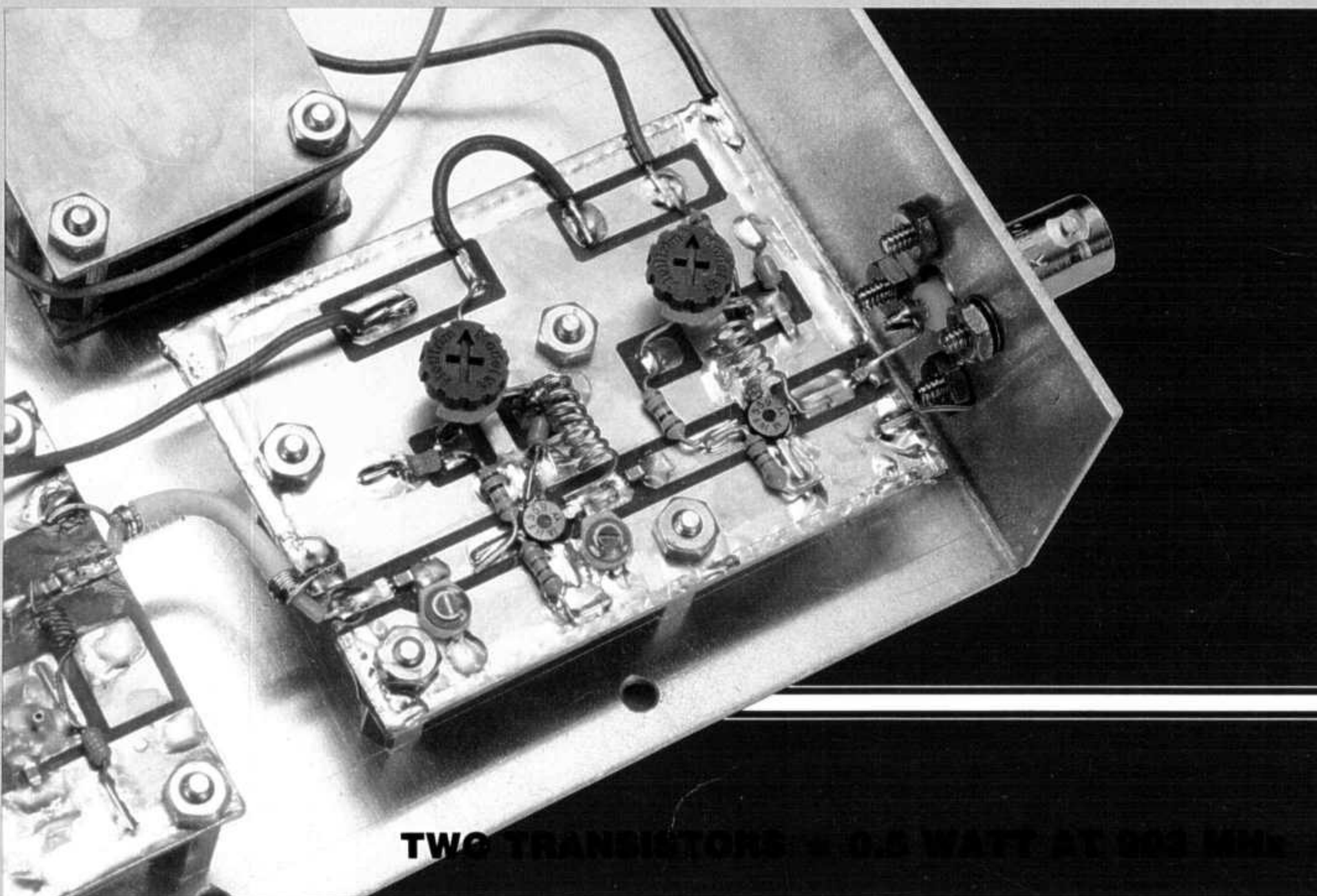
QEX 76

$1.75

JUNE 1988

ARRL Experimenters' Exchange and AMSAT Satellite Journal



QEX: The ARRL
Experimenters' Exchange
American Radio Relay League
225 Main Street
Newington, CT USA 06111

Bulk Rate
U.S. Postage
PAID
Hartford, Conn
Permit No. 2929

QEX (ISSN: 0886-8093) is published monthly by the American Radio Relay League, Newington, CT USA.

Offices

225 Main St, Newington, CT 06111 USA
Telephone: 203-666-1541
Telex: 650215-5052 MCI
Electronic Mail: MCI MAIL ID:215-5052
(user name ARRL)

Subscription rate for 12 issues:
In the US by Third Class Mail:
ARRL/AMSAT Member $8, nonmember $16;
US by First Class Mail:
ARRL/AMSAT Member $13, nonmember $21;
Elsewhere by Airmail:
ARRL/AMSAT Member $23, nonmember $31.

QEX subscription orders, changes of address, and reports of missing or damaged copies may be marked: *QEX* Circulation.

Members are asked to include their membership control number or a label from their *QST* wrapper when applying.

TABLE OF CONTENTS

COLUMNS

ABOUT THE COVER

Ten milliwatts in, half a watt out: Microstriplines and two MRF559s on a 2- × 3-inch PC board work 903-MHz magic in this linear amplifier designed and built by KH6CP. To find out how to build yours, see page 3.

THE AMERICAN RADIO RELAY LEAGUE, INC

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interest in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of the radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

ARRL is an incorporated association without capital stock chartered under the laws of the State of Connecticut, and is an exempt organization under Section 501(c)(3) of the Internal Revenue Code of 1954. Its affairs are governed by a Board of Directors, whose voting members are elected every two years by the general membership. The officers are elected or appointed by the Directors. The League is noncommercial, and no one who could gain financially from the shaping of its affairs is eligible for membership on its Board.

"Of, by, and for the radio amateur," ARRL numbers within its ranks the vast majority of active amateurs in the nation and has a proud history of achievement as the standard-bearer in amateur affairs. As its membership journal, ARRL publishes *QST* monthly.

A bona fide interest in Amateur Radio is the only essential qualification of membership; an Amateur Radio license is not a prerequisite, although full voting membership is granted only to licensed amateurs in the US and Canada.

Officers

President: LARRY E. PRICE, W4RA
PO Box 2067, Statesboro, GA 30458

Executive Vice President: DAVID SUMNER, K1ZZ

AMSAT®
RADIO AMATEUR SATELLITE CORP

President: VERN "RIP" RIPORTELLA, WA2LQQ

Purposes of *QEX*:

1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio field
3) support efforts to advance the state of the Amateur Radio art.

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and correspondence for publication in *QEX* should be marked: Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and double spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black-and-white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

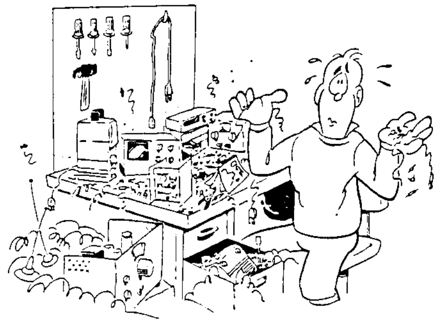
Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking. . .

Wear A TC Hat

Dear ARRL:

I wish to thank the ARRL for helping me locate Mr Lyle Wilson, KØKSE, Technical Coordinator in Jordan, Minnesota. I have been having difficulty with TVI and RFI and Mr Wilson has guided me through the procedures to correct the problem. Mr Wilson has given me excellent advice concerning these issues. I feel he is to be commended for his efforts and hope there are many more people like him available for other Amateur Radio operators.
73, Jeffrey L. Dube, NØGZK



Like to help others? Like to help yourself? Yes? Then, try on a TC hat for size. League *Technical Coordinators,* and *Assistants* are volunteers who help others with technical questions, rig problems, and RFI troubles. The salary: Letters like the above, and the warm feeling in knowing that you are giving something back to Amateur Radio. The fringe benefit is the community you develop with other technical volunteers, broadening your own learning opportunity.

Appointed by your Section Manager (see page 8 of *QST* for a list and brief description of the ARRL Field Organization), you work within the section to do several things. Encourage amateurs to share their technical achievements with others at club meetings, hamfests and through newsletters. Work closely with VHF/UHF and specialized-mode pioneers for technical advancement. Advise instructors on technical parts of amateur exam courses. Handle technical information requests. Sit on cable TV or RFI committees. Give talks at club meetings and hamfests. And more—use your imagination as guide.

A new *Technical Coordinator Manual* by Wisconsin Section Manager Rich Regent (K9GDF), gives a wealth of practical tips for the novice and veteran; it's the major tool in your tool box. You're also provided a copy of the newest *ARRL Handbook,* and the *Radio Frequency Interference* book. You get *QEX* and *FIELD FORUM,* which offer national forums for Technical Coordinator information exchange. And, of course, you've got support from the ARRL Technical Department.

Currently there are 62 Technical Coordinators and more than 320 Assistant Technical Coordinators nationwide, and counting. Since the inception of the enhanced Field Organization in 1983, interest in section technical programs, and information has blossomed. Packet radio, and other digital applications have been contributory, and there's no end in sight. TCs are helping us to fulfill our promise to FCC and the public of advancing the technical phases of the radio art.

It's no accident that this little recruitment pitch appears in this editorial slot. We know your profile: as a *QEX* technically proficient, interested in experimentation, and what others are doing. You like to learn, you care about Amateur Radio and want to promote its ability to further technology. You probably like to help others. We know you're the best suited for a Field Technical Coordinator, or Assistant Technical Coordinator position.

You've read this far. Now, take the next step and contact your Section Manager. He or she will welcome you as an official ARRL volunteer, and you will welcome your new opportunity to help others like Jeffrey Dube, while helping yourself. We're sure you'll find your new TC hat to be just your size!—*K1CE*

A 0.5-Watt 903-MHz Amplifier

By Zack Lau, KH6CP
ARRL Laboratory Engineer

The two-transistor amplifier shown in Fig 1 is an inexpensive way to amplify the 10 mW of RF output available from a monolithic microwave integrated circuit (MMIC) to 0.5 W on the 903-MHz band. Q1 and Q2, MRF 559s, are plastic-cased devices selling for less than $2 each. When driven with an Avantek MSA-0304 MMIC (around 10-mW saturated output), the saturated CW output is 0.5 W as measured with an HP435/8482A power meter and a Bird 10-dB attenuator. A two-tone IMD test indicates that the MMIC saturates before this amplifier does, since the higher order IMD products are down 47 dB. See Fig 2.

The amplifier is built on a 1/16-inch, thick, double-sided glass-epoxy PC board (Fig 3). All components mount on the etched side. The other side is left unetched to act as a ground plane. The 40-ohm striplines (Z1 through Z4) are made using 0.15-inch-wide traces.

Use good VHF/UHF construction techniques and keep the leads short when mounting components. Drill oversize holes for Q1 and Q2 so the leads lie flush against the board traces. Use copper foil wrapped through the transistor mounting holes to connect the top and bottom ground planes at the emitters of Q1 and Q2. Wrap the board edges with copper foil to connect the top and bottom ground planes.

