cold mountain breezes out.

Fig 9 shows the output spectrum of the ×4 multiplier board

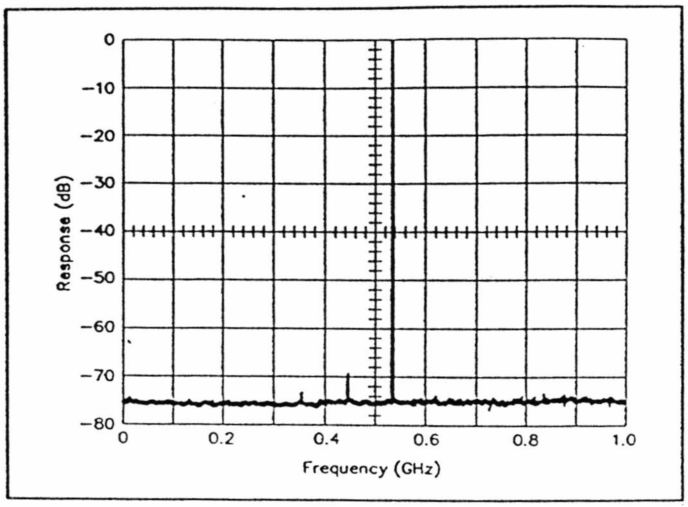


Fig 8—Output spectrum of the crystal oscillator/×6 multiplier board. The desired signal is +12 dBm at 540 MHz.

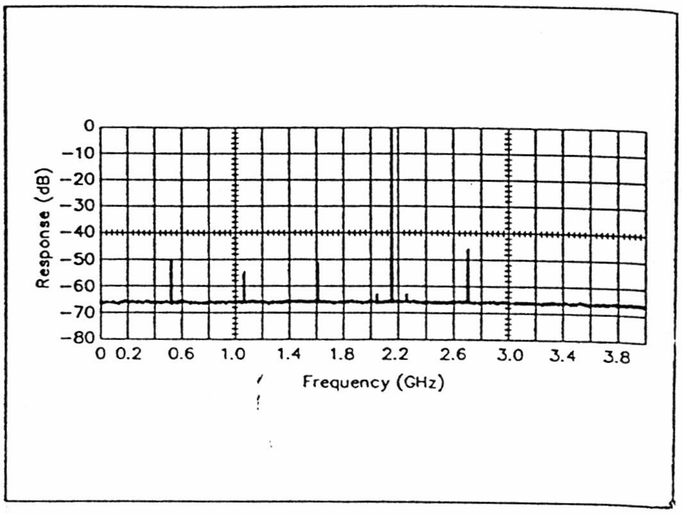


Fig 9—Output spectrum of the ×4 multiplier board. The desired signal is +8 dBm at 2160 MHz.

driven with the signal shown in Fig 8. The plot covers dc to 3.7 GHz. The largest spur, at 2.70 GHz, is 45 dB below the +8-dBm, 2.16-GHz output. The lower frequency spurs, at 0.54, 1.08 and 1.62 GHz, are more than 50 dB down. The second harmonic, at 4.320 GHz (not shown), is only about 25 dB down. However, harmonic spurs are not too important on LO outputs, since the mixer generates harmonics of the LO signal anyway. If the output of the ×4 multiplier board is used to drive an antenna or another multiplier stage, then a filter (as described in reference 5) may be added.

A word about LO drive level is in order here. Many engineers, both amateur and mercenary, agonize because they have a +7-dBm mixer and only +6 dBm of LO drive. It's true that 1-mW (0 dBm) is probably not enough drive for a 5-mW (+7-dBm) mixer, and that 100 mW (+20 dBm) is too much—but there is some latitude. A few decibels either way won't make any measurable difference in most systems. If you really don't want to get on the air, having only +5-dBm drive for your +7-dBm mixer is as good an excuse as any. But the guy down the street with 10 countries worked via moonbounce is probably running +3 dBm into the RF port of an unknown surplus mixer.

Applications

These two boards have been used for a surprising number of applications. I use the 2160-MHz LO described here in a pair of 2304-MHz transverters with 144-MHz IFs, and in a 3456-MHz transverter with a 1296-MHz IF. Simply change the crystal to 96 MHz, and you have a low-power CW transmitter for 2304 MHz. A 94-MHz crystal will provide a 2256-MHz LO for OSCAR Mode S. The ×4 multiplier board may be used with a suitably modified 70-cm FM exciter to generate 2304-MHz FM (460.8 × 5 = 2304 MHz).

The crystal oscillator/×6 multiplier board is even more versatile. With a 96-MHz crystal providing 576-MHz output, it serves as the LO for my single-board 1296-MHz transverter. It may be used with a ×4 multiplier for 2304-MHz CW output, a ×6 multiplier for 3456-MHz output, a ×10 multiplier f 5760-MHz output or a ×18 multiplier for 10.368-GHz outp If an untuned Schottky diode multiplier is used, it will prov useful signal levels for receiver and filter alignment on calling frequencies of all the amateur bands from 2.3 throu 10 GHz. With 92-MHz crystal and an MSA-1104 provid +16 dBm at 552 MHz, it can serve as the LO board for WA8NLC single-board 3456-MHz transverter describe June 1989 *QST*.

Acknowledgments

This work would not have been possible withou forum for amateur microwave technology exchange prov by Don Hilliard, WØPW, in the form of the Microwave U Conferences. The speakers and attendees at Micro Updates '85 through '88 provided much of the basic info tion and all of my motivation for pursuing this wo particular, I thank Al Ward, WB5LUA, for the wealth of mation on MMICs he has made available to the an community, and Jim Davey, WA8NLC, for bringi hairpin filters and encouraging me to keep pushing the s the art.

Notes

[1]A. Ward, "Simple Low-Noise Microwave Preamplifiers May 1989, pp 31-36.

[2]R. Campbell, "2.3 GHz Transverters," *Proceedings c wave Update '88* (Newington: ARRL, 1988), pp 9-

[3]J. Davey, "A No-Tune Transverter for 3456 MHz," C 1989, pp 21-26. *Proceedings of the 21st Conferen Central States VHF Society* (Newington: ARRL, 1 51-57. (Reprinted in this book.)

[4]See note 2.

[5]R. Campbell, "A Clean Microwave Local Oscillator," in *Proceedings of the 1296 and 2304 Conferen* Park, CO, Sep 1985. Reprinted in *Proceedings c Conference of the Central States VHF Society*. N ARRL, 1987), pp 51-57.

[6]J. Wong, "Microstrip Tapped-Line Filter Design," *IEEE Transactions on Microwave Theory and Techniques*, Vol MTT-27, No. 1, Jan 1979, pp 45-51.

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

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

[9]J. Davey, "Microwave Filter Update," *Proceedings of Microwave Update '88*, pp 1-8. See note 2.

[10]Most of the parts for this project are available from Microwave Components of Michigan. Etched PC boards and parts kits are available from Down East Microwave.

PREAMPLIFICATEURS 6 cm

LNA JE1AAH DUBUS 3/89

1. Introduction

In the following a 5.7 GHz LNA and a 5.7 GHz HPA (Power Amplifier) with GaAs-Fets are described and some measurements for performance are given. The LNA design stems from the original design, published by CEL (California Eastern Laboratory) in MICROWAVE JOURNAL, Nov. 1984, titled '6 GHz LNA using NE70083'. The LNA uses the very same transistors. In case of the HPA, it's an own design and uses the cheap MGF1402 transistors. All PCB's are made from 0.8 mm Cu-Clad from Keene.

(Le HPA est dans le § Amplis)

2. 5.7 GHz LNA

The diagram is shown in Fig. 3. No tuning is provided. Without tuning the NF is 2.4 dB and the gain is 21 dB at 5.7 GHz (Figure 4). The etching pattern can be seen in Figure 1. Since there are only quarter wave transformation lines the amplifier is very simple and straightforward.

Each amplifier has a negative bias supply circuit. It uses the ICL7660 voltage inverter IC. It requires only two external capacitors to invert the applied positive voltage. Tantalum capacitors are preferred because of their lower losses. Drain current is set to 10 mA with the bias pots. That's easily verified by measuring the resulting drain voltage as 3 V.

For interstage match and output match additional tuning stubs have been provided. The whole PCB is housed by machined aluminia box, to assure mechanical stability (Figure 2).

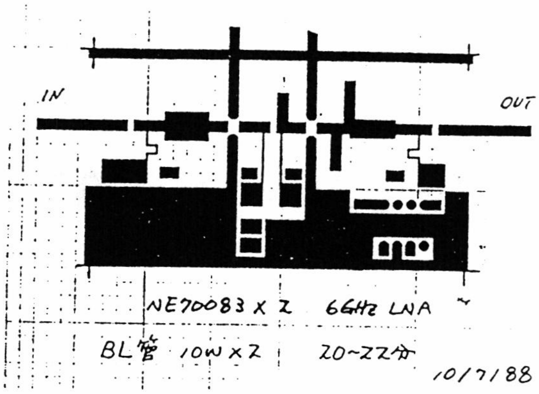


Figure 1/Bild 1: Etching Pattern/Layout of LNA

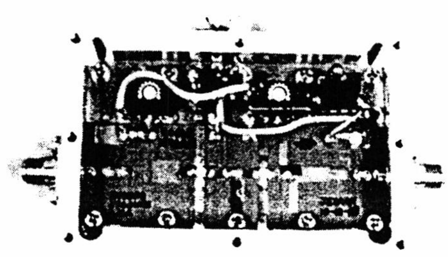
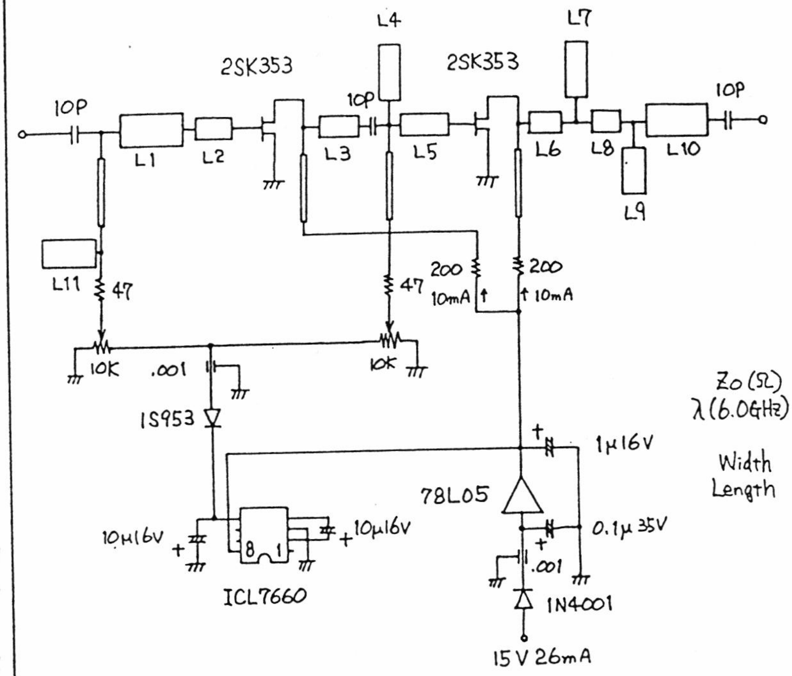


Figure 2/Bild 2: 5.7 GHz LNA

FIGURE 3 (BILD 3): Diagram/Schaltung LNA

f 5760 MHz

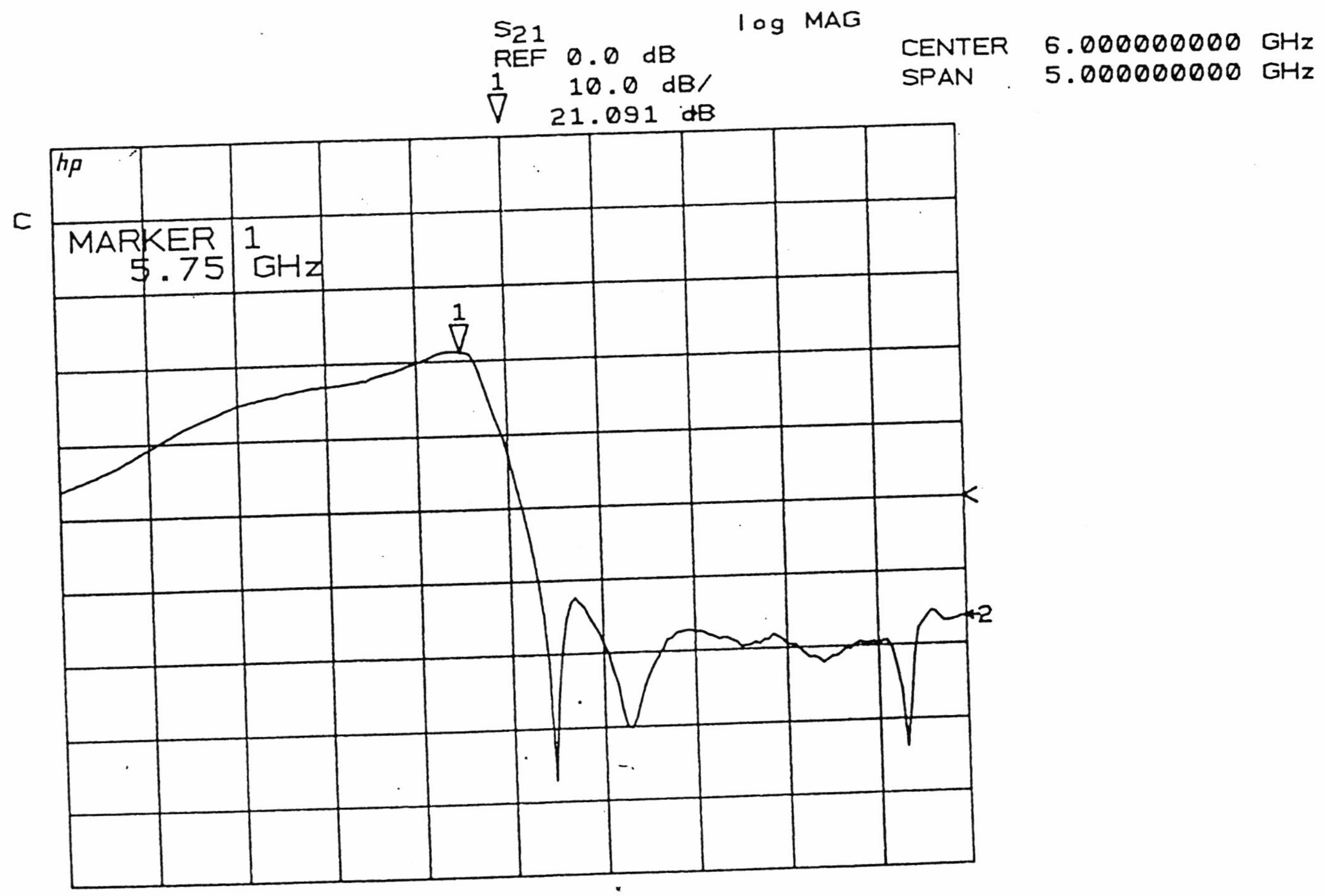


IN −18.8dBm OUT 2.2dBm Gain 21.0dB

(−80dBm 入力位では、25~6dB 利得ありそう (at -80dBm input gain is 25~26dB)) • (ケースのふたをはずすと 23dB~26dB位に上がる) (without cover plate gain increase to 23~26dB)

	L1	L2	L3	L4	L5	L6	L7	L8	L9	L10	L11	BIAS
Zo (Ω)	25	50	→							35	30	130
λ(6.0GHz)	.25	.10	.172	.15	.157	.107	.197	.06	.224	.25	.25	.25
Width	5.5	2.2	→							3.6	4.5	0.3
Length	8.4	3.5	6.0	5.2	5.4	3.7	6.8	2.1	7.8	8.5	8.5	9.1

FIGURE 4 (BILD 4): Gain @ F/Frequenzgang LNA



GaAs MMIC Amplifier

Until recently, affordable MMIC devices only worked well at frequencies below 4 GHz. Some devices still had a little gain at 5760 MHz, so it was possible to make an amplifier using multiple low-gain stages.[1 18] A new GaAs MMIC, the MGA-86576 (also available from Down-East Microwave) offers excellent performance to 8 GHz at a very reasonable price. I built an amplifier using one on a scrap of Teflon PC board with a 50-Ω transmission line printed on it. It is important to keep the ground leads very short, so I cut a tight-fitting hole through the board and mounted the MMIC with the ground leads on the ground side of the board and the input and output leads bent up through the hole in the board to the input and output transmission lines. A schematic diagram of this simple amplifier is shown in Fig 7.