Use R3 and R7 to adjust bias currents to 22 mA for Q1 and 87 mA for Q2. Note: These are the *total* currents as indicated by the voltage drops across R4 and R8. The bias currents do drift slowly without RF drive, but this is not a problem as long as the supply voltage is switched off during receive. The prototype works fine without a bypass capacitor at the junction of R6 and R7.

Tune-up is simple. Apply drive and adjust C2 and C6 for maximum output. Attempts at matching the output with a tunable network failed to improve gain or output power. Better results may be obtainable by using Teflon® -dielectric board and porcelain chip capacitors.

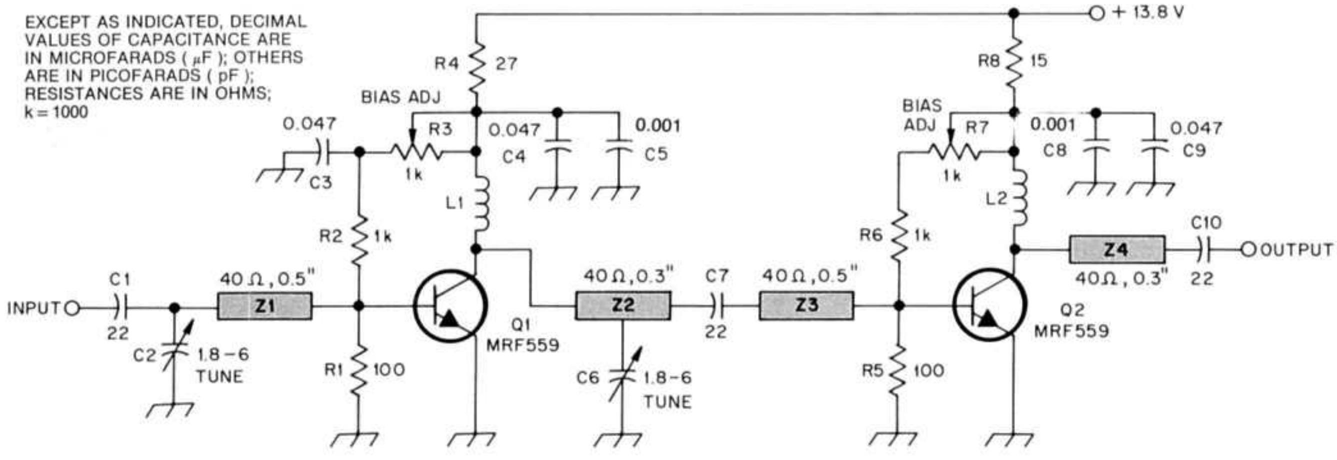


Fig 1—Schematic of an inexpensive 0.5-W 903-MHz amplifier. Resistors are ¼ W. Capacitors are NP0 chip capacitors unless noted, although more expensive porcelain capacitors may work better.

C2, C6—1.8-6-pF trimmer capacitors. Mouser 24AA070.
C3, C4, C9—tiny ceramic capacitors or chip capacitors.
L1—7 turns no. 22 tinned copper wire, space wound; no. 33 drill bit used as temporary form.
L2—5 turns no. 22 tinned copper wire, space wound; no. 33 drill bit used as temporary form.
Q1, Q2—MRF 559.

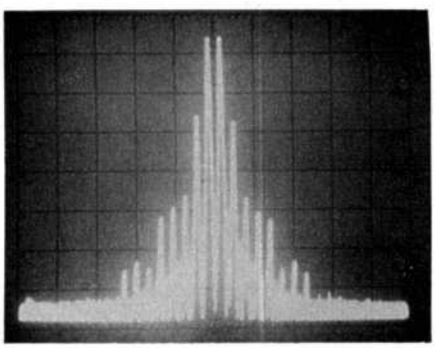
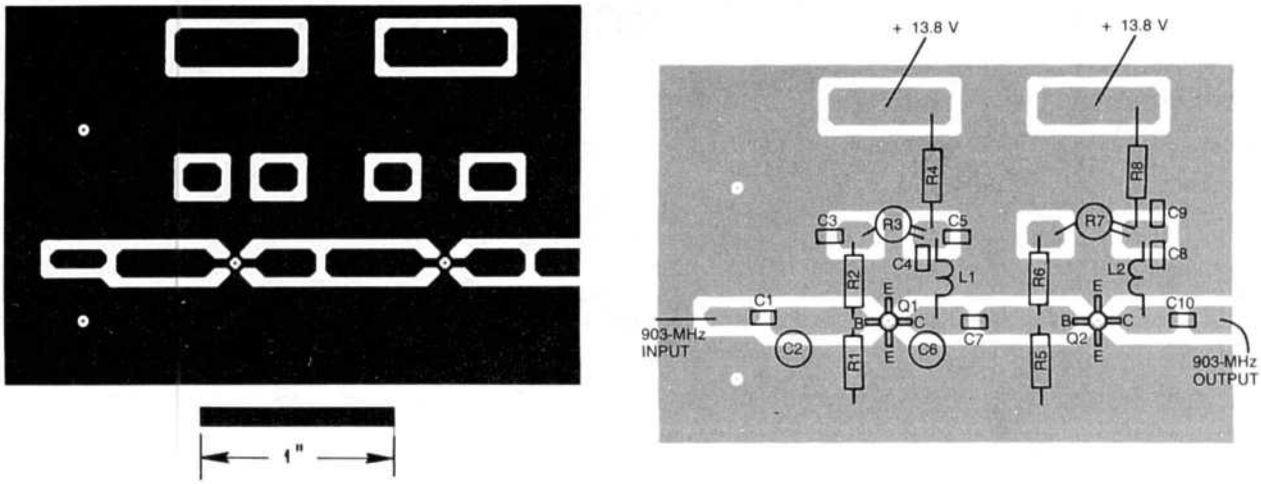


Fig 2—Spectral display of the 0.5-W 903-MHz amplifier during two-tone intermodulation distortion (IMD) testing. Third-order products are approximately 26 dB below PEP output, and fifth-order products are approximately 47 dB down. Vertical divisions are each 10 dB; horizontal divisions are each 10 kHz. The amplifier was being operated at 560-mW PEP output on 903.1 MHz.



(A) (B)

Fig 3—Circuit-board etching pattern (A) and parts-placement diagram (B) for the 0.5-W 903-MHz amplifier. The etching pattern is shown full size from the etched side of the board. Black areas represent unetched copper foil. Board material is 1/16-in. G-10, double sided. The other side of the board is unetched to form a ground plane. The shaded area in B represents the copper pattern. All components are mounted on the etched side of the board.

Bits

Satellite-Tracking Program for Commodore Computers

Many Amateur Radio operators use C64 of C128 computers as part of their satellite communications systems or for receiving TV and weather-satellite signals. In each of these cases, knowing when the satellite is accessible and where to aim their antennas are essential to successful operation.

SATCOMM-64 provides the C64/C128 owner with several useful features: It has a master menu to allows quick selection of any of 12 options, can store information on up to 15 different satellites and quickly confirm W1AW reference orbits. The program will provide a printed report of up to 31 days of access times (for a specific time bracket) for any satellite. Also available is a printed report of a single day's access times for up to three different satellites.

For each user-specified time interval, SATCOMM-64's printed reports include: relative azimuth and elevation, altitude, longitude and latitude, local time, UTC day, geographic areas that are within the satellite's communication range, Doppler shift, minimum and maximum communication distance, operating frequencies, orbit number and phase.

At this writing, SATCOMM-64 can print out detailed reports covering OSCARS 9, 10, 11 and 12, RS-5, 7, 10 and 11, MIR, Kivant, Salyut-7 and weather/research satellites (GOES/WEFAX, NOAA, Meteor). Whenever desired, you can replace any of these with new satellite choices.

Additional program features include: choice of screen and printed report or screen alone; easily altered defaults (start time, time increment and so on) and automatic changeover from Standard to Daylight time and vice versa (this feature may be bypassed). SATCOMM-64 also handles the annual rollover period during which the previous year's Keplerian elements have not yet been updated.

SATCOMM-64 requires a C64/128 computer, 1541 disk drive and 1525-compatible printer. Price: $15.95 (Missouri residents add sales tax), plust $3 for postage and handling. To order or obtain additional information, contact Strategic Marketing Resources, Inc, PO Box 2183, Ellisville, MO 63011, tel 314-256-7814.—*Paul K. Pagel, N1FB*

Macket Software for Macintosh Computers

Macket provides power and flexibility for the packet-radio operator who uses a Mac®. There are windows for entering text, displaying the receive buffer and logging transmitted text. These windows support all of the features expected by Mac users such as scrolling and text selection. The input window also allows mouse-based editing. Other features include text uploading and downloading, printing and the use of macros.

Macket works with all Pac-Comm TNCs: the TNC-200, -220, Tiny-2 and Micropower-2 as well as any other TNC that is equipped with an RS-232-C port. When used with a TNC-2 clone that supports the RXBLOCK command, Macket can display the user's conversations in a special window so that the conversation will not be mixed with monitored text.

The program, developed by S Fine Software, is available from Pac-Comm Packet Radio Systems, Inc, 3652 W Cypress St, Tampa, FL 33607, tel 800-223-3511 (orders only, except in Florida); technical information, 813-872-2980.—*Paul K. Pagel, N1FB*

Solid-State Linear Amplifiers for 33 cm

By David Hallidy, KD5RO
2118 W Jefferson Rd
Pittsford, NY 14534

Since the 33-cm band (902 to 928 MHz) became available for amateur use several years ago, a number of articles have described construction of 33-cm transmitters, receiving converters and transverters. Most of these projects, however, feature low-power (1 W or less) linear transmitting amplifiers, or else they use class-C amplifier modules to generate higher RF-output levels (typically 5 to 10 W). These approaches generally preclude the use of SSB because the effective communications range of low-power SSB is minimal at UHF, and the nonlinearity of a class-C amplifier causes objectionable distortion of an SSB signal.

I tried low-power and class-C transmitters, and I wanted to find something better—an amplifier capable of delivering 10 to 20 W of linear 33-cm power. My goal was an amplifier that could be used for portable "grid-peditions," and it had to provide an output level sufficient to drive a 2C39 cavity amp to a couple hundred watts output for long-haul tropo or EME work.

This article describes construction of a pair of solid-state 33-cm power amplifiers that meet my requirements:

- They are linear and can be used with any emission mode.
- They are compact and run from a 13.8-V dc supply—ideal for portable use.
- When the modules are cascaded and driven with about 1.5 W, the output level is about 18 W—more than enough to drive a 2C39 amplifier to full output.
- They are stable.
- They are easy to duplicate.