The amplifier has about 15 dB of gain and is capable of a few milliwatts of output power, so it would be suitable for a low-power rover station or as a driver for a power amplifier. It also works well as a receiving preamplifier, with a noise figure around 2 dB. The amplifier is quite broadband, with similar noise figure down to 1296 MHz (gain falls off below 1 GHz); measured performance is shown in Table 1.

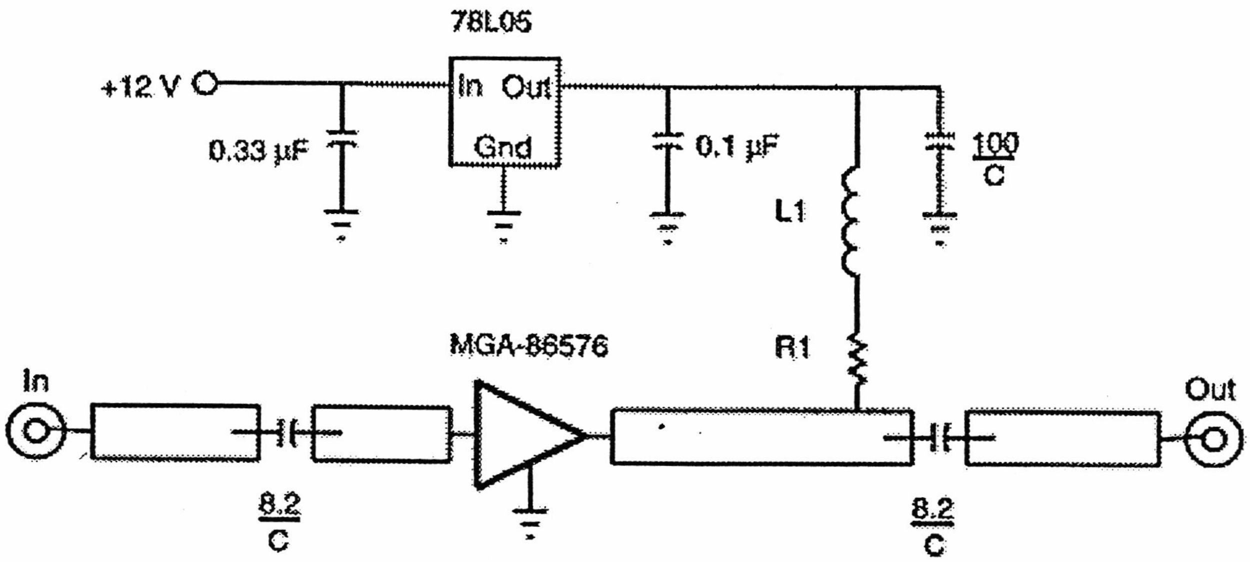


Fig 7

Table 1—GaAs MMIC Amplifier Performance

Freq MHz	*Gain*	*Return loss input*	*Return loss output*	*|S12|*
1296	19.2 dB	-3.9 dB	-9.7 dB	-48 dB
2304	20.6	-5.8	-9.7	-37
3456	21.0	-6.4	-10.8	-32
5760	15.4	-7.5	-14.5	-28
10368	7.1	-7.4	-13.9	-20

Noise figure was about 5 dB at 10368 with a 3-dB second stage.

Simple Low-Noise Microwave Preamplifier

By Al Ward, WB5LUA

(From May 1989 *QST*)

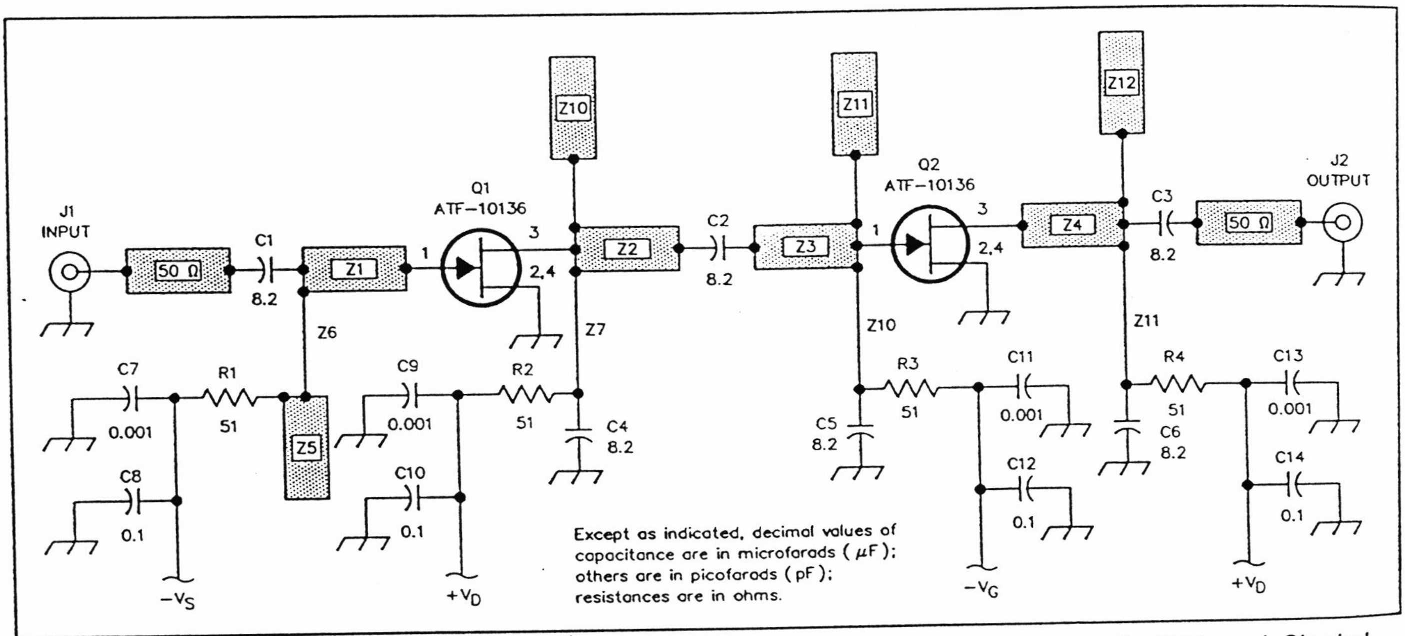


Fig 3—Schematic of the 5.7-GHz preamplifier. Z1 through Z12 are microstriplines etched on the PC board. Shaded rectangles marked "50-Ω" are 50-Ω transmission lines etched on the PC board. All resistors and capacitors are chip types. C1-C6 are 0.05-in. square. C8, C10, C12 and C14 enhance "low-frequency" bypassing. J1 and J2 are SMA female connectors; see text.

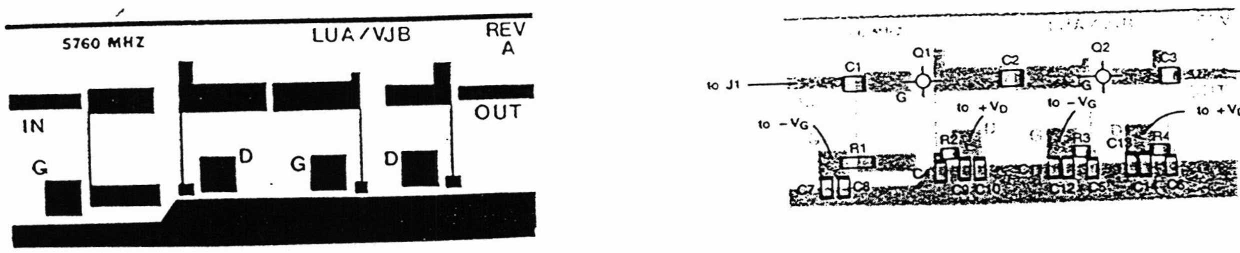


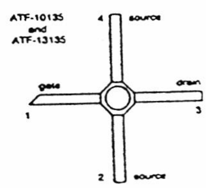
Fig 7—Circuit-etching patterns and parts-placement guides for the preamplifiers. The etching patterns are shown at full size. PC board material is double-sided, 0.031-in-thick Rogers Duroid 5880 or Taconic TLY-5 (dielectric constant, 2.2). Black areas represent unetched copper foil. The back side of the board is left unetched to act as a ground plane. The parts-placement guides are not shown at their actual size. All components mount on the etched side of the board.

Table 2

Actual Preamplifier Performance Versus Computer Simulation and Projected Worst-Case Performance

Device	Freq (GHz)	Bias per device	Gain (dB) Type	Gain (dB) Worst Case	Gain (dB) Simul	Noise Figure (dB) Typ	Noise Figure (dB) Worst Case	Noise Figure (dB) Simul
ATF-10136*	2.3	2 V @ 20 mA	13.5	12.0	13.9	0.5-0.6	0.8	0.60
ATF-10136**	2.3	2 V @ 20 mA	13.0	12.0	13.0	0.65	0.9	0.70
ATF-10136	3.4	2 V @ 20 mA	23.0	22.0	24.1	0.8-0.9	1.0	0.58
ATF-10136	5.7	2.5 V @ 15 mA	18.0	17.0	20.6	0.9-1.0	1.2	0.85
ATF-13136*	10.4	3 V @ 20 mA	8.5	7.5	10.6	1.25-1.5	1.7	1.25
ATF-13136	10.4	3 V @ 20 mA	18.0	15.0	21.9	1.5-1.7	2.0	1.35

*Single-stage amplifier
**Self-biased, single-stage amplifier



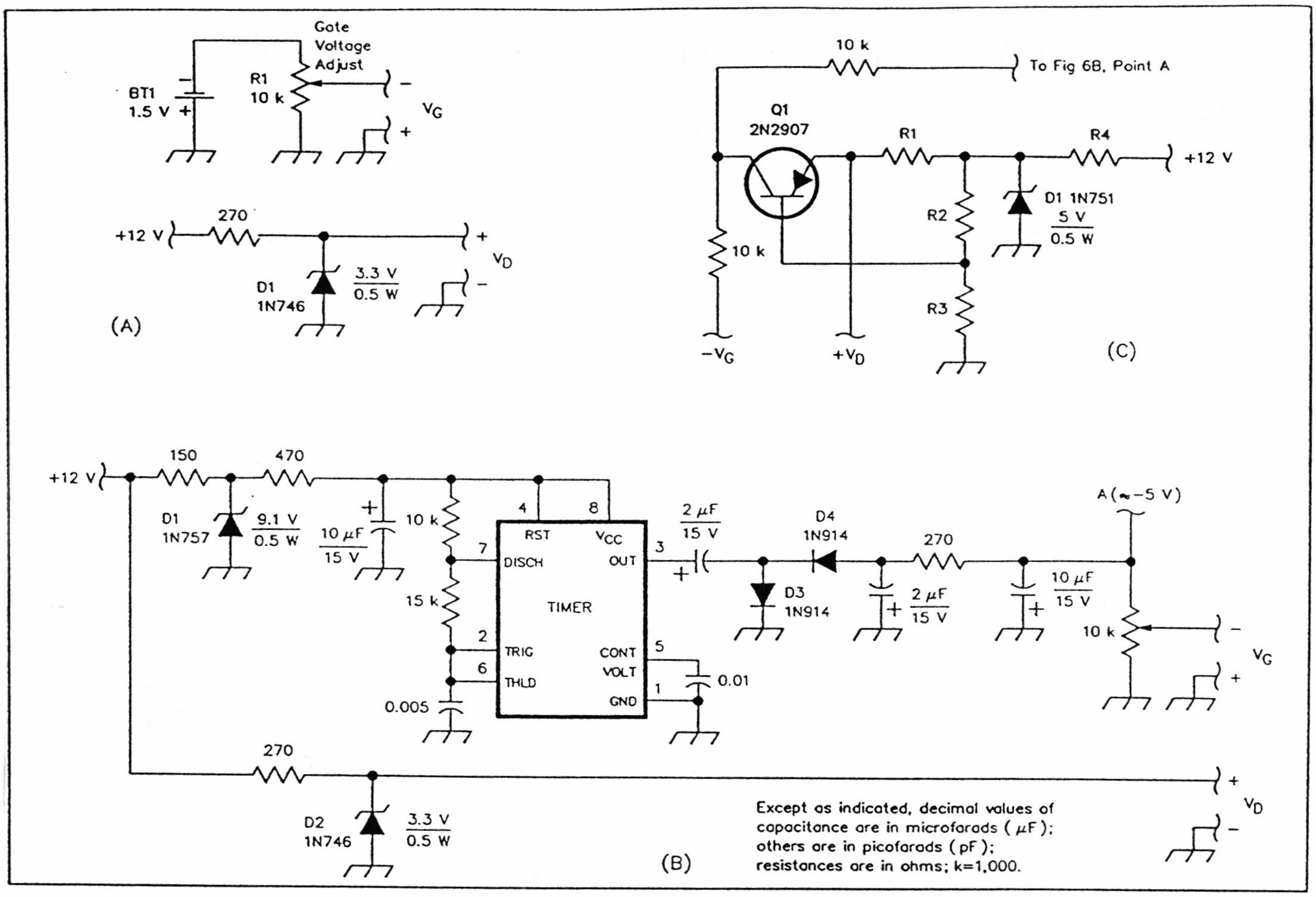


Fig 6—Bias circuits for the preamplifiers. See text for discussion. The passive circuit at A uses a 1.5-V cell for the gate supply and a Zener diode to stabilize the drain supply. The circuit at B, another passive arrangement, uses a 555 timer IC to generate negative gate bias, and a Zener diode to stabilize the drain supply. C shows an active bias circuit. The values of R1, R2 and R3 can be varied for different FET operating conditions; see text and Table 1. The value of R4 should be chosen so that 10-15 mA of Zener current flows when the FET (or FETs, for a two-stage design) is powered at rated bias.

Three basic bias circuits are shown in Fig 6. Two are passive. The third—and most desirable—is an active bias network that uses a PNP transistor to set the GaAsFET drain voltage and current.

The simplest bias network, shown in Fig 6A, uses a 3.3-V Zener diode to set the drain voltage. A 1.5-V AA cell is used for the bias supply. Bias, applied through R1, sets the gate voltage, which in turn determines the drain current. Generally, the AA cell is connected so that there is always a negative voltage applied to the gate. The preamplifier is then turned on by connecting V_D to a positive voltage source. Because there is a 51-Ω resistor in series with the drain, there is some interaction between drain voltage and drain current: Greater drain current produces a lower drain voltage. The gate voltage required to properly bias the device varies from unit to unit because of slight variations in pinch-off voltage. (Pinch-off voltage is the gate voltage required to turn the FET off.) A disadvantage of this simple bias circuit is that it lacks compensation for bias changes over temperature variations. Although not optimum, this technique has been used at WB5LUA and WA5VJB with good results. A high-grade, long-life alkaline AA-size cell should last several months before its voltage drops low enough to cause the FET to draw excessive drain current.

An adaptation of the simple passive bias configuration is shown in Fig 6B. Drain voltage is again provided by a 3.3-V Zener diode, but this circuit replaces the AA cell with a positive-to-negative voltage inverter. Several manufacturers make suitable inverter ICs. A less-expensive approach is to use a common 555 timer in the simple inverter circuit shown. With simple battery bias, the negative supply is continuously applied to the FET gate. With the inverter of Fig 6B, gate and drain supplies are simultaneously applied to the FET. This approach has been used by manufacturers of satellite TV receiving equipment for years. Although there can be problems if the drain voltage is applied before the negative gate voltage, the 51-Ω resistor in series with the drain safely limits the maximum drain current—even if the application of gate voltage to the FET is delayed.

Table 1

Active Bias Circuit Values for Various Drain Currents

V_{DD} (V)	V_{DS} (V)	I_D (mA)	R1 (Ω)	R2 (kΩ)	R3 (kΩ)
3.5	2.5	20	75	2.2	2.8
3.25	2.5	15	117	2.2	2.3
4.0	3.0	20	50	2.2	4.3

An improvement on both passive circuits is the active bias circuit shown in Fig 6C. The negative gate supply uses the inverter of Fig 6B, but the drain supply employs an inexpensive PNP transistor in a circuit that effectively sets both the drain voltage and drain current regardless of device variations. It also offers a more constant bias over temperature than the passive designs. Drain voltage is set by R2 and R3, a voltage divider at the base of Q1. The voltage is then raised by the emitter/base junction voltage of Q1. Drain current is set by R1. The gate voltage required to sustain the drain voltage and current is set automatically by the voltage divider set up by the emitter/collector junction of Q1 and the negative voltage source. About –1 V is supplied to the FET gate.

Table 1 gives resistor values for various bias conditions. I suggest building a separate bias network for each FET stage to properly set each device's bias point. I have, however, used a single active bias supply to power a two-stage amplifier with good success. If the devices are fairly well dc matched (drain current vs gate voltage), this technique will be okay. It will not, however, keep the device drain currents equal if they are not dc matched.

The active bias arrangement can also be used with a battery instead of the voltage inverter. Since the active bias network automatically adjusts gate voltage for a required bias condition, the circuit will adjust the gate voltage as it drops with battery age. The gate requires about –1 V, so the battery can age significantly before the FET bias condition is altered significantly. If this technique is used, it is best to start out with a 5- to 6-volt battery source. The active bias network will compensate for a battery voltage deteriorating to 1-1.5 V. Active bias networks are discussed in greater detail in Hewlett-Packard application note ANA002.

Construction

Construction of all amplifiers is similar. Part-placement guides and etching patterns are shown in Fig 7. All amplifiers are etched on 0.031-in.-thick Duroid 5880 or Taconic TLY-5 PC-board material with a dielectric constant of 2.2. The etched PC board can be installed in a housing such as a die-cast aluminum box. Another method, one that I prefer, is to solder thin (0.02-in.-thick) brass side walls to the PC board to form a shielded enclosure. The brass walls also connect the top and bottom ground planes, which is essential for low-loss "low-frequency" bypassing. Power connections for V_G and V_D can be made via 0.001-µF feedthrough capacitors soldered to the brass walls. See Figs 8 and 9.

SMA-type end-launch connectors are used for J1 and J2 to provide a transition from coaxial cable to the microstripline. Two- or four-hole gold-plated connectors are easily soldered to the PC board or brass side walls, depending on your assembly technique. End-launch connectors are preferred to the right-angle type because of the impedance discontinuity associated with the right-angle transition. Additional amplifier tuning may be required if right-angle connectors are used.

The type and size of the chip capacitors used in these amplifiers becomes increasingly important as frequency increases. For the blocking capacitors, I strongly recommend using good-quality RF-type ceramic chip capacitors, such as those made by ATC. The values specified are common and should not be hard to find. The physical size of the capacitors is especially critical at 10 GHz, where the 0.05-in.-square type *must* be used. Anything larger produces a sizable mismatch on the microstripline.

The value of the "low-frequency" bypass capacitors is less critical. Anything in the 820- to 1500-pF range will work fine. The value of the "high-frequency" bypass capacitors is somewhat more critical, though. Stay within 10% of the values indicated. Again, use good-quality capacitors for "high-frequency" bypassing.

To obtain a low noise figure, the preamp's FET source leads must be properly grounded. In the case of the 3456-MHz and higher-frequency preamplifiers, bend the FET source leads down right at the case and insert them through slots in the PC board. See Fig 10A. This technique works fine for the long leaded devices. For the short leaded devices, I would suggest using 0.050 inch to 0.070 inch wide copper wire straps to bring

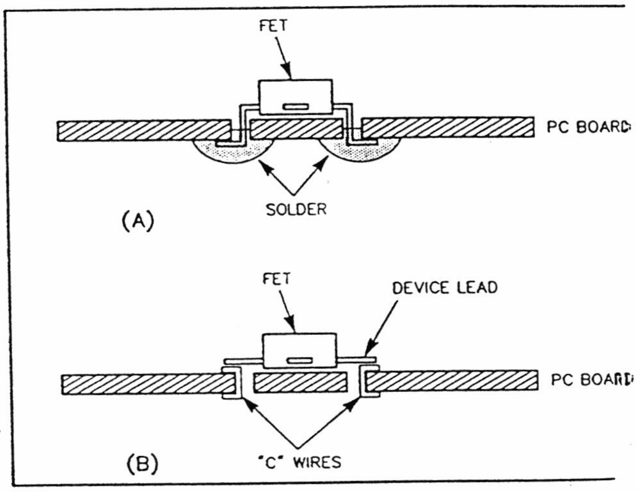


Fig 10—The ATF-10135 and -13135 FET source leads are bent, inserted through slots cut in the PC board and soldered to the ground plane. The ATF-10136 and -13136 leads are not long enough to pass through the board, so you must add "C wires." See Fig 10B.

the bottom ground plane to the top of the printed circuit where the device can be soldered directly to the "C wire". The "C wire" strap should be positioned close enough edge of the package such that the device source lead lay minimal, i.e. less than 0.010 inches of source lead 3456 MHz through 10368 MHz preamplifiers.

The slots can be made with a sharp hobby knife or something similar. Be sure to clean up the area where the leads will pass through the slots by removing any extra dielectric material or copper. The source leads should be passed through the board, again bent at right angles and laid neatly along the bottom foil. Solder them to the bottom ground plane and try to force the solder to cover the length of the slot if possible.

PHEMT LNA for 6cm

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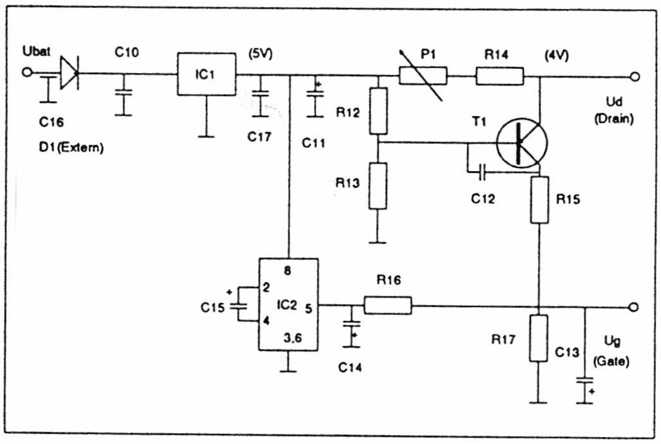
Abstract: Preamplifiers equipped with PHEMTs provide top notch performance in noise figure and gain as well as unconditional stability for the 13 cm, 9 cm and 6 cm amateurbands. Typical noise figures range from 0.35 dB on 13 cm, 0.45 dB on 9 cm to 0.65 dB on 6 cm. The 13 cm and 9 cm models utilize the C-band PHEMT NEC NE42484 and provide a facility for an optional second stage on board. The second stage with the new HP GaAs-MMIC MGA86576 can boost the gain to about 40 dB in one enclosure. The 6 cm version is a one stage preamp which utilizes a quarterwave coupler in the input circuit and a X-band PHEMT NEC NE32684. The features render them useful for EME, satellite and tropo work on these bands.

On Board Bias Circuit



Measurements LNAH-5.7-N326	
Device:	NE32684
Noise Figure:	0.65 dB typ. @ 5760 MHz
Gain:	13 dB typ. @ 5760 MHz
Input RL:	6 dB
Output RL:	20 dB
Bandwidth:	NF < 0.8 db from 5.3..6GHz
Stability K:	> 1 from 0.2..20 GHz

Measurement Results

Fig. 13 show the measurement results for gain and noise figure for the 1-stage. A typical noise figure of 0.65 dB at a gain of 13 dB can be measured on 5.76 GHz. The preamp is rather broadbanded. A low noise figure is provided from 5.4 to 6 GHz.

Stability

Stability is excellent. This has been achieved by a carefully controlled mixture of inductive source feedback, resistive loading in the drain and appropriate DC-feedstructures for drain and gate. A broadband sweep from 0.2 to 20 GHz shows a stability factor K of not less than 1 and the B1 measure is always greater than zero. These two properties indicate unconditional stability.

Description

This LNA utilizes a specially designed quarter-wave stripline coupler in the input. This innovative technique provides about 0.1 dB less loss than a high-Q ATC-chip, hence 0.1 dB less noise figure when used for LNAs and last but not least it's very inexpensive. Using the NEC32684 PHEMT a noise figure of 0.65 dB can be measured on 5.760 GHz at a gain of 13 dB.

Fig. 12 shows the circuit diagram and Fig. 14 the PCB with the dimensions 34x47 mm. It's printed on RT-Duroid 5870 with 0.508 mm thickness. The quarterwave coupler at the input has a length of 9.3 mm and a slit width of 100 µm. The substrate thickness makes the coupler low loss and provides low inductance plated through holes necessary for stability.

A separate bias circuit as described in [2] provides an independent adjustment facility for drain current and voltage as well as regulation.

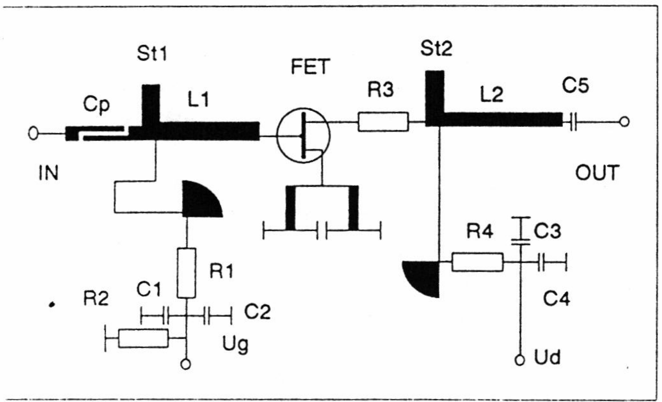


Figure 12: Circuit for LNAH-5.7-N326

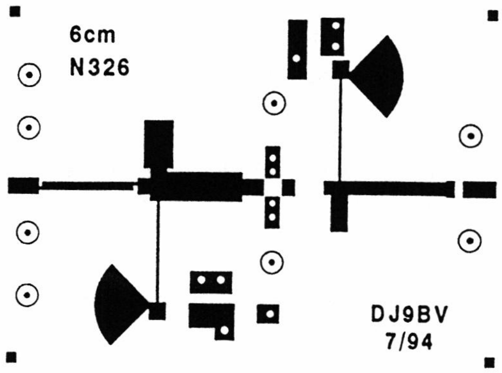
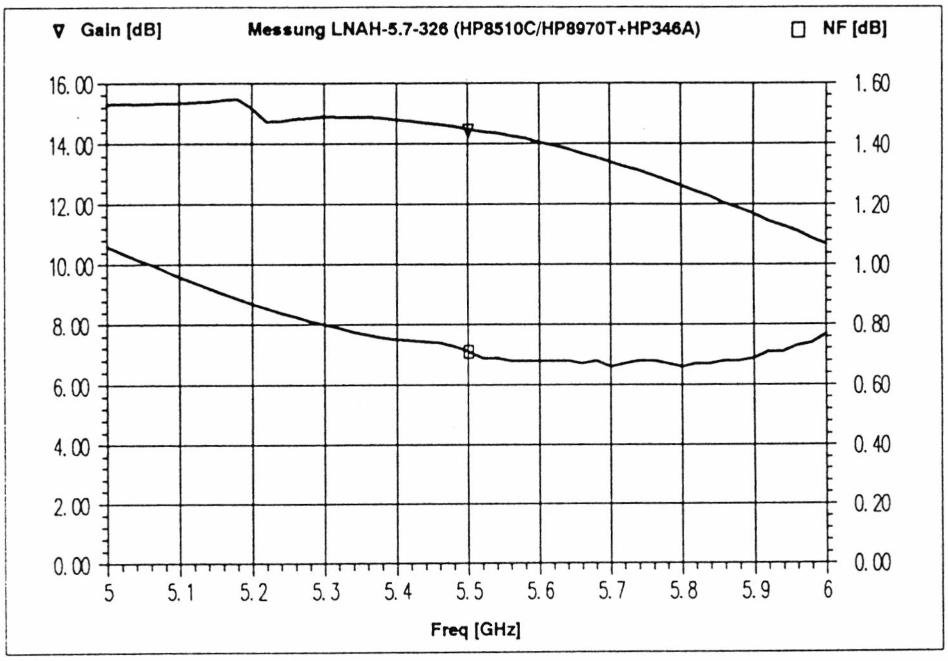


Figure 14: PCB LNAH-5.7-N326

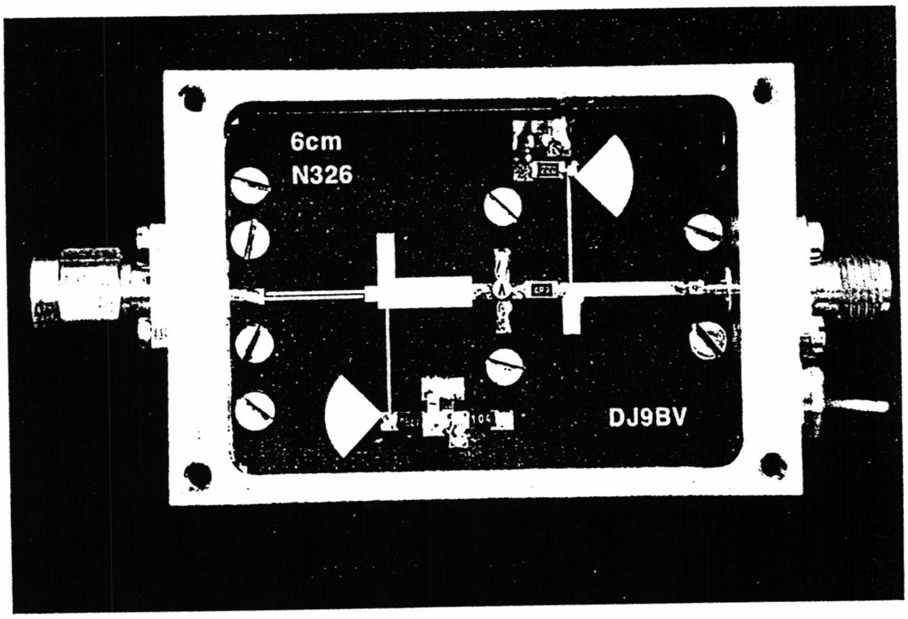
Figure 13: Measured Noise Figure and Gain of LNAH-5.7-N326



References

[1] R. Bertelsmeier, DJ9BV, "HEMT LNAs for 23cm", DUBUS 4/1993, pp. 47-60

[2] R. Bertelsmeier, DJ9BV, "HEMT LNA for 13cm", DUBUS 3/1992, pp. 35-41



6cm Preamp LNAH-5.7-326

Parts/Teile

Parts, Kits, PCBs and detailed descriptions are available from Rainer Jäger, DC3XY, Breslauer Str.4, D-25479 Ellerau. Tel.: (++49)4106-73430.

Ready made and calibrated preamps are available from Frank Schreyer, DD1XF, Maimoorweg 32, D-22179 Hamburg, Tel. (++49)406428253.

Table 1: Parts List of LNAH-5.7-N324

Teile-Nummer/Part-No.	Art/Sort	Wert/Value	Hersteller/ Manufacturer	Herst.-Bez./Manuf.-No.
C1,3	SMD-C	100pF	Sie	
C2,4	SMD-C	1000pF	Sie	0805
C5	Chip-C 50mil	1 pf (500 CHA 1R0 JG)	Tekelec	CHA
R1	SMD-R	47	Sie	1206
R2	SMD-R	100k	Sie	1206
R3	SMD-R	4R7	Sie	1206
R4	SMD-R	22	Sie	1206
FET	GaAs-FET	NE32684	NEC	
Bu1,2	Coaxial	SMA	Radiall/Suhner	
PCB	Teflon PCB, Duroid 5870	mm, 0.508 mm, Er=2.3, - LNAH-5.7-N326	DC3XY	
G	Box, Aluminium	35x74x30 mm		

Acknowledgements

I have to thank Klaus Eichel, DL6SES from TSS for lending the MICROWAVE HARMONICA software, which proved to be an invaluable tool for designing the preamps; Frank Schreyer, DD1XF and Uwe Nitschke, DF9LN for building some of the prototypes; and last but not least Dieter Briggmann, DC6GC for measuring the S- and noise parameters of the preamps. Without their help and their valuable ideas and contributions this work would not have been possible.

RF

by Zack Lau, KH6CP/1 QEX 09/94

A Low-Noise PHEMT Amplifier for 5760 MHz

With the growing interest in the 6-cm band, for both terrestrial and satellite work, I've decided that this band could use more circuits to copy. Unlike at VHF, where the sky is pretty noisy, cold-sky temperatures can be quite low on this band, sometimes getting close to the 3-Kelvin limit associated with the Big Bang. Because of the low sky temperature, a 0.7-dB system noise figure can result in as much as a 7-dB improvement over a 3-dB system. In contrast, a 0.7-dB system noise figure at 2 meters is likely to have only a 2-dB improvement over a 3-dB system. As amateurs in heavily populated areas have found, even this much of an improvement may not be available at 2 meters because of the high noise level from electrical interference on VHF. The design presented here has a 0.65-dB noise figure with 22 dB of gain. This noise figure is approximately 0.3-dB less than that realized from good designs that use the ATF 10135 GaAs FET.