My amplifiers are based on a design presented several years ago in *QST* by Al Ward, WB5LUA.[1] (These amplifiers are also described in Chapter 32 of *The ARRL Handbook* from 1986 to the present edition.) Al described the use of the NEL1306 and NEL1320 transistors at 1296 MHz. Manufactured by NEC, these NPN bipolar transistors are specified for operation from 1.0 to 1.5 GHz. The '1306 is rated at 6.5 dB gain and 7 W output,

[1]Notes appear on page 7.

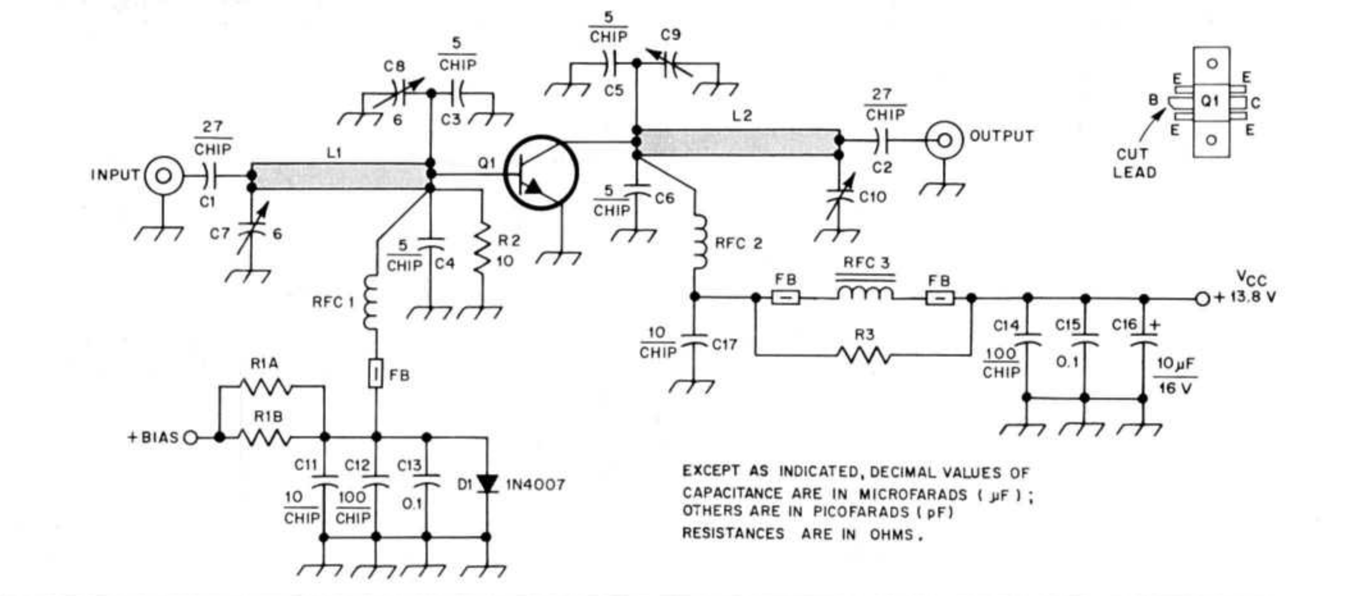


Fig 1—Schematic diagram of the 33-cm solid-state power amplifiers. The schematic is identical for both versions. Component values are the same except as noted.

C1, C2—27-pF chip capacitor.
C11, C17—10-pF chip capacitor.
C3, C4, C5, C6—3.6- to 5.0-pF chip capacitor.
C7, C8—1.8- to 6.0-pF miniature trimmer capacitor (Mouser 24AA070 or equiv).
C9, C10—Same as C7 and C8 for the NEL1306 amplifier. For the NEL1320 version, 0.8- to 10-pF piston trimmers are used (Johanson 5200 series or equiv.).
C12, C14—100-pF chip capacitor.
C13, C15—0.1-µF disc ceramic capacitor.
C16—10-µF electrolytic capacitor.
D1—1N4007 diode.
L1, L2—30-ohm microstrip line, ¼-wavelength long (see text).
Q1—NEC NEL130681-12 (6 W) or NEL132081-12 (18 W) transistor.
R1—82- to 100-Ω resistor, 2-W minimum. Vary for specified idling current.
R2—10-Ω, ¼-W carbon-composition resistor with "zero" lead length. See text.
R3—15-Ω, 1-W carbon-composition resistor.
RFC1—4 turns no. 24 wire, 0.125 inch ID, spaced 1 wire diam.
RFC2—2 turns no. 24 wire, 0.125 inch ID, spaced 1 wire diam.
RFC3—1-µH RF choke; 18 turns no. 24 enam. close-spaced on a T-50-10 toroid core.

and the '1320 is rated at 6.0 dB gain and 18 W output. (Power output ratings are at the 1-dB compression point.) Current prices make these devices reasonable for amateur work; as of early 1988, the '1306 is $32.50 and the '1320 is $54.00 in single quantities.[2]

Al's circuit, as originally presented, was very straightforward. He used microstriplines each one-quarter wavelength long to match the input and output of the amplifiers. He also used a simple bias network and clean construction techniques. It will be helpful for you to refer to Al's original article in *QST* or *The ARRL Handbook* for background information regarding construction of these amplifiers. This article concentrates on my efforts to modify Al's original design for operation at 33 cm.

Construction

Most of the components shown in the original article are retained in the 33-cm design. See Fig 1. All components mount on a pair of 0.031-inch, double-sided, glass-epoxy circuit boards.[3] No hard-to-find parts are used. If you don't have suitable chip capacitors and trimmer capacitors on hand, you should be able to find everything you need at Microwave Components of Michigan.[4]

I decided that it would be a simple matter to redesign the input and output circuits to tune to 902 MHz. The most important change is lengthening the 30-Ω, quarter-wave lines used in the base and collector circuits. These lines (L1 and L2 of Fig 1) are etched on the PC board.

As originally laid out, these lines are each 1.125 inches long, representing a quarter wavelength in microstrip at 1296 MHz. Calculations show that the new lines must be 1.63 inches long to resonate at 902 MHz.[5] To preserve the general layout of the original circuit, I extended the lines toward the input- and output-connector ends of the boards. To maintain the overall symmetry, I also increased the size of the ground areas extending to the board edges. This resulted in an overall increase of about ½ inch in the length of each board. Fig 2 shows a full-size etching pattern and parts-placement diagram.

A couple of component changes must be noted. In the original article, RFC1 and RFC2 were selected for "the lowest possible reactance that will not affect power gain or output power." At 902 MHz, these values are a little small. The amplifiers work well enough without changing these chokes, but increasing the size of each by one turn resulted in about 0.5 dB more gain at the new frequency. In addition, coupling capacitors C1 and C2 are increased to 27 pF from the 10-pF units used at 1296 MHz.

Al's 1296-MHz design included a 10-Ω,

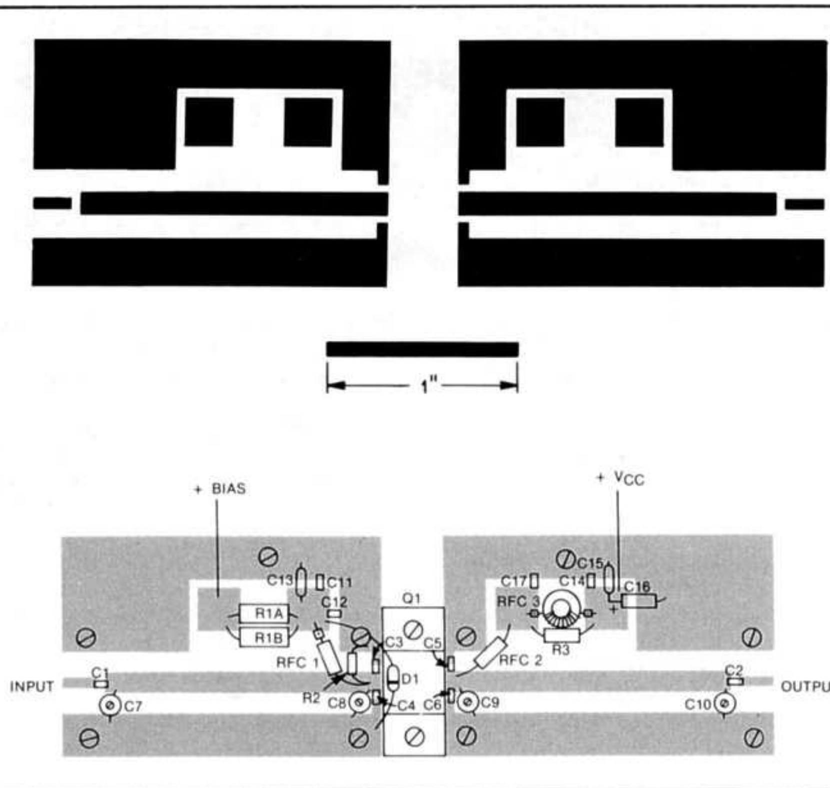


Fig 2—Etching pattern (A) and parts-placement diagram for the solid-state 33-cm power amplifiers. PC board material is double-sided 0.031-inch glass-epoxy. One side is left unetched to act as a ground plane. Wrap the board edges with conductive foil and solder it on both sides of the board to make a good connection between the top and bottom ground planes. All components mount on the etched side of the board. The same PC boards are used for each version.

¼-W resistor (R2) in the base circuit to ensure stability. I found I could remove this resistor without causing any noticeable instability, and removing this resistor further increased the gain. I suggest first trying the circuit with the 10-Ω resistor in place. If gain and power output are sufficient, leave R2 in. If you need a little more gain, remove R2 or increase its value. Remember that if you do this, you will need to check the idling current and readjust the value of R1 to bring it back to the specified level (50 mA for the '1306 and 150 mA for the '1320).

I strongly recommend following the construction tips given in the original article—particularly those pertaining to grounding of the boards and minimizing lead lengths. I found that the output capacitor, C10, was not required in my version of the '1306 amplifier, but it was necessary in the '1320. C10 is shown on the schematic and parts placement diagram for both units, however. There is enough variation among devices and substrate characteristics that C10 may be necessary to provide optimal matching to the load.

The body of D1, in the bias network, should be placed against the case of the power transistor to ensure best thermal stability. I recommend the use of a small amount of thermal compound here to reduce the thermal resistance of this joint and maximize heat transfer to the diode. Also be sure to use thermal compound underneath the transistor before bolting it to the heat sink. Fig 3 shows the finished amplifiers. The heat sinks may look large, but the devices do dissipate quite a bit of heat, so I felt that a lot of heat sink would only aid reliability.

Tune-Up

You are again urged to refer to the original article for details of amplifier tune-up. Initially set all trimmer capacitors to minimum capacitance. You'll need a 13.8-V power supply, a 902-MHz signal source (1.5 W maximum), an appropriate wattmeter and a dummy load. It may be helpful to insert an ammeter capable of reading about twice the expected operating current of the device in the collector supply lead. This allows you to keep an eye on idling current, power dissipation and efficiency, and it also helps you to see if there is any "funny stuff" (for example, oscillations) happening.