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System Design Considerations

While the goal is to minimize the noise figure of the preamp as much as

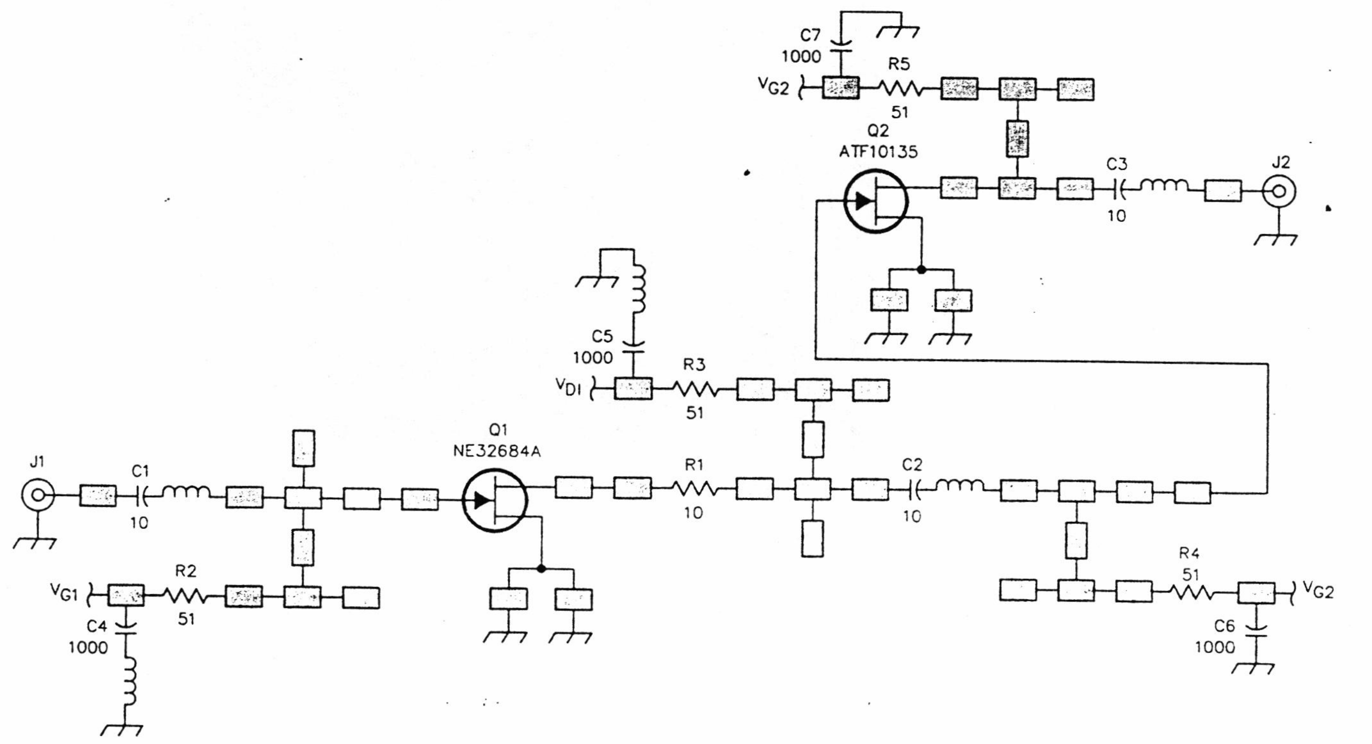


Fig 1—Schematic diagram of the 5760-MHz, two-stage, low-noise preamplifier.

C1-C3—10-pF ATC 100A chip capacitors.
C4-C7—1000-pF chip capacitors.
J1, J2—SMA panel jacks.
Q1—NEC 32684A PHEMT, available from California Eastern Laboratories. Long-lead device preferred.
Q2—ATF 10135 GaAs FET, available from Hewlett Packard. Short-lead devices are usable (see text for details).
R1—10-Ω chip resistor.
R2-R5—51-Ω chip resistors.

Microstrip transmission lines are indicated in the drawing by rectangles. These are modeled in the *Microwave Harmonica* simulation of the circuit, as are the stray (unlabeled) inductances shown.

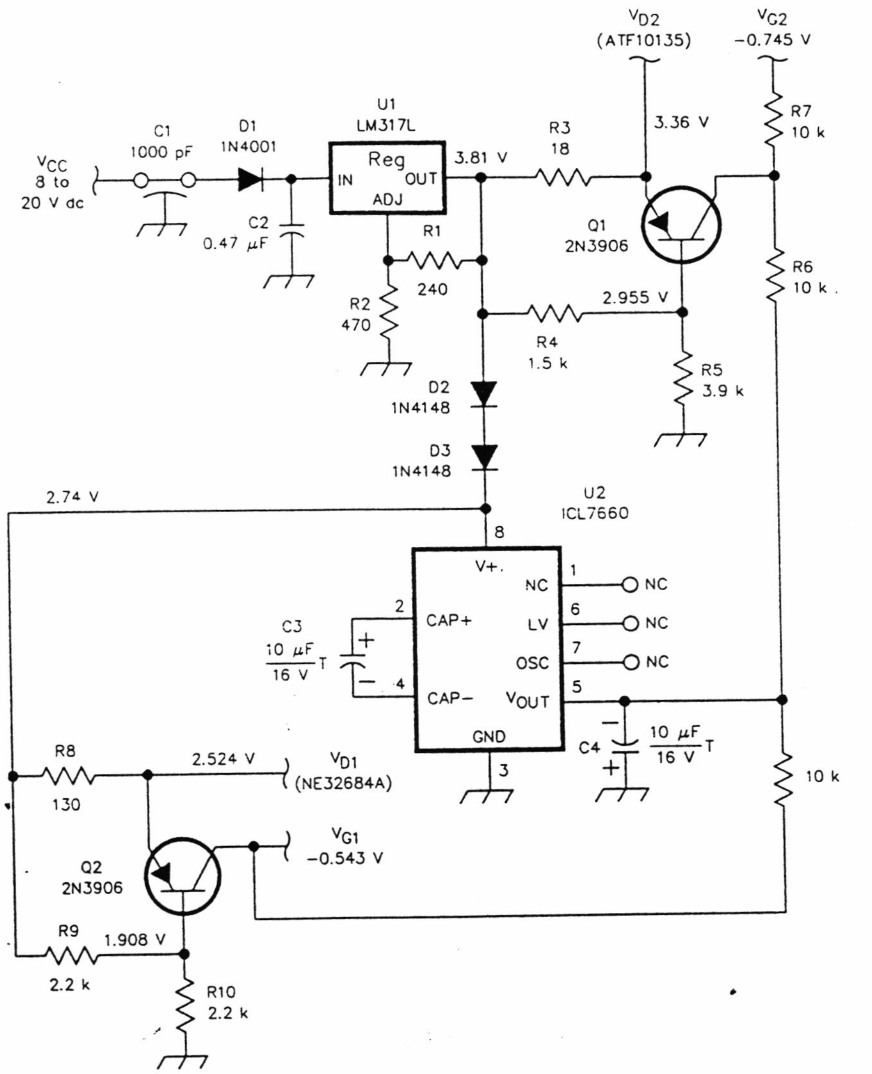


Fig 2—Bias supply circuit for the GaAs FETs.

C1—Feedthrough capacitor. The value is not critical.
Q1, Q2—2N3906, 2N2907, or other general-purpose PNP transistors.
U1—LM317L adjustable voltage regulator IC.
U2—Intersil ICL7660CPA or SI7660CPJ CMOS voltage converter IC.

possible, it is often wise to figure out how the preamp fits into the system. Then, when you fit everything together, you should get the theoretical performance you calculated with a minimum of tweaking.

Perhaps the most important factor is stability. If your preamp oscillates, you are better off without it in your system. Some people actually get it backward; they only consider stability at the design frequency. Actually, the design frequency is where you have the most flexibility. After all, amateurs typically take great pains to match their antenna to the transmitter, so the impedance of the antenna is often pretty close to 50 Ω, at least on the band you are using. So, conditional stability, or stability with good terminations, may be all you need at the design frequency. (This assumes that the second stage has a well behaved input impedance, like that of a broadband MMIC.) On the other hand, impedances can get pretty wild once you get out of band, so your preamp should be unconditionally stable outside the design frequency range.

The gain of the amplifier is another important consideration; the noise figure of a microwave LNA doesn't necessarily set the *system* noise figure. If you do a few calculations, considering

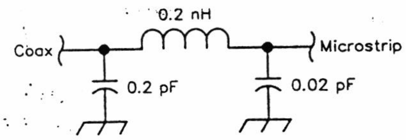


Fig 3—Model of the SMA-to-microstrip transition.

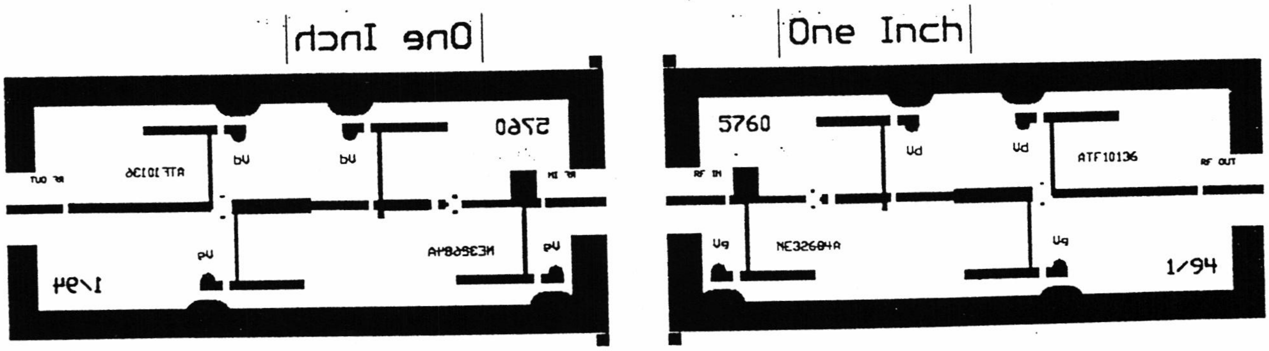


Fig 4—Etching pattern for the preamplifier circuit board. Use 15-mil 5880 Duroid.

that a typical transverter noise figure is 3 or 4 dB, it is pretty obvious that preamps with gains of 10 dB are more useful for winning noise-figure contests than for setting system noise figures. To get that ultra-low 0.4-dB system noise figure, you really need 20 dB of preamplifier gain, rather than 10 dB. To achieve that, you need a multistage amplifier. But the dynamic range of high-gain preamps tends to be less than that of lower-gain designs. Fortunately, the degradation in dynamic range is often acceptable as you go up in frequency, since signals can be separated using antenna directivity. Of course, if you already have a good system or just want to evaluate a device, single-stage pre-amps make more sense.

The gain-versus-frequency curve is also important. Ideally, your design shouldn't have a low-frequency gain peak; it should have more of a bandpass response, centered at the design frequency. This is particularly important if you are using broadband antennas near a cellular phone site. If you need to fix a system with this problem, you might consider using a horn feed, which will act as a waveguide below cutoff for low-frequency signals.

The Actual Design

The RF circuit of the preamp is shown in Fig 1, and the bias circuit is in Fig 2. Based on my success with the device at 10 GHz, I chose the NEC 32684A PHEMT for the low-noise stage. For the second stage I use a Hewlett Packard ATF 10135 GaAs FET. While the GaAs FET has a higher noise figure than the PHEMT, it is much easier to match for both noise figure and return loss. It is also significantly cheaper, while only having a slightly worse noise figure. Using the PHEMT as a second stage would lower the noise figure by just a few hundredths of a dB. If you had another PHEMT device to use, I'd recommend you build a second two-stage preamp and select the best one. It's not unusual for seemingly identical preamps to vary by a tenth of a dB in noise figure.

The circuit is etched on 15-mil 5880 Duroid board. I first built the GaAs FET stage on this board to see if there was any significant reduction in board loss compared to a design on 31-mil stock—there wasn't. And 15-mil board

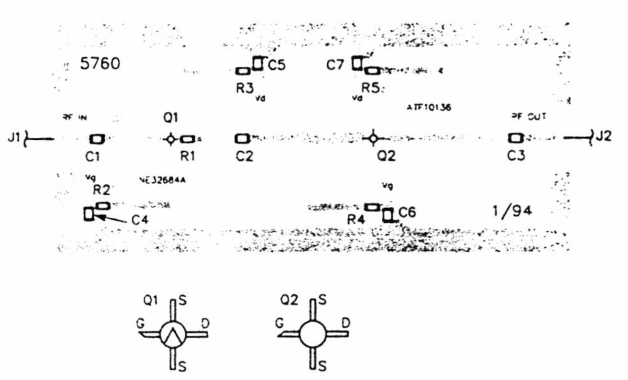


Fig 5—Parts-placement diagram for the 5.7-GHz GaAs FET preamplifier.

allows shorter-length source leads when through-board mounting the transistors—and less consequent inductance. This is an important consideration for the PHEMT input stage, since this device has a lot of gain, even through Ku band.

While it made the design a bit more difficult, I wanted the flexibility of using either stage by itself. With calculated input and output return losses of at least 19 dB, or a 1.24:1 SWR, there shouldn't be any difficulty in using the stages of this design separately. However, if you use something other than SMA transitions, such as an N-to-microstrip transition, you may want to modify the matching circuitry slightly. It was designed for the SMA transition shown in Fig 3. The gain of the PHEMT stage is undoubtedly a little lower than what could be obtained if the stages were designed as a single unit, since some resistive loading is needed to get a stable single-stage design.

Construction

I advise reading about Al Ward's microwave preamps before building this one. You can find his article in either the May 1989 *QST* or the 1992 or 1993 *ARRL Handbook*. The transistor mounting is almost identical, though I decided to mark where to cut the slots. You want your slots to just touch the outside edges of the tiny PC board pads marking the slots. Ideally, you should get the long-lead versions of the devices. Otherwise, I'd recommend you use thin copper foil to connect the pads marking the slots to the ground plane, then solder the transistor to the copper foil. The foil should be 20 mils wide. I used EMI shielding tape with the adhesive removed. The tape is only 1 mil thick.

As with all my preamps, the bias circuitry is built over the ground plane side of the board. I prefer an active bias circuit, such as that shown in Fig 2, but you can use any circuit that doesn't overstress the PHEMT. According to the data sheet, the absolute maximum V_{ds} is 4 V, while the absolute maximum V_{gs} is –3 V. It is a good idea to test the negative voltage generator before connecting a bias supply to a microwave FET. Since the negative voltage is needed to turn the FET off, some designs can stress the transistor with too much drain current if there is a problem with the bias supply.

The file LNX5670.ZIP contains a Postscript file that can be used to print the circuit-board pattern, as well as the Microwave Harmonica circuit model description. This file can be downloaded from the ARRL BBS (203-666-0578), or via Internet from ftp.cs.buffalo.edu in the /pub/ham-radio directory.

Table 1—*Microwave Harmonica* Output Listing

Freq (GHZ) LNZ	NF (dB) LNZ	MS11 (dB) LNZ	MS21 (dB) LNZ	MS22 (dB) LNZ	K LNZ	NF (dB) AMP1	MS11 (dB) AMP1	MS21 (dB) AMP1	MS22 (dB) AMP1	K AMP1	NF (dB) AMP2	MS11 (dB) AMP2	MS21 (dB) AMP2	MS22 (dB) AMP2	K AMP2
0.50	7.05	-15.79	3.79	-29.64	184.66	7.78	-8.61	3.73	-10.40	44.05	8.35	-15.80	8.12	-10.42	999.90
1.00	6.24	-10.50	5.04	-21.42	63.23	6.60	-9.00	4.92	-20.25	18.86	7.21	-10.54	10.28	-20.05	999.90
1.50	5.63	-7.03	5.83	-16.12	30.77	5.77	-7.44	5.17	-25.74	10.69	6.36	-7.04	11.27	-25.18	772.91
2.00	5.46	-5.24	5.64	-12.89	20.31	5.58	-6.95	4.47	-29.42	8.46	6.14	-5.13	10.89	-28.92	376.64
2.50	5.67	-4.53	4.76	-11.13	17.35	6.13	-7.98	2.43	-14.24	10.11	6.66	-4.44	7.96	-13.78	375.61
3.00	5.68	-4.76	5.17	-10.59	12.97	6.85	-12.57	-0.09	-9.17	14.66	7.19	-4.72	4.42	-8.97	506.78
3.50	3.05	-2.75	11.40	-11.81	1.93	3.31	-14.96	6.05	-10.42	3.24	3.25	-2.53	17.07	-9.51	12.94
4.00	1.73	-3.10	13.39	-17.28	1.26	1.23	-7.19	10.16	-13.55	1.19	1.78	-4.24	23.24	-11.17	3.28
4.50	1.32	-4.31	13.95	-42.51	1.21	1.03	-6.32	10.10	-10.51	1.06	1.35	-4.59	24.07	-10.13	1.98
5.00	0.98	-6.48	14.35	-18.10	1.21	0.95	-7.03	9.91	-10.66	1.06	1.01	-4.08	24.53	-10.13	1.53
5.50	0.68	-13.37	14.74	-16.73	1.23	0.86	-12.42	10.13	-20.79	1.06	0.71	-16.24	25.04	-26.48	1.71
5.76	0.65	-20.96	14.61	-19.40	1.25	0.88	-19.03	10.01	-20.12	1.06	0.69	-23.25	24.32	-19.60	1.79
6.00	0.80	-10.27	13.87	-15.08	1.28	0.97	-14.74	9.56	-12.13	1.06	0.83	-13.57	23.69	-17.65	1.78
6.50	1.84	-2.86	10.00	-6.08	1.37	1.35	-8.32	8.15	-8.60	1.08	1.98	-2.45	17.97	-5.31	2.04
7.00	3.75	-1.18	5.10	-3.29	1.50	2.02	-7.07	7.16	-12.92	1.10	4.37	-1.03	11.45	-12.18	4.06
7.50	5.89	-0.71	0.84	-2.19	1.64	2.90	-4.64	5.94	-14.42	1.11	7.18	-0.67	5.89	-8.45	7.62
8.00	7.84	-0.55	-2.58	-1.70	1.83	3.44	-2.51	3.88	-5.62	1.13	9.07	-0.59	2.07	-3.54	8.70
8.50	9.45	-0.50	-5.30	-1.43	2.10	2.83	-2.78	2.99	-4.13	1.22	9.98	-0.61	5.04	-6.58	4.83
9.00	10.84	-0.50	-7.71	-1.26	2.66	3.43	-6.25	1.89	-4.75	1,75	14.18	-0.47	-7.37	-3.80	45.31
9.50	13.04	-0.65	-11.82	-1.04	6.60	14.02	-1.97	-13.85	-3.88	21.03	27.87	-0.66	-27.77	-3.84	999.90
10.00	9.29	-1.50	-7.04	-1.75	7.68	21.31	-1.33	-24.91	-1.89	109.64	23.17	-1.51	-26.53	-1.90	999.90
10.50	9.72	-0.68	-4.50	-2.60	2.32	7.75	-2.18	-8.29	-2.35	4.16	11.57	-0.51	-7.98	-2.12	25.43
11.00	9.62	-0.78	-4.46	-2.79	2.19	4.00	-4.60	-1.14	-5.05	2.10	12.80	-0.77	-9.03	-4.07	63.68
11.50	9.16	-0.83	-4.36	-3.03	2.16	3.99	-3.87	0.12	-7.40	1.73	12.54	-0.90	-8.90	-6.54	80.44
12.00	8.65	-0.89	-4.13	-3.21	2.10	4.68	-2.37	-1.10	-5.91	1.63	10.54	-0.95	-7.21	-5.90	50.40

LNZ: low-noise PHEMT amplifier stage
AMP1: GaAs FET amplifier state
AMP2: Complete amplifier (LNZ and AMP1 cascaded)

NEC

HETERO JUNCTION FIELD EFFECT TRANSISTOR
NE32584C

C to Ku BAND SUPER LOW NOISE AMPLIFIER
N-CHANNEL HJ-FET

DESCRIPTION

The NE32584C is a Hetero Junction FET that utilizes the hetero junction to create high mobility electrons. Its excellent low noise and high associated gain make it suitable for DBS, TVRO and another commercial systems.

EATURES

- Super Low Noise Figure & High Associated Gain
  NF = 0.45 dB TYP., G_a = 12.5 dB TYP. at f = 12 GHz
- Gate Length : $L_g \leq 0.20\ \mu m$
- Gate Width : W_g = 200 μm

ORDERING INFORMATION

PART NUMBER	SUPPLYING FORM	LEAD LENGTH	QUALITY GRADE
NE32584C-SL	STICK	L = 1.7 mm MIN.	Standard
NE32584C-T1	Tape & reel	L = 1.0±0.2 mm	

Please refer to "Quality grade on NEC Semiconductor Devices" (Document number IEI-1209) published by NEC Corporation to know the specification of quality grade on the devices and its recommended applications.

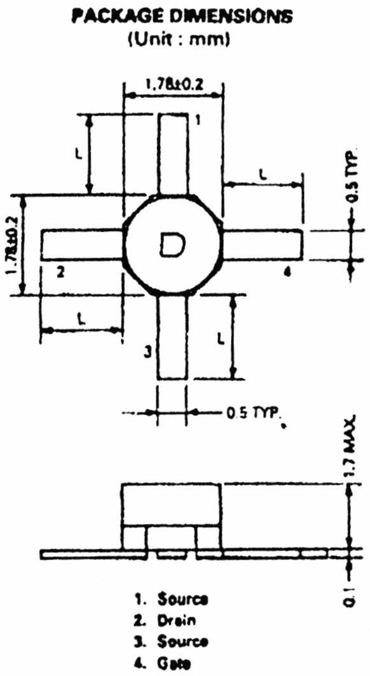
ABSOLUTE MAXIMUM RATINGS (T_A = 25 °C)

Drain to Source Voltage	V_{DS}	4.0	V
Gate to Source Voltage	V_{GS}	−3.0	V
Drain Current	I_D	90	mA
Gate Current	I_G	100	μA
Total Power Dissipation	P_{tot}	165	mW
Channel Temperature	T_{ch}	150	°C
Storage Temperature	T_{stg}	−65 to +150	°C

RECOMMENDED OPERATING CONDITION (T_A = 25 °C)

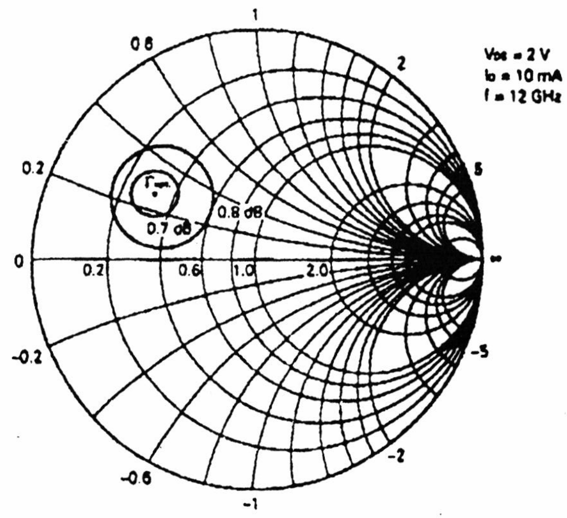
CHARACTERISTIC	SYMBOL	MIN.	TYP.	MAX.	Unit
Drain to Source Voltage	V_{DS}		2	3	V
Drain Current	I_D		10	20	mA
Input Power	P_{in}			0	dBm

PACKAGE DIMENSIONS
(Unit : mm)



1. Source
2. Drain
3. Source
4. Gate

NOISE PARAMETER
<TYPICAL CONSTANT NOISE FIGURE CIRCLE>



<NOISE PARAMETER>
V_{DS} = 2 V, I_D = 10 mA

Freq. (GHz)	NF_{min} (dB)	G_a (dB)	Γ_{opt} MAG.	Γ_{opt} ANG. (deg)	$R_n/50$
2.0	0.29	20.0	0.86	22	0.27
4.0	0.30	18.3	0.76	45	0.25
6.0	0.33	16.5	0.69	70	0.18
8.0	0.36	15.0	0.63	96	0.11
10.0	0.40	13.6	0.59	122	0.08
12.0	0.45	12.5	0.54	147	0.04
14.0	0.54	12.0	0.48	171	0.04
16.0	0.68	11.8	0.40	−165	0.05
18.0	0.85	11.6	0.31	−144	0.06

AMPLIFICATEURS 6 cm

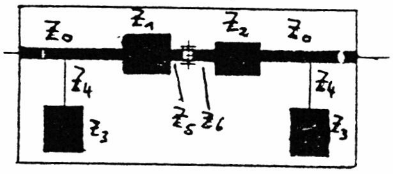
6cm AMPLIFIER

as pre-and low power amplifier

The circuit shown at fig. 2 uses the transistors of the MGF 1400 series. The circuit uses the same material as the other, RT 5870; h=0.79mm. At the input he cam optimizied to minimum noice or to maximum gain. For experiments it is better to do not print Z1 and Z2, but to make them out of 0.3mm copper stipes.Through sliting of Z1 and Z2 on the 50 ohm stripes it is possible to match the input to minimum noice or maximum gain. So it is also possible to compensate reactances in the output or input of following stages. Also other GaAs Fets can be matched by this way.The transformer are soldered very careful now at the experimental founded place (distance from the transistor). The source must be soldered to ground as short as possible. The negative pre voltage can be done by an IC or a battery.

The shown low power amplifier at fig. 3 uses the MGF 1801, input and output are 50 ohm. It is here also better to print not Z1 and Z2, but to test where they are belonging. So you can win some last dBs, and mostly input and output impedance are rare 50 ohm. Tuning is also possible with the drain current. More information also from transveter with subharmonic mixer by DC8UG, Marksburgerstr.58, 5409 Becheln Tel. 02603/70709

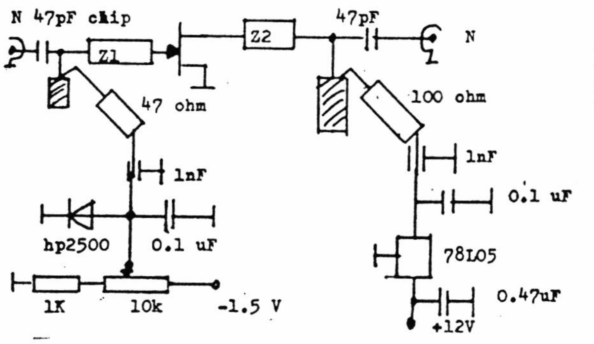
fig.2



Layout MGF 1400-12
Vp max =11 dB
at 1400 , NF=3.3 dB at Z1=18 ohm

Zo=50 ohm l= equal beliebig
Z1 noice =27 ohm; w=6 l=10mm
" power ≤18 ohm; w=9 l=10mm
Z2=22 ohm;w=7 l=10mm
Z3≤20 ohm;w=9 l=10mm
Z4≥100 " w=0.3mm l=11mm
Z5= 50 " l=2mm
Z6= 30 " l=4mm

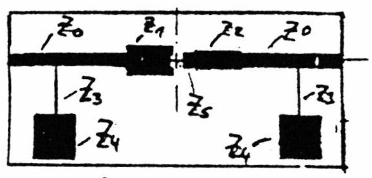
echelle 1:2



negative Spannung muß vor + anliegen!!
negative voltage has to be switched on before + !!!

FIG. 2

power amplifier for 6cm



Layout MGF 1801
I=100mA Uds=6V

Zo=50 ohm w=2.3mm l=equal
Z1=25 " w=6.5mm l=10mm
Z2=37 " w=4 " l= "
Z3≥100 " w=0.3mm l=11mm
Z4≤20 " w=9mm l=10mm
Z5=50 " w=2.3 l=3mm

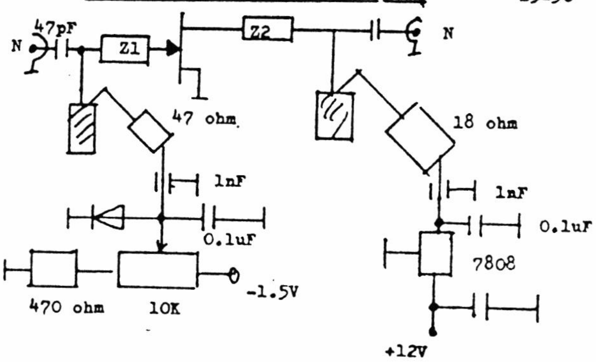


FIG. 3

Two Stage 5760 MHz GaAs FET Amplifier

by DL 7 QY

E.This amplifier is usable as preamp. as well as smallpower amplifier. Maximum rf output on 5760 MHz is abt. 60 mW at 24 dB gain. Best noise figure is adjustable with the open stub in the inputcircuit (min noise figure abt. 1.5 dB). If the amplifier is adjusted for highest gain, the noise figure increases to about 4 dB. The gates are driven by a voltage converter IC ICL7660 (Intersil). Normally gate voltage should stay before the supply voltage reaches the drain of transistor. In this case it isen't necessary because the voltage converter operates already with very low supply voltage and the negative output voltage arrives the gate in time because the RC combination 15 Ohm and 10 uF. For lowest noise figure adjust the drain current to 10 mA of each transistor. When operating as small power amplifier increase the current to 20-30 mA. The variable coupling capacitor is home brew, a silverplated copper band (3mm x 7mm x .5mm) is soldered on the printed transmission line, gate T2. Take care under adjustment of this capacitor, because you may not touch the drain transmission line from T1 (4 V positive voltage!). The best protection under adjustment is to isolate the drainline T1 using a thin Teflon foil. Variing of this capacitor moves the center transmission frequency.
As PCB material is used DUROID RT5870. Transistors NE72089 are obtainable from DL7QY (74,--DM each). Fig 1 shows the circuit diagram. Fig. 2 shows the PCB layout and Fig 3-5 the transmission curves and gain characteristics.

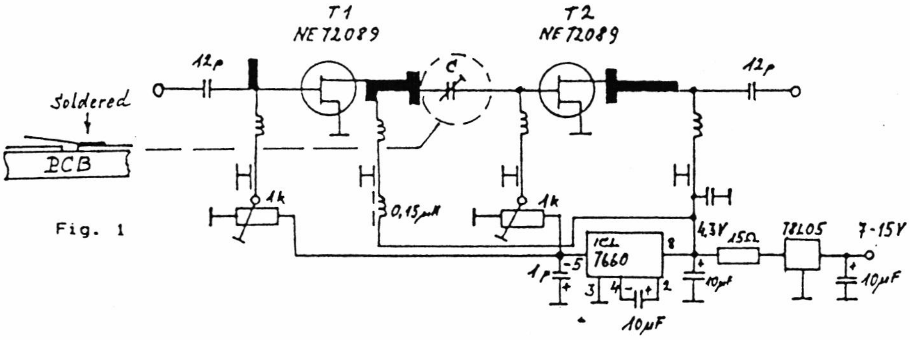


Fig. 1

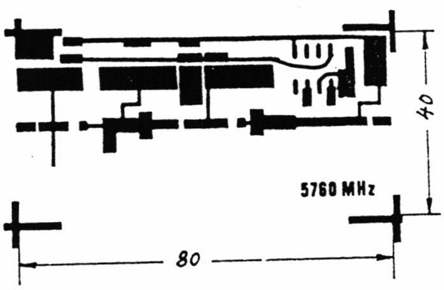


Fig. 2 layout

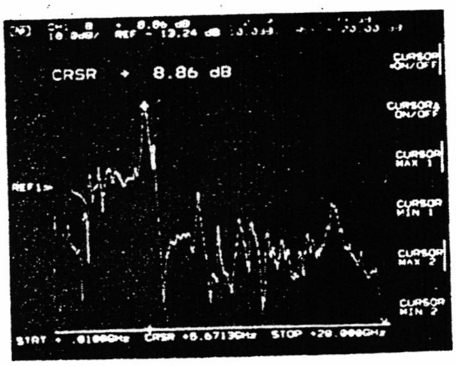


Fig. 3 Transmission curve
10 MHz - 20 GHz.

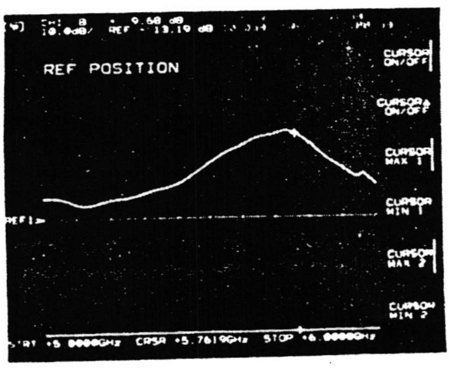


Fig. 4 Transmission curve
5 - 6 GHz.