Apply drive gradually, peaking the input and output circuits as the level is increased. With the '1306, start with

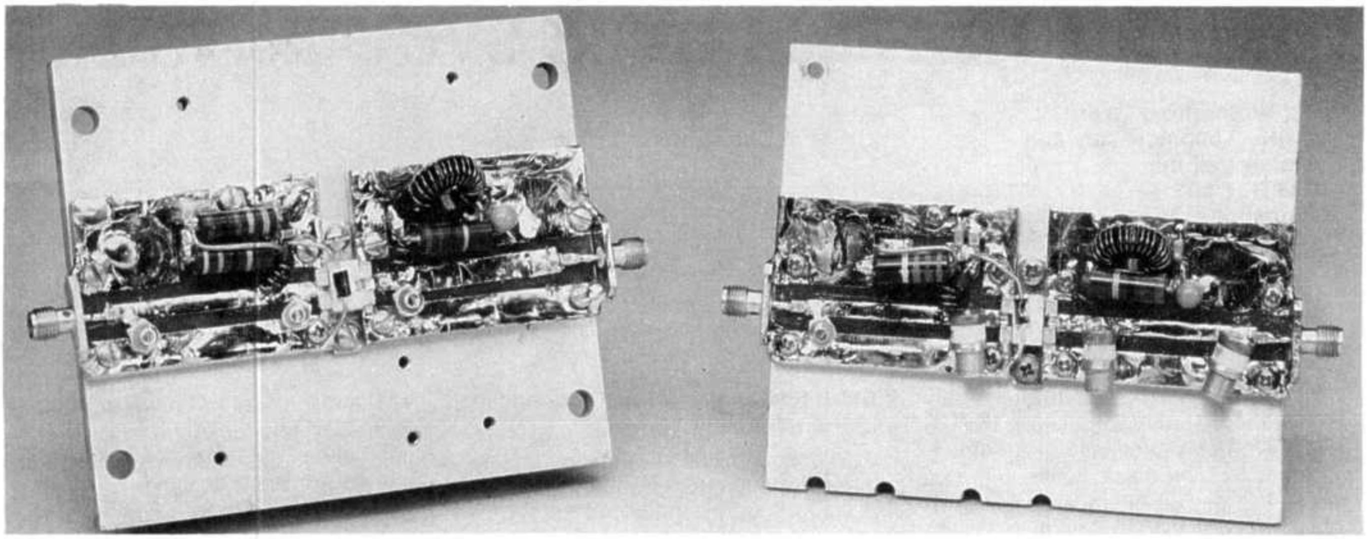


Fig 3—The 6-W (left) and 18-W (right) 33-cm amplifiers are virtually identical.

about 100 mW and end up at the 1- to 1.5-W level. With the '1320, start with about 1 W and finish with about 6 W. Table 1 lists the measured output level, gain and collector current for each device at various drive levels.

Summary

These amplifiers are a solution to the problem of developing moderate linear RF power for the 33-cm amateur band. The modules are easy to build and make operational. On-the-air reports have been gratifying—my signal has been reported to be comparable in quality to a popular commercially made 33-cm transverter. I now feel confident that I can provide enough drive to a 2C39 amplifier to maximize its potential.

Table 1
Amplifier Performance Characteristics

Drive	*Drive (W)*	*Gain (dB)*	I_c *(A)*	*Power Output (W)*
NEL130681-12	0.1	7.5	0.28	0.6
	0.2	7.0	0.40	1.0
	0.5	7.0	0.60	2.5
	1.0	7.0	0.88	5.0
	1.5	6.5	1.20	6.8
NEL132082-12	0.6	5.0	1.0	1.8
	1.0	5.0	1.3	3.1
	2.5	5.0	2.1	7.9
	5.0	4.5	2.8	14.1
	6.8	4.2	3.0	17.8

Notes

[1]Al Ward, "1296-MHz Solid-State Power Amplifiers," *QST*, Dec 1985, pp 41-44.

[2]NEC transistors are available from California Eastern Laboratories, 3260 Jay St, Santa Clara, CA 95050. Contact CEL for ordering information or for the address of the closest CEL sales office.

[3]If you have trouble locating 0.031-inch PC board material, the author can provide suitable unetched pieces for a nominal charge.

[4]Microwave Components of Michigan, 11216 Cape Cod, Taylor, MI 48180, tel 313-941-8469 (eves).

[5]Calculating the length of a quarter-wavelength line in microstrip is fairly simple. In this case, a very close (and usable) approximation may be made by scaling the line length to the ratio of the two frequencies. That is, since 1296/902 equals a ratio of 1.44, the 902-MHz line will have to be 1.44 times as long as the 1296-MHz line. Therefore, 1.125 inch × 1.44 = 1.62 inches. The correct method, however, is to determine the shortening factor for the line and multiply this factor by the length of a quarter wave in air. Determining the shortening factor is fairly involved. You must first determine the *effective* dielectric constant for the substrate material and line width being used. The shortening factor (SF) is then given by

$$SF = \frac{1}{\sqrt{\epsilon_{eff}}} \quad \text{(Eq 1)}$$

where ϵ_{eff} is the effective dielectric constant. For 0.031-inch G-10 material with a line width of 0.121 inch (30-Ω line), this comes out to be about 0.5. This factor is then multiplied by the length of a quarter wavelength in air at the design frequency to determine the actual length. Thus,

$$L = \frac{2952\ SF}{F} \quad \text{(Eq 2)}$$

where
L = line length in inches
SF = shortening factor (from Eq 1)
F = frequency in MHz

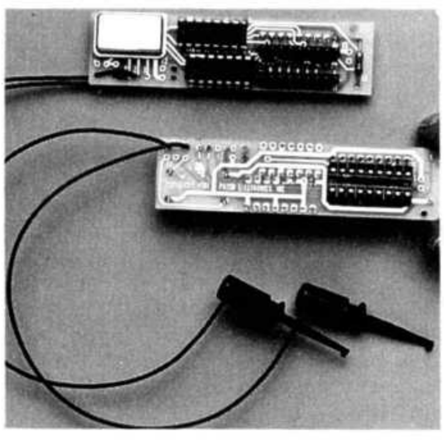
In this case, L = (2952 × 0.5)/902 = 1.64 inches. This result is very close to the length determined by the scaling method, and the scaling method is simpler.

Bits

The PC-BANDIT

PRISM Electronics has developed a low-cost accelerator board for IBM® PC, PC-XT, and PC compatibles. PC-BANDIT™ resides in the computer's current 8284 clock chip position to provide additional clocks with its own clock IC. No expansion slot is required.

The 0.8-in × 3.1-in board uses software to toggle between speeds, and it is compatible with certain BIOS hot-key sequences. The board does not interfere with the computer's real-time clock, which continues to operate at its correct rate. Two versions of the PC-BANDIT are available: the high-speed 7.4 MHz for 150-ns memory chips, and the 6.7 MHz for machines using 200-ns chips. Price: $69.95. Contact PRISM Electronics, Inc, 14682 NE 95th St, Redmond, WA 98052, tel 206-881-1088.



The PC-BANDIT. This small board plugs into your IBM PC (or compatible) computer to boost processing speed by as much as 60%.

Designing LC Filters Using SVC Filter Tables

By Ed Wetherhold, W3NQN
ARRL Technical Advisor
Honeywell Inc
PO Box 391
Annapolis, MD 21404

Filter-design tables that employ standard-value capacitors (SVCs) have been published in a number of places.[1-8] Now these tables can also be used to design filters with standard-value *inductors* (SVLs). The filter-design procedure described in this article uses both standard-value capacitors and inductors, and is helpful in designing passive LC audio filters. It is possible to use a standard-value LC filter design for virtually any cutoff frequency because the impedance level of the equally-terminated audio filter can be varied until a unique condition is achieved where both the capacitor and inductor values are standard. The penalty for this convenience is that the cutoff frequency may not be exactly that desired, but it will be close enough to satisfy amateur filtering requirements. Don't be concerned with the impedance level; I will explain later why this is not a problem.

RF filters generally use a standard impedance level of 50 Ω. Equally terminated audio filters may be designed for any impedance level. For example, assume an LC filter is placed between two op amps where any value of filter termination resistance can be specified. That resistance value can be placed in series with the output of the driver op amp, and in parallel with the input of the following op amp, to properly terminate the filter input and output. This is possible because the op amp output and input impedances are less than ten ohms and more than 100-kΩ, respectively. The filter impedance that uses SVCs and SVLs for a given cutoff frequency can be calculated using a special design procedure.

An example of how to use a passive LC filter in combination with two op amps was discussed by Gary Breed, K9AY, in January 1988 *QST*.[9] Breed used a different approach than what I describe here, however. By following the procedures explained in this article, you can use the SVC design tables to obtain optimum Chebyshev or elliptic audio filter designs in which both the capacitor and inductor values are standard.

[1]Notes appear on page 10.

Filter Impedance, Cutoff Frequency and Inductance

For this application, the acceptable impedance and cutoff frequency ranges are about 500Ω to 10 kΩ and 500 Hz to 20 kHz, respectively. Inductance values of 22 and 88 mH are readily available[10-12] because these are the standard values of surplus telephone-line loading coils. Each of these high-Q molybdenum permalloy toroidal inductors has two windings. If the windings are connected in "series aiding," the resulting inductance is 88 mH. If the windings are connected in parallel, the inductance is 22 mH. Values less than 88 mH can be obtained by removing turns. The ability to get non-standard values of inductance is important when designing a standard-value LC, 5-branch elliptic filter, or a 7-branch Chebyshev filter. In such cases, one of the inductor values can be 22 or 88 mH, and the other value will be smaller. You can get a smaller value by removing the proper number of turns from a standard inductor.

For optimum convenience in filter construction using these surplus inductors, let's consider an inductance value of 88 mH to be standard. This value is used in the following design example to demonstrate how the low-pass elliptic SVC filter table can be used to design a standard-value LC 2.8-kHz low-pass filter. The same procedure is applicable to the other SVC tables.

General Design Procedure

1) Select a low-pass or high-pass filter, a filter family (Chebyshev or elliptic), the desired ripple cutoff frequency (F_{co}) and standard inductance value (L_s).

2) Find the published SVC filter table corresponding to the filter type and family selected.

3) Scale the tabulated data from MHz to kHz, pF to nF and μH to mH, by replacing the heading units with kHz, nF and mH. Next, scale the impedance level from 50 Ω to 500 Ω by dividing the tabulated capacitor values by ten and multiplying the inductor values by ten. The other data such as frequency (kHz), SWR and attenuation remain unchanged.

4) Using the original tabulated frequencies (in kHz) and the 500-Ω scaled inductance values, calculate the SVL scaling factor, S, for each design

$$S = F_{cot} \times (L_t) \qquad \text{(Eq 1)}$$

where F_{cot} is the tabulated filter cutoff frequency in kHz, and (L_t) is the 500-Ω scaled tabulated L value in mH. For future reference, write the calculated S value at the end of the row of data for each tabulated SVC.