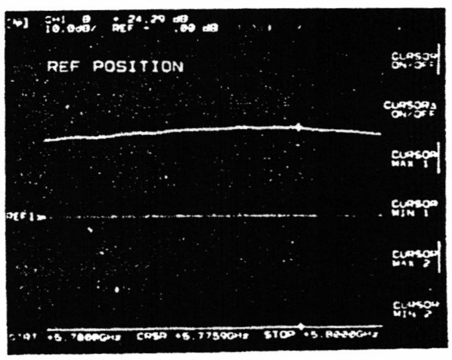


Fig. 5 Transmission curve
5.7 - 5.8 GHz

5.7 GHz Amplifiers

Toshihiko Takamizawa, JE1AAH
Parktown 21-502 946-16
Kitahassaku-cho, Midori-ku
Yokohama, 226 Japan

Toshihiko, JE1AAH, has provided some excellent articles covering his work as a microwave amateur. Because he is a new author in DUBUS, we would like to introduce him. Toshihiko works as a computer systems engineer for CocaCola, Japan.
In his free time he designs and builds amateur microwave gear. DUBUS will publish his 5.7 and 10 GHz systems and some interesting PLL-stabilized DRO-oscillators.

Toshihiko, JE1AAH, at microwave work

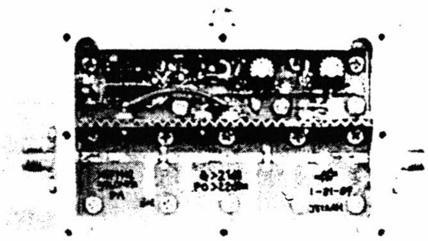


Figure 6/Bild 6: 5.7GHz HPA

Substrat: 0,8 mm
Cu-Clad
Keene

3. 5.7 GHz HPA

The HPA is a medium power amplifier for the 6 cm amateur band. It uses the low price MGF1402 transistors and achieves a maximum power output of 22 dBm (150 mW). At 100 mW output (20 dBm) the gain is 21 dB. See Figure 8 for gain versus frequency. Input and output matching networks are made from 50 Ohm series and parallel open shunt stub lines. The interstage matching first transforms to 50 Ohms and then to the conjugate of the second stage input impedance. The bias feed lines are made from fine copper wire with 0.15 mm diameter and 12 mm length instead of printed quarter wave lines. This is used in professional amps also. See layout on picture on the right (Figure 5) and circuit diagram in Figure 7.

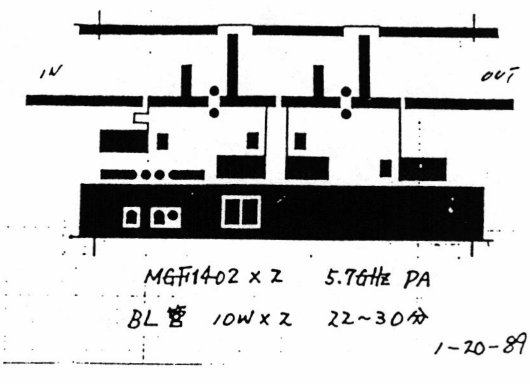
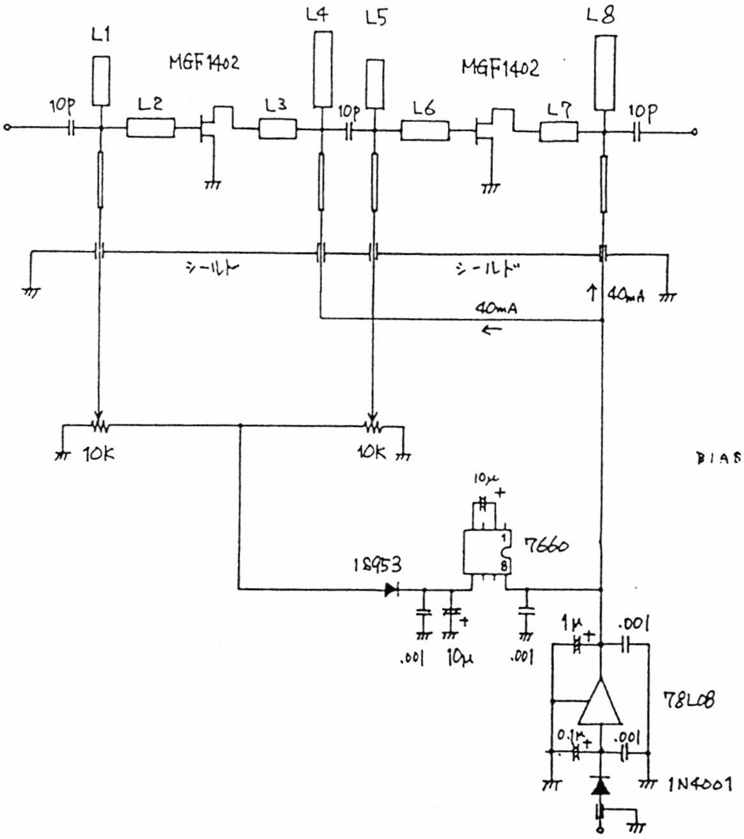


Figure 5/Bild 5: Etching Pattern/Layout HPA

The MGF1402 is a reasonable medium power FET by applying 8 Volts to the drain. Idling cureent for each FET is 30 mA, which increases to 40 mA with 22 dBm output. So it's a good alternative to the very expensive MGF1601 transistors by operating on the edge of the maximum ratings, which has been 'good' amateur style from the very beginnung of our hobby. Mechanical construction is like the LNA's. See Figure 6.

FIGURE 7(BILD 7): Diagranm/Schaltung HPA

5.7 GHz HPA



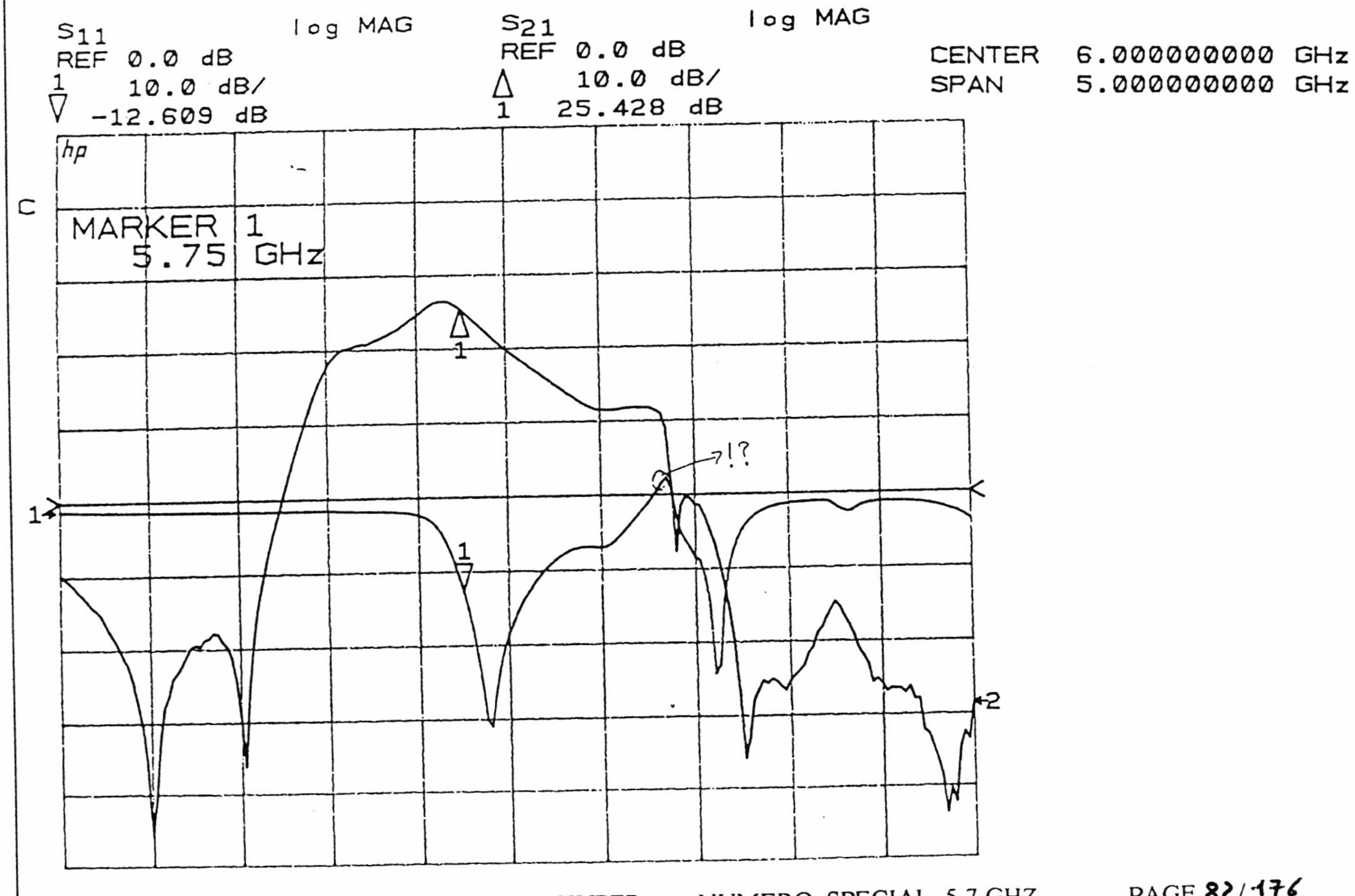
	L1	L2	L3	L4	L5	L6	L7	L8	
λ	.163	.116	.068	.321	.163	.116	.068	.321	5.76GHz
Width	2.2	→							mm
Length	5.9	4.2	2.45	11.6	5.9	4.2	2.45	11.6	mm

all 50Ω imp. line

Length of
between L4 & L5 is free,
it is 50Ω imp. line

BIAS is feeded by fine copper fire, its Length is 11~13mm

FIGURE 8 (BILD 8): Gain @ F/Frequenzgang HPA



AMPLIFICATEURS D'EMISSION 5760 MHz FC1GAS

(Hurc Infos No 31 Juin 1988)

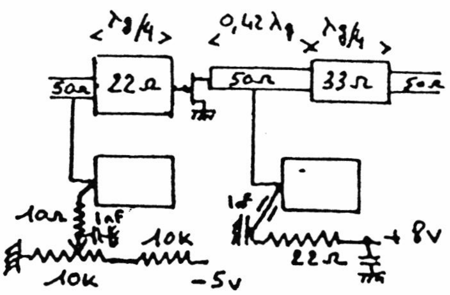
La description suivante concerne l'utilisation du Transistor GaAsFet de moyenne puissance MGF 1601 de MITSUBISHI

I MODELE A 1 TRANSISTOR

A 5760 MHz, avec $V_{DS} = 6V$ $Id = 100mA$, le MGF 1601 est pratiquement inconditionnellement stable. c'est à dire qu'aucune charge ne peut l'entrainer vers l'oscillation. (Rappels des conditions de stabilité d'un transistor

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2\,|S_{12}|\,|S_{21}|} \quad \text{avec } \Delta = S_{11} S_{22} - S_{12} S_{21}$$

- si $K > 1$ le système est inconditionnellement stable
- si $K < 1$ le système est conditionnellement stable, c'est à dire qu'une sélection des charges est nécessaire.) Dans le cas present $K = 1,26$



Performances

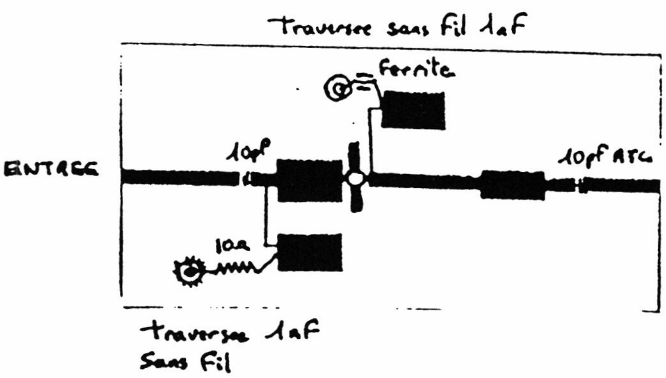
$V_{DS} = 6V$ $Id = 100mA$

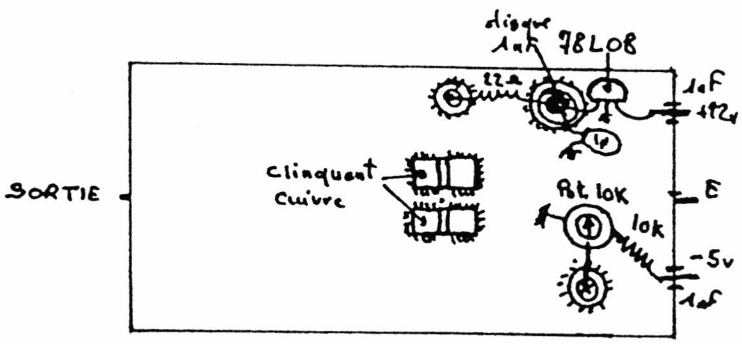
Pin .. jusqu'à 50mW pour 1dB de compression

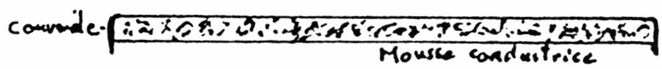
Gain 8 à 8,5 dB

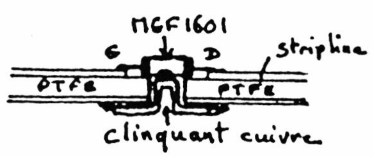
Réalisation sur support verre-téflon $\varepsilon_r = 2,45$ épaisseur 0,762 mm

Boitier 37 x 74 x 30









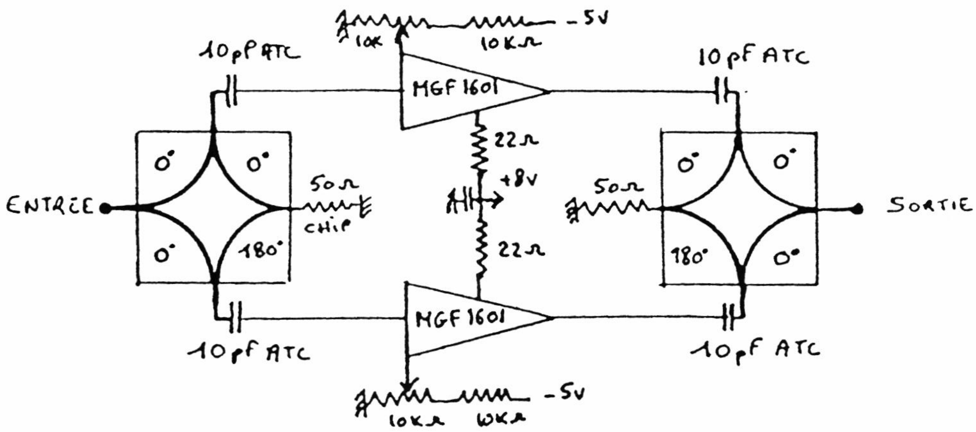
Sur le couvercle du boitier coté circuit prevoir un absorbant (Mousse anti statique pour CI)

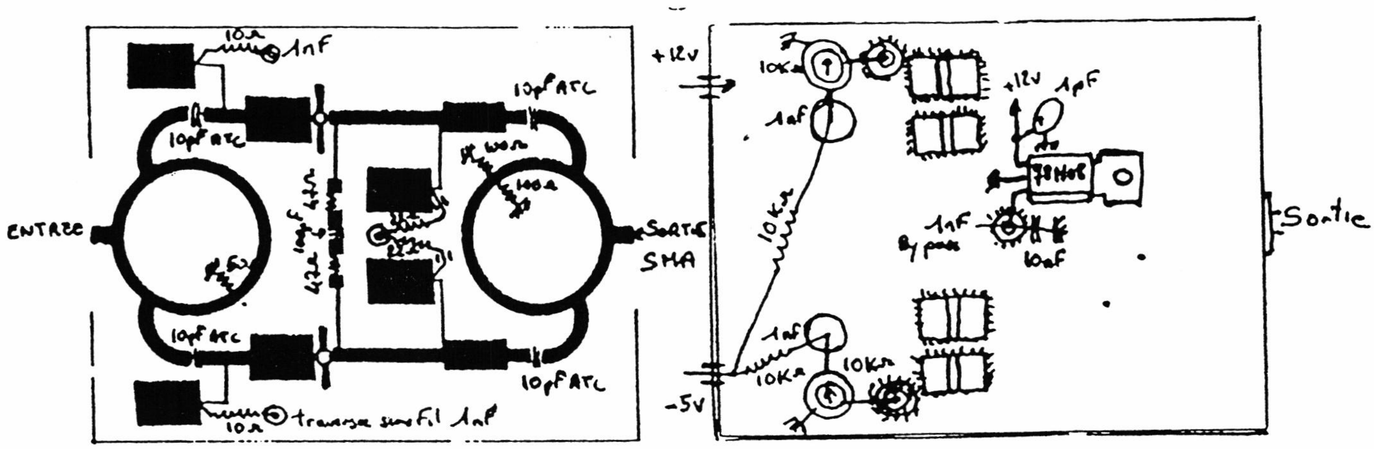
Montage du transistor : conserver toute la longueur des pattes "Source" (Il y a 600mW à dissiper..) La dissipation max du transistor est de 1W

II Modèle à 2 transistors en parallèle

Il s'agit tout simplement du couplage en // de deux amplificateurs. Le couplage est réalisé grâce à 8 coupleurs 3 dB en 6λ/4 qui introduit une bonne isolation entre les deux voies.

La réalisation est encore faite sur un support verre téflon $\varepsilon_r = 2{,}45$ $e = 0{,}762$ m dans un boîtier 55 x 74 x 30





L'argenture des 2 faces du CI est recommandée

Les résistances chips sont des 1/8 W CMS

Le réseau 2 x 47Ω + 100pF entre les 2 drains, limite les risques d'"accrochages" en VHF..

Même recommandations que pour le montage précédent : clinquant cuivre pour les sources des transistors ; mousse absorbante sur le couvercle côté circuit.

Afin de conserver les qualités des capas ATC, il est recommandé de les positionner de manière à ce que les "armatures" soient perpendiculaires à la ligne afin d'éviter les "résonances parallèles parasites"

- Réglages : Appliquer ~ -1V sur les 2 grilles, connecter le + 12V et ajuster les tensions "grille" de manière à avoir les 2 courants "drain" à 100mA chacun. injecter la HF, ça marche...

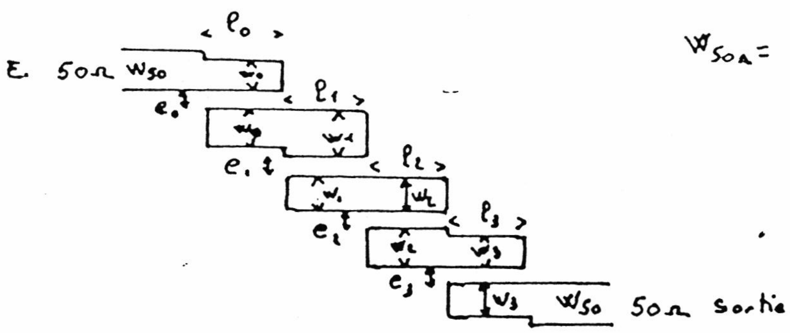
- Performances : Le gain est au moins de 8dB et on peut sortir jusqu'à ≈ 700mW avec ≈ 1,5 dB de compression.

La bande passante étant de plusieurs centaines de MHz, il est préférable d'interconnecter les différents étages par l'intermédiaire de filtres (stripline ou cavité). Pour ma part j'utilise dans la chaîne émission 3 cavités DK 2 AB (voir DUBUS 2-84 p. 84/85)

Filtre stripline sur verre téflon $\varepsilon_r = 2,45$ e = 0,762 mm

Fcentrale : 5760 MHz Bande passante -3dB : 30 MHz Ondulation : 0,1 dB

Réjection OL (5616 MHz) : -40 dB

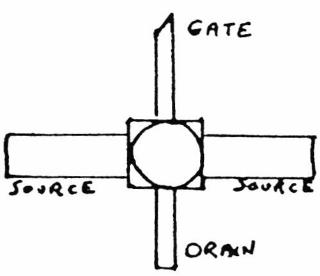


$W_{50\Omega}$ = 2,19 mm

	w_x	e_x	ℓ_x
x = 0	2,1634	0,8662	8,776
x = 1	2,1892	5,819	8,743
x = 2	2,1892	5,819	8,743
x = 3	2,1634	0,8662	8,776

Le dessin doit être fait à l'échelle 4 (ou plus) puis réduit au banc photo....

MGF 1601



Valeurs a ne pas depasser

V_{GD} = -8V

I_D = 250 mA

P_{tot} = 1W

A 5760 MHz 6V 100mA

S_{11} = 0,665 $\angle -178°$ S_{12} = 0,063 $\angle 26°$

S_{22} = 0,370 $\angle -126°$ S_{21} = 2,63 $\angle 34°$

5.7 GHz HPAs using IM5964-3A

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+81-45-931-5757

Feed Point No Aout/Sept 94

I have discovered Avantek IM5964-3A from States several years ago and it is very cheap indeed. I have bought many of these this year and decided to utilize it for 5.7 GHz application. I developed various HPAs for 5.7 GHz using this hot device at reasonable cost by utilizing C band power GaAs FET at one of 2 stage HPA. There have been 4 HPAs developed. They are:

A1 MGF1302-MGF1601A 20 dB gain, 25 dBm Pout

A2 MGF0904A-5964-3A 12dB gain, 35.5 dBm Pout

A3 NE800295-5964-3A 16 dB gain, 36 dBm Pout

A4 5964-3A 8 dB gain 36 dBm Pout

Common data: All of HPA utilize Matsushita R4726 PCB for the circuit board. It has about 3.5 of permitivity and I selected thickness of 0.8 mm. The PCB is available from JE1AAH upon request. Machine milled case was used for all HPA and it is also available from JE1AAH. You should use heat sink as big as possible for the 4W HPA. Because thermal stress is huge and machine milled case itself is absolutely NOT enough size for heat capacity of the HPA.

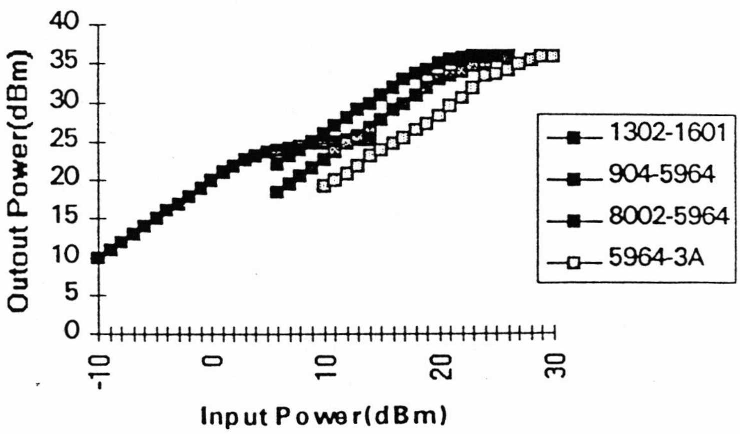
A1: MGF1302-MGF1601A 2 stage HPA It employs previously developed MGF1402-MGF1402 2 stage general purpose amplifier board. Only one tuning is required at output circuit of MGF1601A by additional open circuit stub. Cut-out Source lead of MGF1601A is a good material for the stub. Position of the stub is just after current open stub and length of 3 to 4 mm is needed. Vds I have set is 9V by 78M09. A 22 ohm resistor is inserted at MGF1302 Drain circuit to protect it from over voltage. Ids for MGF1302 is about 50 mA and it makes 1V drop of Vds. MGF1601A is running under 9V of Vds. Eventually, both 8V and 9V is over specification, but it is needed to boost saturated power level. Vds of 6V for MGF1601A only produces about 150 mW and it just matches to manufacture's specification.

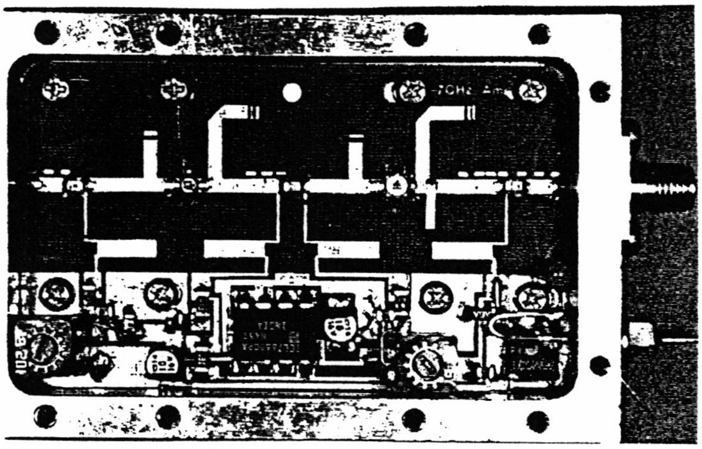
A2: MGF0904A-IM5964-3A 2 stage HPA MGF0904A has been originally developed for S band application. It is very good device for 2.4 GHz application with reasonable price. Additionally, very good point of MGF0904A is that it has several dB of gain at 5.7 GHz while MGF0905A is completely NO good at 5.7 GHz. The board used is etched and has bias circuitry. However, tuning is required by small copper foil on the 50 ohm track because S parameter of MGF0904A is unknown at 5.7 GHz due to out of range from manufacturer's specification. Tuning points for MGF0904A are at gate and at drain just next to the device by single open circuit stub for one each. The stub is attached(closed) to the device for both gate and drain. Length of stub is about 3 to 4 mm. Tuning is also required at 5964-3A gate, too. Vds was set to 8V and Ids for 0904A was set to 400 mA and 1A for 5964-3A with NO RF power added. As estimated, gain is about 12dB due to lower gain at 0904A stage. However saturated power reaches at 4W with help of more driving level than A3 HPA.

A3: NE800295-IM5964-3A 2 stage HPA Since NE800295 is designed to be used at C band application, this amplifier performs best. Gain is over 15 dB and Psat exceeds 4W. Tuning is required at NE8002 gate and 5964-3A gate. Vds was set to 8.5V and Ids for 8002 was set to 300 mA and 1A for 5964-3A with NO RF power added.

A4: IM5964-3A single stage. This amplifier performs just close to manufacturer's specification. It was needed 1W to fully drive the device and over 4W of saturated power could be obtained. Drain current was set to 1A with NO RF power added.

Input/Output performance of various 5.7GHz HPA by JE1AAH 06/12/94





MGF 1302 – 1601 A1

Single Stage 5964

JE1AAH

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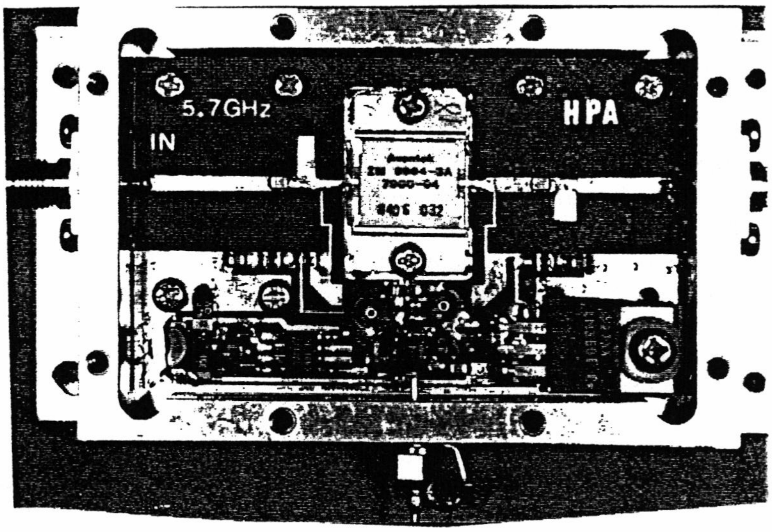

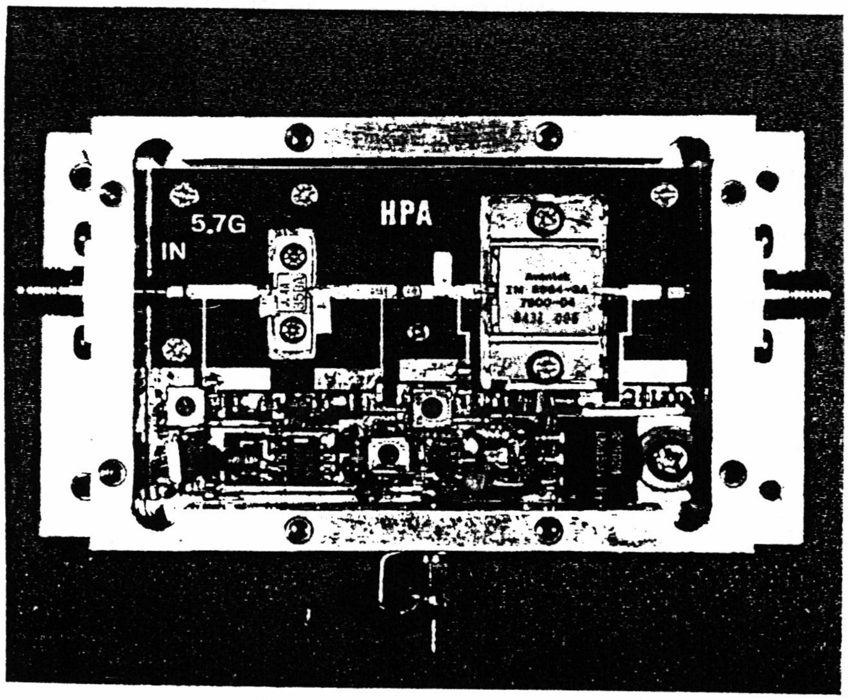

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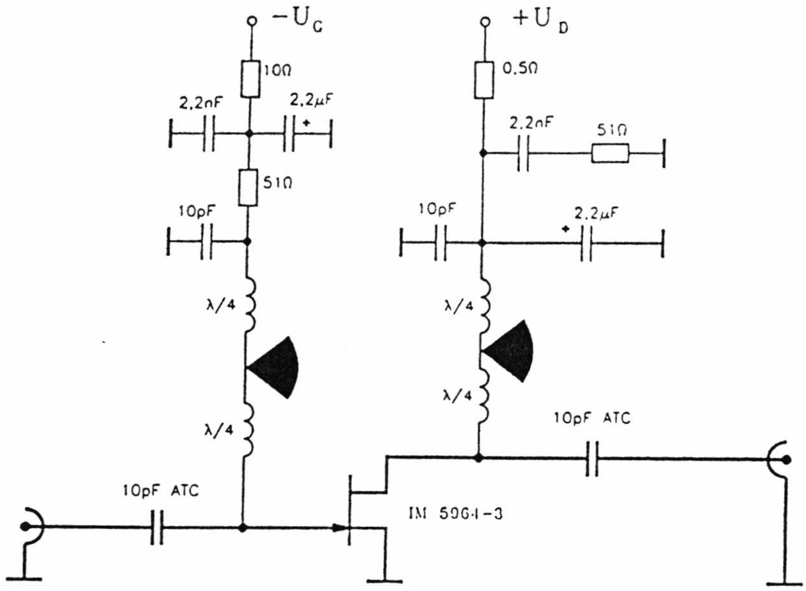
5.7GHz Power Amplifier IM57-4

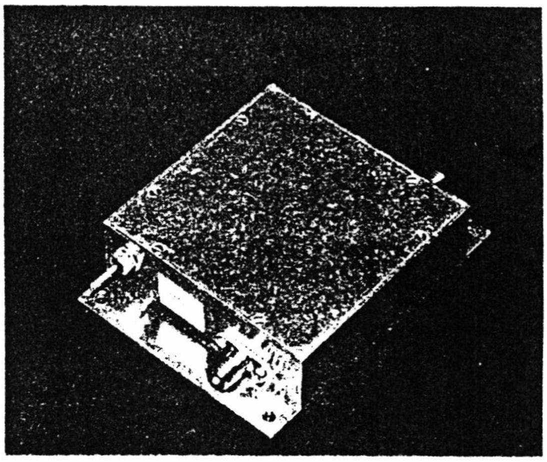
Abstract: The power Amplifier IM57-4 provides a 1db compression output power of 4.5W on 5.76 GHZ with a supply voltage of 12 and 8.5dB gain. It uses a IM5964-3 GaAs FET from Avantek und is constructed on Teflon board in a machined aluminum box. Philipp Prinz, DL2AM, produces and sells this amplifier.