5) Calculate the S factor for your desired F_{co} and the standard inductance value (88 mH) to be used in the filter construction.

6) Refer to the SVC table (Table 1). Look through the S values you documented and pick the SVC design that has an S value closest to the S value of your desired design. The 500-Ω capacitor values are used directly in the new standard-value LC design. The inductance and frequency values are scaled.

7) Calculate the actual cutoff frequency (F_{coa}) based on the tabulated cutoff frequency (F_{cot}), the tabulated 500-Ω L-value (L_t) and the standard L value (L_s) using

$$F_{coa} = F_{cot} \times (L_t/L_s) \qquad \text{(Eq 2)}$$

Calculate the other frequencies ($F_{3\,dB}$, and so on) in a similar manner.

8) Calculate the actual impedance level, Z_a, using

$$Z_a = 500 \times \sqrt{(L_s/L_t)} \qquad \text{(Eq 3)}$$

9) If a second and smaller inductance value is required, the scaled smaller value is calculated in direct proportion to the larger value.

a) If the second inductance value is less than 88 mH, but greater than 22 mH, calculate how many turns must be removed from the inductor using

$$T_s = 380.7 - 40.58 \times (L_s) \qquad \text{(Eq 4)}$$

where L_s is the desired inductance (mH) with the windings in the *series*-aiding connection, T_s is the turns to be removed from each of the two windings, and the total turns removed is 2 × T_s.

b) If the second inductance value is less than 22 mH, calculate the turns to be removed using

$$T_p = 380.7 - 81.15 \times (L_p) \qquad \text{(Eq 5)}$$

where L_p is the desired inductance (mH) with the *parallel*-aiding connection, T_p is the turns removed from each of the two windings, and the total turns removed is $2 \times T_p$.

These equations are based on the data obtained from Table 1 of my September 1968 *QST* article[13] and are good for the core type used in generating the data in that article. For different cores, other turn-removal equations are calculated using the procedure described in Appendix A of my February 1984 *ham radio* article.[14]

10) Assemble the components in accordance with the schematic diagram associated with the SVC table that you use.

This completes the general description of the standard-value LC filter-design procedure. The following example uses actual values and more clearly demonstrates the simplicity of the procedure.

Example of the Standard-Value LC Filter Design Procedure

Let's say that we have chosen to work with a fifth-degree elliptic low-pass filter. The desired F_{co} is 2.8 kHz and the standard inductance value is 88 mH.

Table 15 on page 2-47 of *The 1988 ARRL Handbook* shows the SVC low-pass elliptic table. The data in Table 15 is frequency and impedance scaled to the 1- to 10-kHz decade and to an impedance level of 580 Ω. To get an understanding of this procedure, see Table 1 in this article for a scaled version of Table 15. Note that only the tabulated numerical values of capacitance and inductance have changed. The C and L values are now, respectively, one-tenth and ten times their original values. In addition, the S factors for each SVC design in Table 15 have been computer-calculated, and they are listed at the end of each row of data in Table 1 of this article.

The S factor for a filter having the desired cutoff frequency of 2.8 kHz and a standard-value inductance of 88 mH is equal to $2.8 \times \sqrt{(88)} = 26.27$. Look through the S-column values in Table 1 for the closest match to 26.27. Design no. 27 has S = 26.09, and is the closest value to that desired. Design no. 27 is therefore selected for scaling to get the values of the final design. The capacitor values of 56, 82, 47, 6.41 and 18.1 nF are copied directly as the capacitance values of the new filter.

The frequencies of the new filter are calculated from

$$F_{coa} = F_{cot} \times \sqrt{(L_t/L_s)} \qquad \text{(Eq 6)}$$

where F_{cot} = 7.18 kHz, L_t = 13.2 mH and L_s = 88 mH. Note that this filter type has *two* inductor values. We select the highest value (L2) for the following scaling calculations so that the smaller inductance (L4) can be obtained by removing turns from another standard 88-mH inductor.

$$F_{coa} = 7.18 \text{ kHz} \times \sqrt{(13.2/88)} = 7.18 \text{ kHz} \times 0.3873 = 2.781 \text{ kHz} \qquad \text{(Eq 7)}$$

$$F_{3\,dB} = 7.68 \text{ kHz} \times 0.3873 = 2.974 \text{ kHz} \qquad \text{(Eq 8)}$$

$$F_{As} = 11.1 \text{ kHz} \times 0.3873 = 4.299 \text{ kHz} \qquad \text{(Eq 9)}$$

$$F2 = 17.3 \text{ kHz} \times 0.3873 = 6.700 \text{ kHz} \qquad \text{(Eq 10)}$$

$$F4 = 11.5 \text{ kHz} \times 0.3873 = 4.454 \text{ kHz} \qquad \text{(Eq 11)}$$

Note that the desired F_{co} was 2.8 kHz, but the actual F_{co} of 2.781 kHz is acceptable in exchange for the convenience of using a standard-value LC design.

The actual impedance level, Z_a, is calculated by

$$Z_a = 500 \times \sqrt{(L_s/L_t)}\ \Omega$$

$$Z_a = 500 \times \sqrt{(88/13.2)} = 500 \times 2.58199 = 1291\ \Omega \qquad \text{(Eq 12)}$$

Standard 1.3-kΩ, five-percent tolerance resistors will be suitable for the filter source and load terminations.

L2 was selected to be 88 mH, and L4 will be proportionately smaller

$$L4 = 88 \times (10.6/13.2) \text{ mH} = 88 \times 0.80303 \text{ mH} = 70.7 \text{ mH} \qquad \text{(Eq 13)}$$

For 70.7 mH, the series-aiding connection is used, and

$$T_s = 380.7 - 40.58 \times (70.7) = 380.7 - 340.5 = 40 \qquad \text{(Eq 14)}$$

The results tell us that 40 turns must be removed from each winding—in other words, that *80* turns must be removed in total.

Fig 1A shows the schematic diagram of the 2.8 kHz standard-value LC elliptic low-pass filter. All parameter values are shown. Fig 1B shows the filter attenuation response.

Summary

This procedure demonstrates how to design equally-terminated LC audio filters having both standard-value capacitors and inductors. Expanding existing SVC filter tables and scaling them for use in

Table 1

5-Branch Elliptic 500-Ω Low-pass Filters, C-in/out with E12 Values for C1, C3 and C5.

No.	F_{co}	F_{3dB}	F_{As}	As	Max	C1	C3	C5	C2	C4	L2	L4	F2	F4	S
		(kHz)		(dB)	SWR			(nF)			(mH)		(kHz)		
1	0.795	0.989	1.57	47.4	1.092	270	560	220	32.4	93.7	121	101	2.54	1.64	8.73
2	1.06	1.20	1.77	46.2	1.234	270	470	220	34.1	98.2	93.6	75.6	2.82	1.85	10.28
3	1.47	1.57	2.15	45.4	1.586	270	390	220	36.4	104.5	63.2	48.8	3.32	2.23	11.70
4	0.929	1.18	1.91	48.0	1.077	220	470	180	25.7	74.3	102	85.9	3.11	1.99	9.37
5	1.27	1.45	2.17	46.7	1.215	220	390	180	27.1	77.9	78.5	63.9	3.45	2.26	11.25
6	1.69	1.82	2.54	45.9	1.489	220	330	180	28.7	82.1	56.4	44.2	3.96	2.64	12.71
7	1.12	1.44	2.41	49.8	1.071	180	390	150	19.2	54.9	84.5	72.5	3.95	2.52	10.25
8	1.49	1.73	2.70	48.8	1.183	180	330	150	20.0	57.0	67.5	56.2	4.33	2.81	12.20
9	2.11	2.27	3.27	47.8	1.506	180	270	150	21.3	60.4	45.5	36.4	5.12	3.40	14.24
10	1.28	1.66	2.63	46.3	1.064	150	330	120	19.2	56.1	72.0	60.0	4.28	2.74	10.89
11	1.79	2.06	2.99	44.8	1.195	150	270	120	20.4	59.2	55.2	44.2	4.75	3.11	13.29
12	2.52	2.70	3.63	43.8	1.525	150	220	120	22.0	63.6	37.1	28.2	5.58	3.76	15.35
13	1.56	2.08	3.55	50.1	1.055	120	270	100	12.7	36.3	58.8	50.7	5.83	3.71	11.97
14	2.23	2.59	4.04	48.8	1.183	120	220	100	13.3	38.0	45.0	37.5	6.50	4.22	14.95
15	3.17	3.41	4.90	47.8	1.506	120	180	100	14.2	40.2	30.3	24.2	7.68	5.10	17.44
16	1.94	2.52	4.15	48.4	1.064	100	220	82	11.5	33.1	47.9	40.6	6.78	4.34	13.43
17	2.73	3.14	4.73	47.0	1.199	100	180	82	12.1	34.8	36.6	29.9	7.56	4.93	16.49
18	3.73	4.02	5.63	46.2	1.491	100	150	82	12.9	36.8	25.6	20.1	8.76	5.85	18.88
19	2.39	3.11	5.20	49.4	1.065	82	180	68	8.93	25.6	39.1	33.5	8.51	5.44	14.93
20	3.26	3.79	5.85	48.2	1.185	82	150	68	9.36	26.7	30.7	25.4	9.39	6.10	18.07
21	4.83	5.17	7.30	47.2	1.569	82	120	68	10.0	28.6	19.5	15.4	11.4	7.58	21.36
22	2.85	3.71	6.15	48.8	1.063	68	150	56	7.66	22.0	32.6	27.8	10.1	6.43	16.26
23	4.16	4.74	7.14	47.3	1.221	68	120	56	8.13	23.3	24.0	19.7	11.4	7.44	20.39
24	5.72	6.13	8.58	46.5	1.547	68	100	56	8.63	24.6	16.5	13.0	13.3	8.91	23.26
25	3.67	4.69	7.95	50.5	1.076	56	120	47	5.76	16.4	25.9	22.3	13.0	8.31	18.68
26	5.02	5.77	9.01	49.4	1.212	56	100	47	6.03	17.1	20.1	16.8	14.5	9.40	22.48
27	7.18	7.68	11.1	48.6	1.582	56	82	47	6.41	18.1	13.2	10.6	17.3	11.5	26.09
28	4.40	5.60	9.24	49.3	1.079	47	100	39	5.14	14.7	21.6	18.4	15.1	9.66	20.48
29	6.17	7.01	10.6	48.0	1.236	47	82	39	5.42	15.5	16.3	13.4	17.0	11.1	24.86
30	8.63	9.20	12.9	47.3	1.604	47	68	39	5.76	16.4	10.9	8.57	20.1	13.4	28.43
31	5.47	6.91	11.8	51.3	1.086	39	82	33	3.85	10.9	17.6	15.2	19.3	12.3	22.96
32	7.55	8.59	13.5	50.2	1.242	39	68	33	4.04	11.4	13.4	11.2	21.7	14.1	27.63
33	10.9	11.5	16.8	49.5	1.659	39	56	33	4.28	12.0	8.62	6.95	26.2	17.4	31.87
34	6.59	8.17	13.0	47.7	1.096	33	68	27	3.90	11.2	14.6	12.2	21.1	13.6	25.20
35	9.10	10.2	15.0	46.5	1.267	33	56	27	4.12	11.8	10.9	8.81	23.7	15.6	30.08
36	12.4	13.2	18.1	45.8	1.635	33	47	27	4.39	12.5	7.41	5.73	27.9	18.8	33.87