1. Description

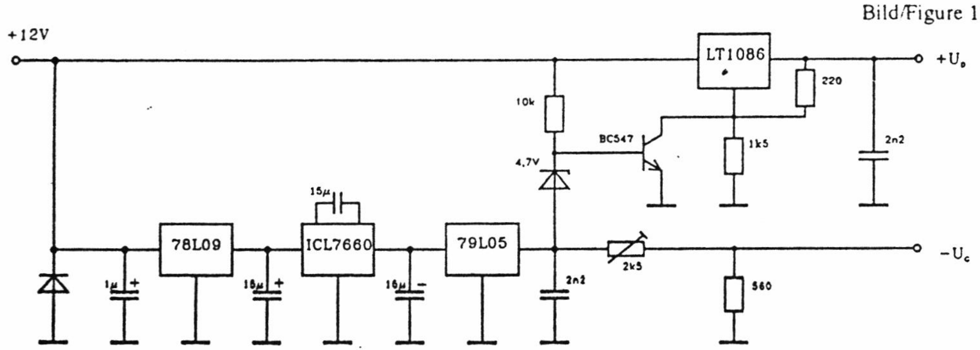
The amplifier is constructed around the internally matched power FET IM5964-3 from Avantek (Fig. 2). Input- and output striplines as well as the DC-feedlines are printed on 0.5 mm



Bild/Figure 2: IM57-4: Circuit Diagram



Bild/Figure 1: 6cm Power Amplifier IM57-4



Bild/Figure 3: IM57-4: DC-Circuit

thick RT-Duroid (Fig. 4). The circuit diagram in fig. 2 shows the special decoupling measures for low frequency stability. These are rather important because with insufficient decoupling the amplifier would be prone to oscillations because of the high gain of the device below 1GHz. The DC-circuit regulates(Fig. 3) the B+, provides negative voltage for the gate and a shutdown protection in case of failure of the gate voltage. Special tuning straps (see Fig. 4) allow for individual tuning for maximum output power and good input return loss.

2. Leistungsdaten/Performance Data

2.1 Mechanische Daten/Mechanical Data

Abmessungen/Dimensions:	79x59x20mm
Gehäuse/Cabinet:	Aluminium Gefräst/Machined Aluminum Box
Kühlkörper/Cooling:	R_{th} < 1,5K/W bei T_U = 25 °C
Anschlüsse/Connectors:	SMA-Buchsen für HF
	Lötstift für 12V

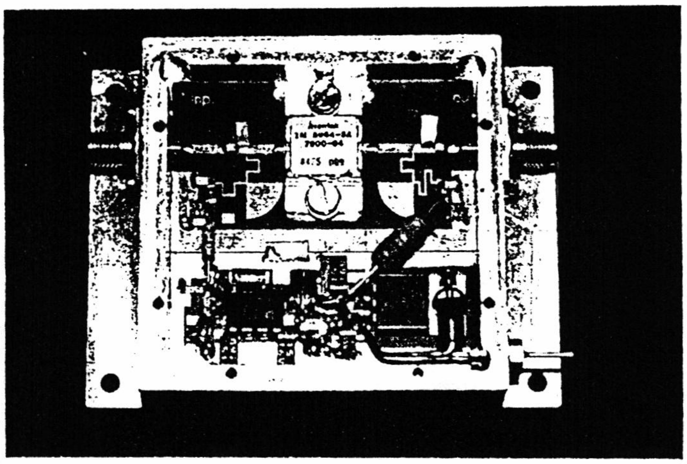
2.2 Elektrische Daten/Electrical Data

Eigenschaft/Item	Symbol/Unit	min	typ	max
Betriebsspannung/Supply Voltage	Ub/V	11	12	15
Stromaufnahme/DC-Current	Ib/A	1	1,2	1,6
Frequenzbereich/Frequency Range	f/GHz	5,6		6,0
Verstärkung/Small Signal Gain	|S21|/dB	8,0	9,5	
Eingangsanpassung/Input Return Loss	|S11|/dB	-10	-13	
Ausgangsleistung/Output Power	P1db/W	3,0	4,5	

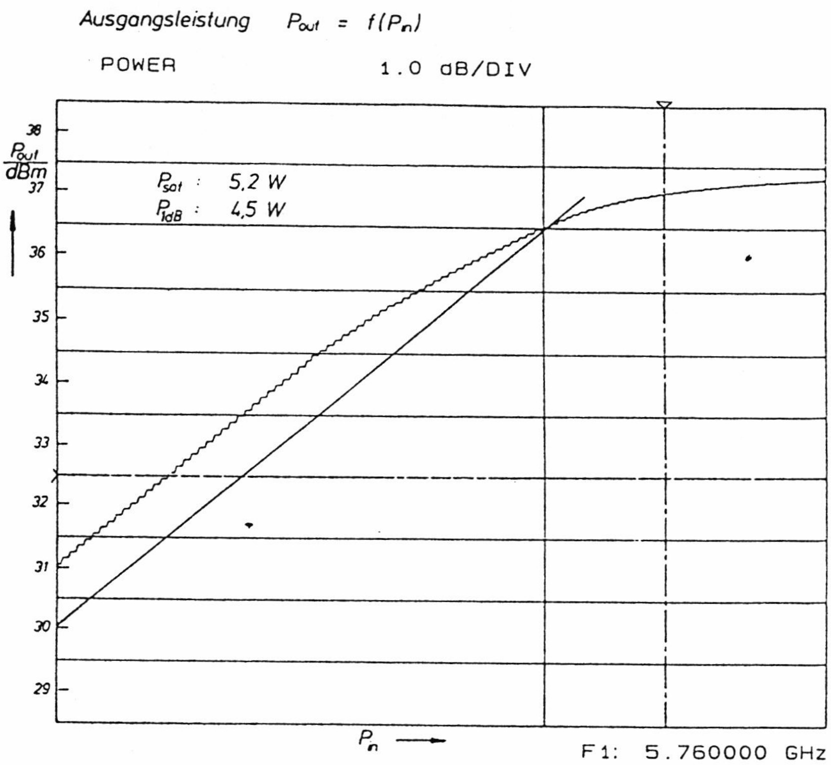
3. Evaluation

Construction of the IM57-4 leaves a good impression. ATC-Chips are used for the coupling caps for low loss and high reliability. The measured performance gives good values for output power (Fig. 5), gain (Fig. 6) and input return loss (Fig. 7).

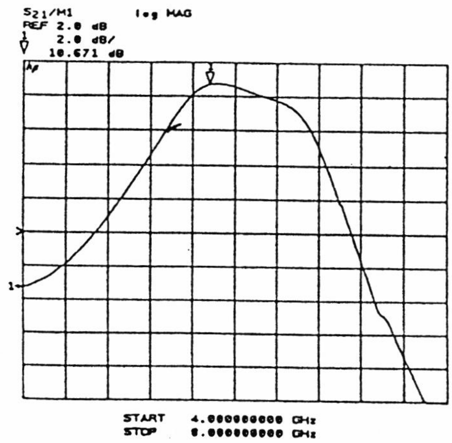
4. Bezug/Manufacturer



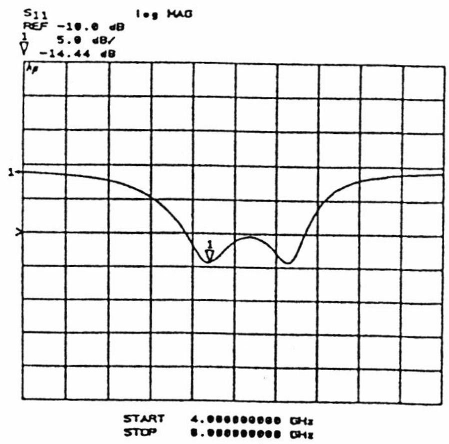
Bild/Figure 4: Inside the IM57-4



Bild/Figure 5: IM57-4: Output Power versus Drive



Bild/Figure 6: Small Signal Gain |S21|



Bild/Figure 7: Input Return Loss

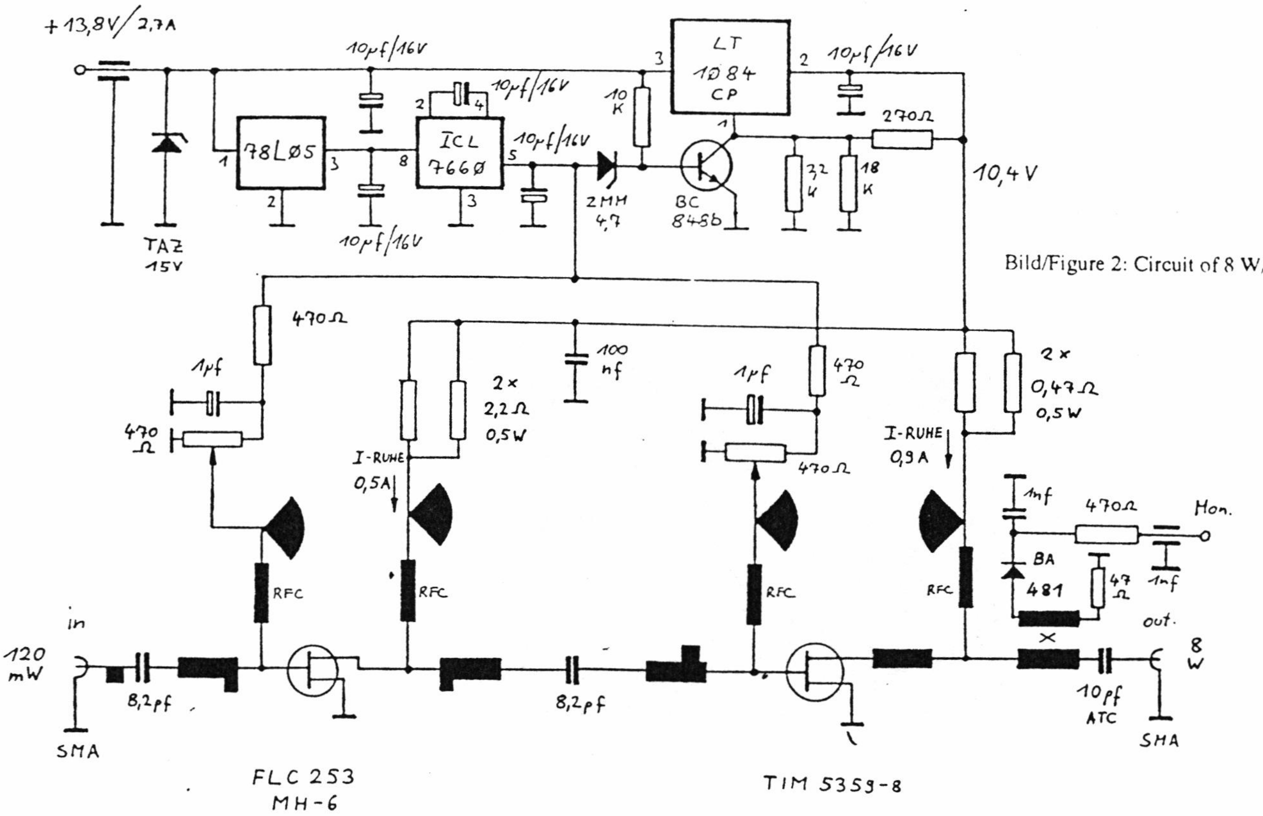
Ready made units can be ordered from Philip Prinz, DL2AM, Riedweg 12, W-7970 Leutkirch 3, Tel.: (++49)7576-294/Fax: (++49)7567-1200 . Price is about DM 500,-.

8 W GaAs-FET Amplifier for 6cm

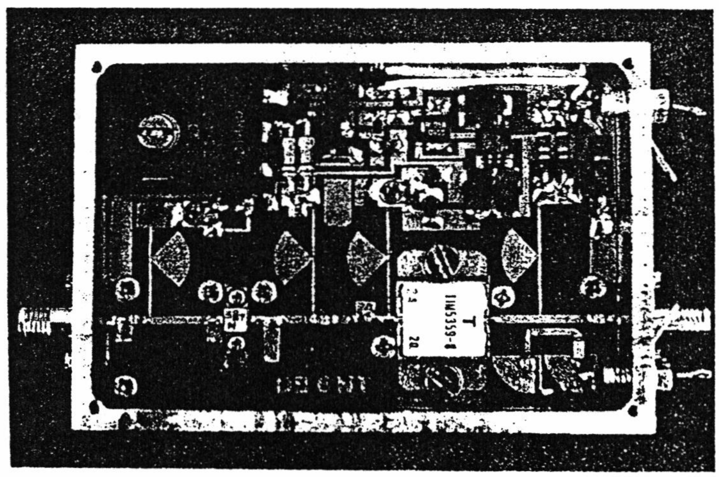
Michael Kuhne, DB6NT
Birkenweg 15, W-8674 Naila 2

1. Design

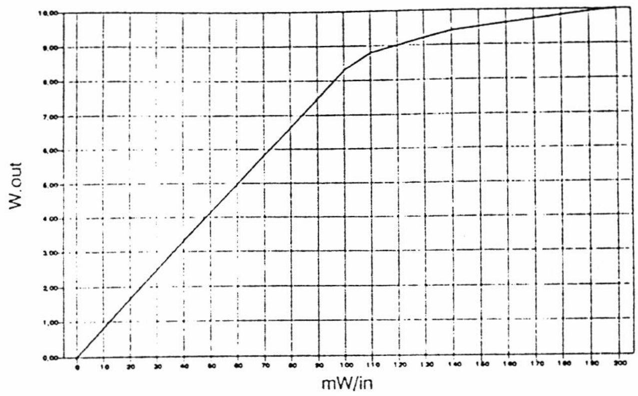
The two stage power amp uses the FLC253MH-6 in the first and the TOSHIBA TIM5359-8 in the final stage. It provides more than 8 W output with 120 mW drive. The TIM5359-8 is internally matched to 50 Ohms. A fail safe circuit provides power shutdown in case of failure of the negative gate voltage. Output power is sampled by an integrated directional coupler and a Schottky-Diode BA481.



Bild/Figure 2: Circuit of 8 W/6cm Amplifier



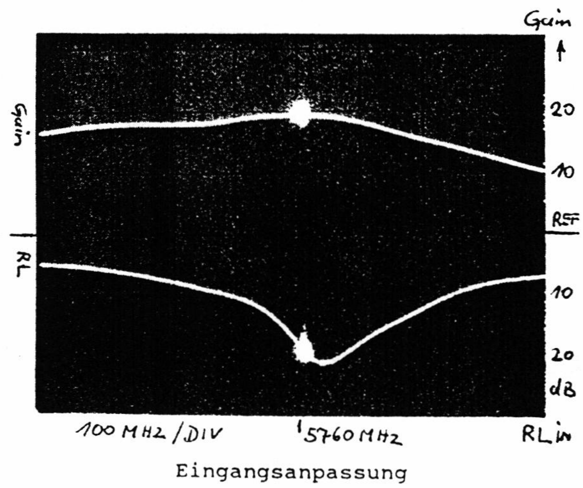
Bild/Figure 1: 8 W Amplifier for 6cm



Bild/Figure 3: Pout versus Pin

2. Construction and Tuning

The PCB (0.5mm RT5870 Duroid) has to be mounted into the machined aluminum cabinet with M2 screws and some silver paste. For the FET's it's necessary to provide nuts to allow for flushing of the leads to the top of the PCB. The connectors are SMA-Stripline variety with a 3 mm PTFE insulator. The coupling caps are simple SMD-type (0807). The aluminum cabinet has to be mounted onto a heatsink with sufficient heat capacity.



Bild/Figure 4: Gain versus F

3. Results

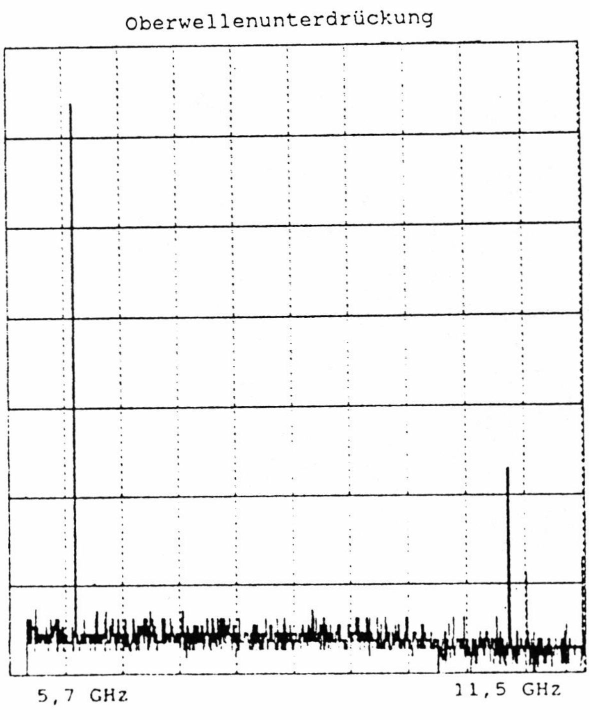
The amp is very stable both from electrical and thermal points of view. No oscillation could be observed even with open input. Output power is in excess of 8W and has a saturation value of 10W (Fig. 3). Gain is flat from 5.4 to 5.8GHZ and is about 19 dB (Fig. 4).

4. Teile/Parts

PCB: From DB6NT

FETs: FLC253MH-6 from MELATRONIK, Tel.: 089/3104076

TIM5359-8 from TRICOM, Tel. 08161/86066



Bild/Figure 5: Harmonic Suppression