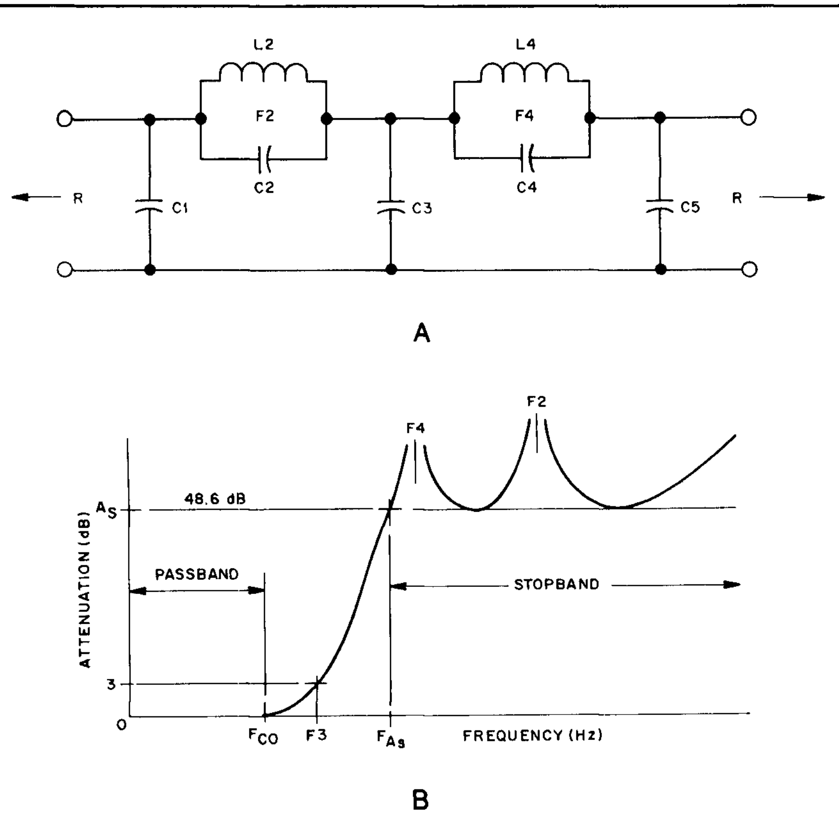


Fig 1—The standard-value LC elliptic, low-pass filter design for 2.8-kHz ripple cutoff frequency. The schematic diagram and associated parameter values are shown at A. At B, attenuation is shown as a function of frequency.

F_{co} = 2.781 kHz	A_s = 48.6 dB	C1 = 56 nF (0.056 μF)	L2 = 88 mH
$F_{3\,dB}$ = 2.974 kHz	Max SWR = 1.582	C2 = 6.41 nF (6410 pF)	L4 = 70.7 mH
F_{As} = 4.299 kHz	F_{As}/F_{co} = 1.546	C3 = 82 nF (0.082 μF)	
F2 = 6.700 kHz		C4 = 18.1 nF (0.0181 μF)	
F4 = 4.454 kHz	R = 1291 Ω	C5 = 47 nF (0.047 μF)	

designing standard-value LC filters was described.

Virtually any cutoff frequency from about 500 Hz to 20 kHz can be selected for standard L and C values if the impedance level is kept constant (by placing the filter between two op amps). The scaling steps were demonstrated for designing a 2.8-kHz elliptic, low-pass standard-value LC filter using an expanded SVC table that included the standard-value L scaling factors. A standard 88-mH inductor and a modified 88-mH inductor were used as the inductive elements. The given equations allow calculation of how many turns must be removed in modifying the inductors. A schematic diagram with all values and a response curve complete the description of the standard-value LC filter design. I recommend this procedure be used to design audio LC filters when used in combination with op amps.

Notes

[1]M. Wilson, ed., *The 1988 ARRL Handbook* (Newington: ARRL, 1987), pp 2-41 to 2-50.

[2]W. I. Orr, ed., *Radio Handbook* (Indianapolis: Howard W. Sams, 1987), pp 3-17 to 3-29.

[3]"Simplified passive LC filter design for the EMC engineer," from the record of the 1985 IEEE International Symposium on Electromagnetic Compatibility, Aug 1985, pp 575-584, IEEE Catalog No. 85CH2116-2.

[4]R. Graf, *Electronic Databook* (Summit: Tab Book Co, 1983), pp 117-143.

[5]E. Wetherhold, "Passive Elliptic Filters using (E24) Standard-Value Capacitors," *Interference Technology Engineers' Master* (ITEM), 1983, published annually by R&B Enterprises, West Conshohocken, PA.

[6]"Low-Pass and High-Pass Filters use Standard-Value Capacitors," *Electronic Designer's Casebook No. 5* (McGraw-Hill, 1982), pp 94-97.

[7]NAVAIR AD 1115, *Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment*, chapter 8, Filtering, Change 1 (7/80), Addendum no. 2, pp 8-58 to 8-67, US Naval Air Systems Command, Washington, DC.

[8]E. Pasahow, *Electronics Ready Reference Manual* (McGraw-Hill, 1985), pp 109 to 124, section 5-3.

[9]G. Breed, "A New Breed of Receiver," *QST*, Jan 1988, pp 16-23.

[10]H. Mitchell, "88-mH Inductors—A Trap!," *QST*, Jan 1983, pp 38-39.

[11]Amidon Associates, 12033 Otsego St, North Hollywood, CA 91607. Send for the *Iron-Powder and Ferrite Coil Forms* catalog.

[12]Typetronics, Box 8873, Ft Lauderdale, FL 33310, tel 305-583-1340.

[13]E. Wetherhold, "Inductance and Q of Modified Surplus Toroidal Inductors," *QST*, Sep 1968, p 37, Table 1.

[14]E. Wetherhold, "Elliptic Lowpass Audio Filter Design," *ham radio*, Feb 1984, p 20. Copies of this article are available from the author for a 4 × 9 ½-inch SASE with two units of first-class postage.

Bits

Multilayer Universal Prototyping Board

The MD12BRD four-layer universal prototyping board is designed to provide engineers with an easy way to evaluate NEC high-speed GaAs logic devices. The board has nine sites for 16-pin flat-pack GaAs ICs, three sites for 20-pin flat-pack GaAs Ics and allows easy access to power-supply and RF interconnections.

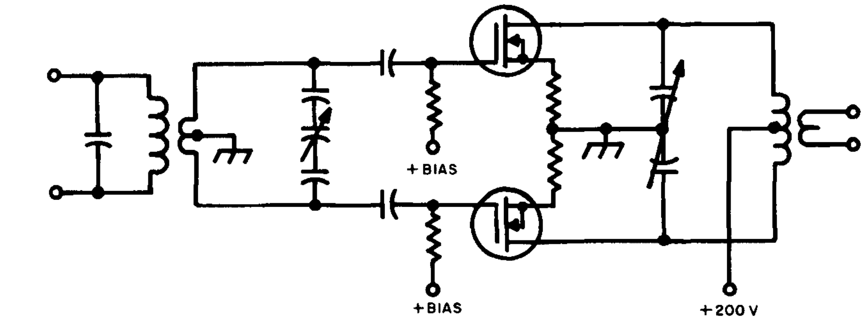
The board has been designed for device applications with clock inputs up to 5 GHz and 50-ohm impedance system operation. It comes with application and assembly information, as well as a list of recommended capacitors, termination resistors, heatsink and RF interconnections.

By employing a variety of NEC GaAs ICs, the MD12BRD can be used to build systems that run at speeds up to 4 GHz. Typical experiments that can be performed with the board are input-clock-signal sensitivity, maximum operating frequency, propagation-delay measurements, rise- and fall-time measurements, output-termination techniques and system prototyping.

The boards cost $100 each in any quantity ordered. For further information, contact Don Apte, Product Marketing Manager, California Eastern Laboratories, Inc, 3260 Jay St, Santa Clara, CA 95054, tel 408-988-3500; FAX 408-988-0279.—*Paul K. Pagel, N1FB*

Notes on a Lightweight, Portable CW Transmitter with a Transformerless Power Supply

By Robert W Vreeland, W6YBT
45 Maywood Dr
San Francisco, CA 94127



If the power output of a solid-state portable station is to be more than a few watts, the problem of power-supply weight must be overcome. This article presents an overview of a portable, solid-state transmitter capable of 15 to 30 W output (depending on band) on 20, 40 and 80 m.[1] Total transmitter weight is 4 lbs: 2 lbs for the driver/final, 1 lb for the exciter and 1 lb for its *transformerless* power supply.

[1]This transmitter design was originally described in R. W. Vreeland, "RF Operation of 450-V Vertical Power MOS Transistors in an Ultralight HF Transmitter," *Proceedings, RF Technology Expo 85* (Cardiff Publishing: Anaheim, 1985), pp 261-265. The paper was reprinted as "An Ultra-Lightweight HF Transmitter Using High Voltage MOSFETs," *rf design,* Aug 1985, pp 46-50.

Driver/Final

See Fig 1. The final amplifier consists of a pair of Supertex VN0360N1 600-V power MOSFETs in a push-pull configuration. These transistors were selected after a long search for a MOSFET with a fast switching time, low input capacitance and a high enough drain-source breakdown voltage to withstand reasonable antenna mismatches. Because the final amplifier transistors operate in push-pull, only modest output filtering (a single, balanced tank circuit) is needed to ensure that the transmitter meets current FCC purity-of-emissions regulations. (On 40 m, for

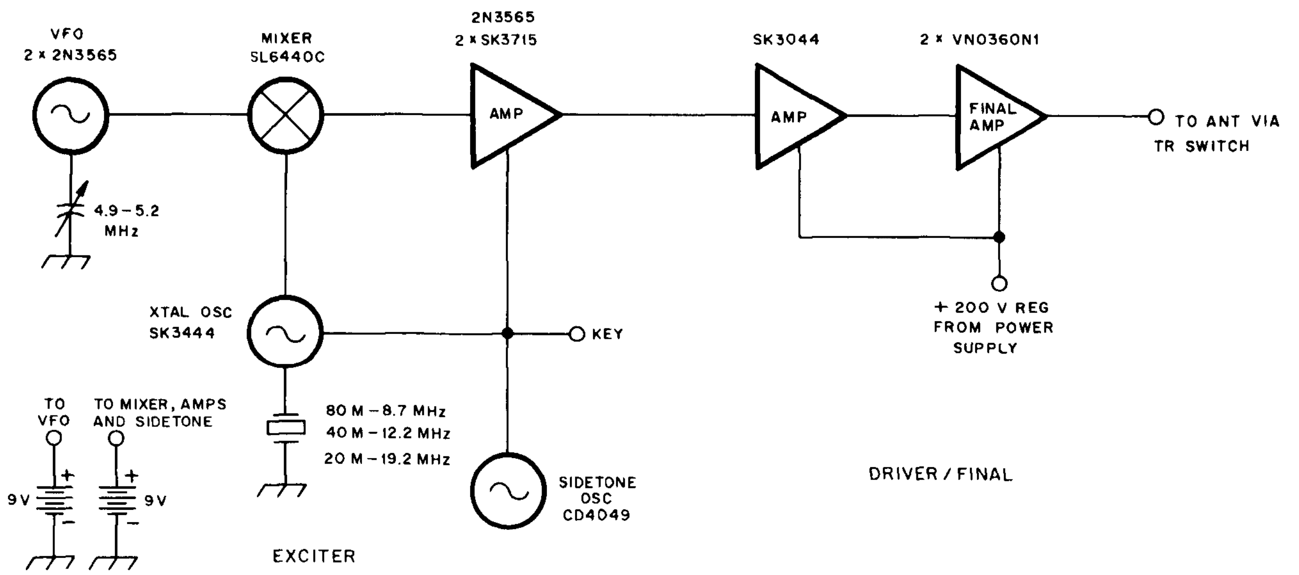
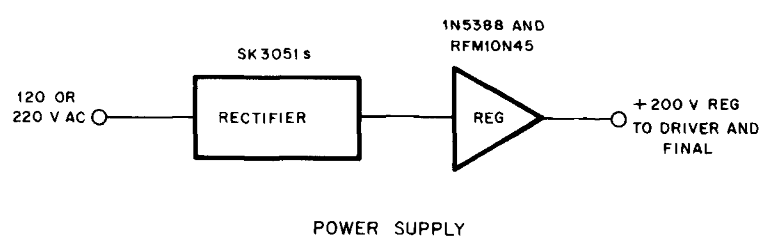


Fig 1—The portable transmitter consists of three units: exciter, driver/final and power supply. Because the driver/final is powered by a transformerless 200-V supply, careful construction and a polarized ac power cord are necessary to prevent operator contact with the ac mains. The exciter is battery-powered to avoid power-supply compatibility problems with the driver/final.



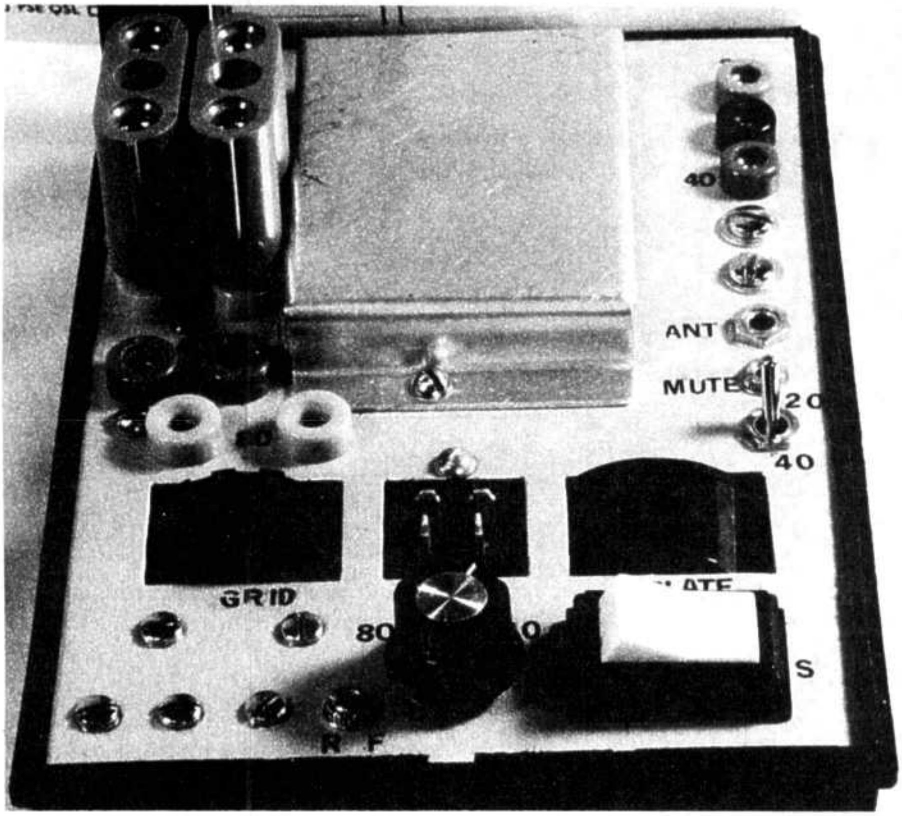


Fig 2—Exterior view of the driver/final module. The action of the final tuning controls is so similar to that encountered with a vacuum-tube amplifier that the controls were labeled GRID and PLATE instead of GATE and DRAIN! The front panel, which serves as a heat sink for the final-amplifier transistors, is *not* common to the negative lead of the driver collector/final drain supply; instead, it is connected to the ground wire of the three-wire ac power cord for safety. The final-amplifier transistors are mounted beneath the square metal shield. Insulated banana and tip jacks are used for most module interconnections and some band-switching functions; connection to the power supply is made by means of the four-pin TRW/Cinch/Jones plug behind the band-switch knob.

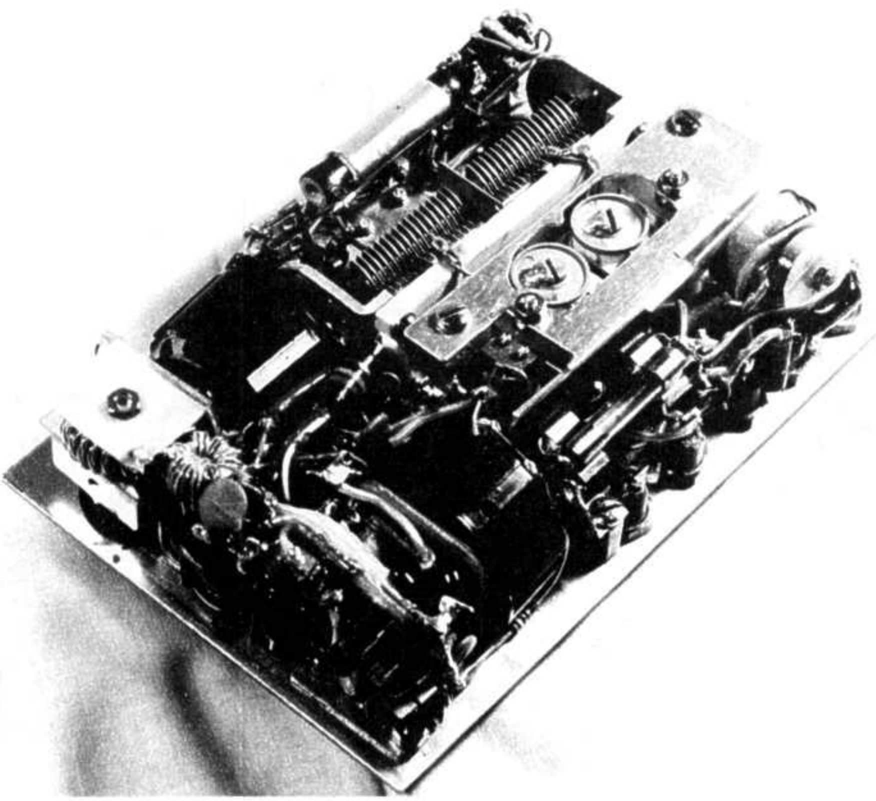


Fig 3—Interior view of the driver/amplifier module. The two-section variable capacitor tunes the balanced final-amplifier output tank. The ceramic trimmer capacitors adjust final-amplifier neutralization. The fins of the driver-transistor heat sink can be seen in the lowermost corner of the module.

example, the second and third harmonics are down 49 and 42 dB, respectively.) Final amplifier efficiency ranges from 25.9 to 56.4%, depending on band (or—on 80 m—on the band segment used). The transmitter is constructed in a 2 × $4\frac{1}{2}$ × $6\frac{1}{2}$-inch (HWD) plastic box (see Figs 2 and 3).

Exciter

See Figs 4 and 5. The exciter uses frequency conversion for stability and constant tuning rate on all three bands. The VFO uses two 2N3565 transistors in a permeability-tuned Hartley configuration. A Plessey SL6440C doubly-balanced mixer IC is used. The mixer LO port is driven by a Pierce crystal oscillator that uses an RCA SK3444 transistor.

The output of the SL6440C drives a 2N3565 tuned amplifier; this, in turn, feeds a common-collector Darlington amplifier consisting of two SK3715s. At this point, the exciter output is ready for application to the driver/final—and to an external frequency counter, if greater calibration accuracy is desired for the VFO. The sidetone oscillator, mixer, crystal oscillator, tuned amplifier and Darlington stages are keyed; ground-return keying is used.

Power Supply

The secret of the low weight and small size of this transmitter lies in its transformerless power-supply circuit. A full-

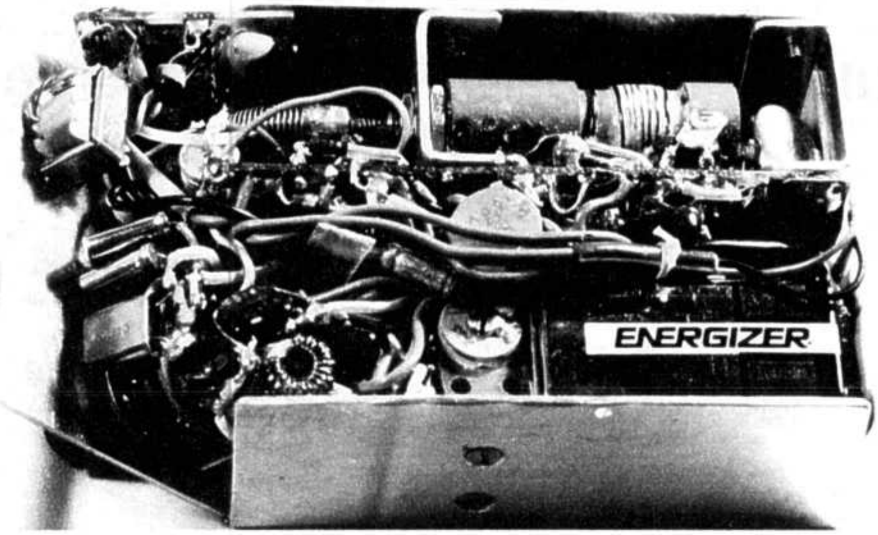


Fig 4—Interior view of the exciter. The band switch is at the lower left. The VFO inductor is in the compartment at the upper right; the inductor slug is turned by means of a flexible coupling and a turns-counting dial.

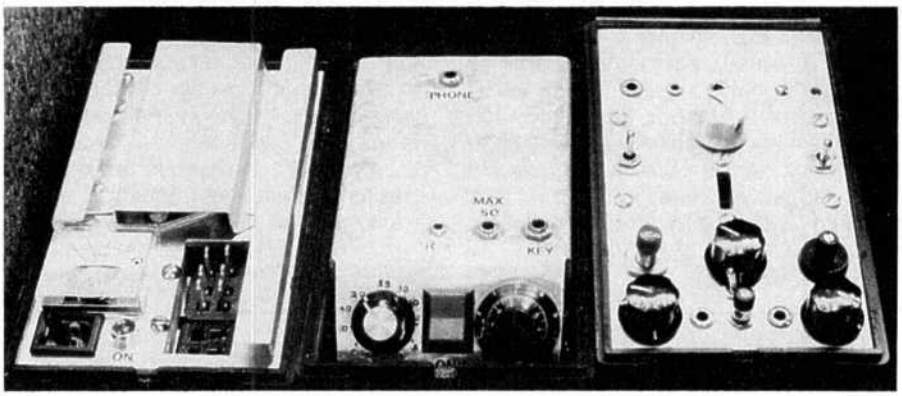


Fig 5—Exterior view of (from left to right) the power supply, the exciter and the receiver used in conjunction with the transmitter. The size of these modules is 2 × 3-3/8 × 5-1/2 inches (HWD). Although the transmitter described in this article covers only the 80, 40 and 20-meter amateur bands, the exciter also provides output on 15 and 10 meters for use with other transmitters at W6YBT. The receiver, a double-conversion design, uses IFs of 2900-3000 and 455 kHz.

wave voltage doubler is used when the ac mains supply is 120 V; a full-wave bridge is used when the mains supply is 220 V (common in Europe). The rectifier output is applied to a regulator designed around an RCA RFM10N45 power MOSFET (pass transistor) and a 1N5388 Zener diode (voltage reference). The supply's 200-V output is stable to within one percent.

Some of the Problems Solved

Designers of power supplies for low- and medium-powered equipment often depend (sometimes unknowingly) on the resistance of power-transformer windings to limit inrush current through filter capacitors. Applying power to a *transformerless* supply may result in disaster unless surge protection is *designed in*: The current-limiting level of the ac mains is far too high to protect power-supply components! Surge survivability in *this* transformerless power supply is achieved by several means. First, rectifier diodes with a high surge rating (SK3051s) are used. Further, the POWER switch is protected against burnout by means of surge-limiting resistors in series with the filter capacitors.

Current limiting *ahead* of the power-supply regulator was insufficient to prevent mysterious regulator-transistor failures, however. Investigation revealed that the charging currents drawn by RF bypass capacitors on the 200-V rail were sufficient to blow the regulator transistor, particularly when the regulator load was hot switched. Installation of surge-limiting resistors in the 200-V lines to the driver and final solved this problem.

That power MOSFETs are "nearly indestructible" is one myth; that they require "virtually no driving power" is another. In fact, the driving-power requirement of power MOSFETs generally increases with frequency. This is so because the capacitive reactance exhibited by a MOSFET's gate capacitance—and, therefore, the driver load impedance—*decreases* with frequency. Push-pull operation places the gate capacitances of two MOSFETs *in series*, making the MOSFETs easier to drive. Even so, the final amplifier is noticeably harder to drive at 20 m than at 80 and 40. Because of this, the amplifier is designed to operate in class B on 20 m and class C on the lower-frequency bands.

Ungraded power MOSFETs can exhibit considerable part-to-part variation in gate threshold voltage and transfer characteristics. Because of this, the current drawn by a parallel or push-pull pair of such parts will probably not divide evenly between the paired transistors unless steps are taken to equalize their current drain. To this end, the bias on each final-amplifier transistor can be adjusted separately.

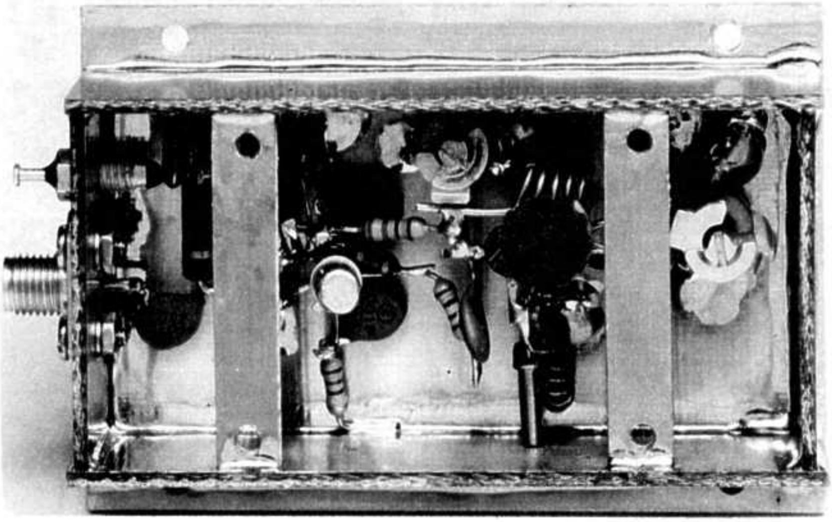
The final amplifier in this transmitter is neutralized for stable operation. Because the gate-to-drain capacitance of a power MOSFET is a function of drain-to-source voltage, neutralization must be carried out with full drain voltage applied to the final. For maximum stability, the neutralization must be touched up while the transmitter is operating at full output power.

Conclusion

High-voltage switching MOSFETs can provide useful CW performance if proper care is taken in their selection, biasing and neutralization. The use of a transformerless dc supply is feasible if: (1) effective surge limiting is included, and (2) care is taken to prevent operator contact with the ac mains *and* high-voltage supplies. The savings in weight and equipment size made possible by combining these techniques means that lightweight portable equipment need not be limited to QRP power levels.

A 192-MHz Local Oscillator for Home-Brewers

By Zack Lau, KH6CP
ARRL Laboratory Engineer



This circuit is the most suitable VHF oscillator design that I have found for construction at home. It avoids the need for frequency multiplication by using a 7th-overtone crystal to generate a 192-MHz signal *directly*. As a result, a costly spectrum analyzer is *not* needed for circuit alignment. A frequency counter is necessary to determine whether the crystal is oscillating on its 5th or 7th overtone—but if you're building an LO, you'll want to measure its frequency anyway.

The Circuit

See Fig 1. The design is basically a retuned version of one described by Robert Matthys in *rf design*.[1] Although Matthys did not push the design beyond 100 MHz, the circuit seems to work well at much higher frequencies, provided that careful construction techniques are used. The low-value base-to-ground capacitor in the original circuit is omitted; stray capacitance suffices at 192 MHz. Properly constructed, the oscillator drifts only a few tens of hertz at most—too little to measure accurately in the ARRL lab. The oscillator signal is amplified by an MMIC to a level of +11 dBm. (My application requires sufficient power to drive a splitter, which in turn feeds two doubly balanced diode mixers; +11 dBm is also suitable for driving frequency multipliers to get to 1152 MHz—a convenient LO frequency for 23-cm operation.)

Readers with a bit of capacitor knowledge may wonder about using disc-ceramic capacitors above their self-resonant frequency. Why not use them? Disc-ceramic capacitors are cheap, and they work just fine. Even at two or three times its self-resonant frequency, a disc-ceramic capacitor has a low enough

[1]Robert Matthys, "A High Performance VHF Crystal Oscillator Circuit," *rf design*, Mar 1987, pp 31-38.

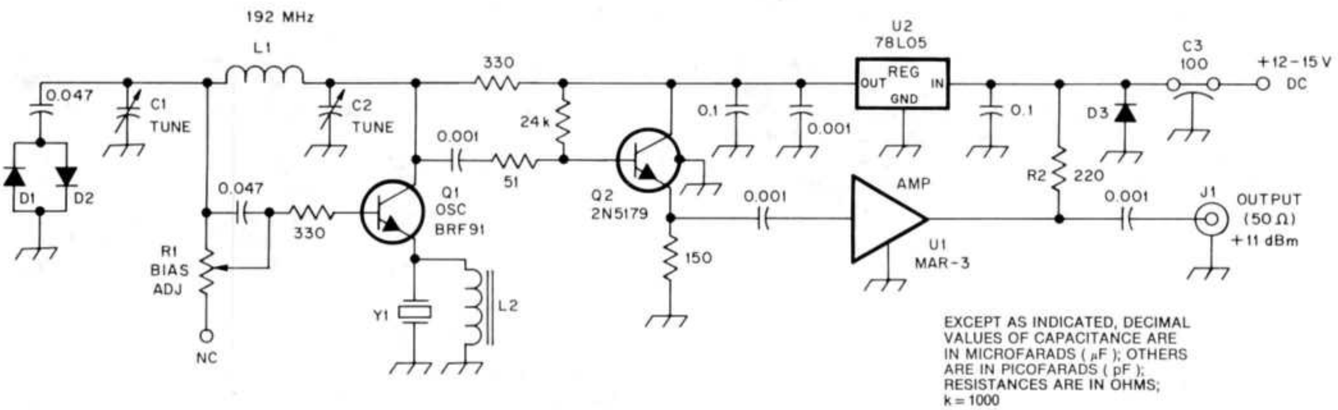


Fig 1—Schematic diagram of the 192-MHz oscillator. Capacitors are small monolithic or disc-ceramic units unless specified otherwise. Resistors are carbon-film or composition units. R2 is ½ W; all other resistors are ¼ W.

C1, C2—1.7- to 11-pF air-dielectric trimmer capacitor. Exact range not critical.
C3—100-pF feedthrough capacitor.
D1, D2—General-purpose, small-signal Schottky diode (1N5711, HP 5082-2835 suitable).
D3—1N4001.
J1—Female SMA connector.
L1—5 turns of no. 20 enameled or tinned wire on ¼-inch form, space-wound. Remove form after winding.
L2—10 turns of no. 28 enameled wire on T-25-12 toroidal core.
Q1—BRF 91, MRF 901.
Q2—2N5179.
R1—100-kΩ trimmer potentiometer.
U1—MAR-3 or MSA-0385 MMIC.
U2—78L05.
Y1—7th-overtone, 192-MHz crystal (International Crystal Mfg Company, Inc, part no. 477390).

series resistance to pass RF without much loss. Save the expensive chip capacitors for use at UHF and above, or power circuits in which RF losses become expensive to replace! At this power level and frequency, you can buy another MMIC for less than the cost of two high-quality chip capacitors!

Careful Construction Ensures Performance

The key to building this circuit and getting it to work without tedious fiddling is to *build it on a single copper ground plane*. Etched-trace boards don't work well at 192 MHz because of stray capacitance. Although you may be able to get the circuit to work on an etched board, the oscillator will probably drift much more than one built over a ground plane using "ugly" construction methods.

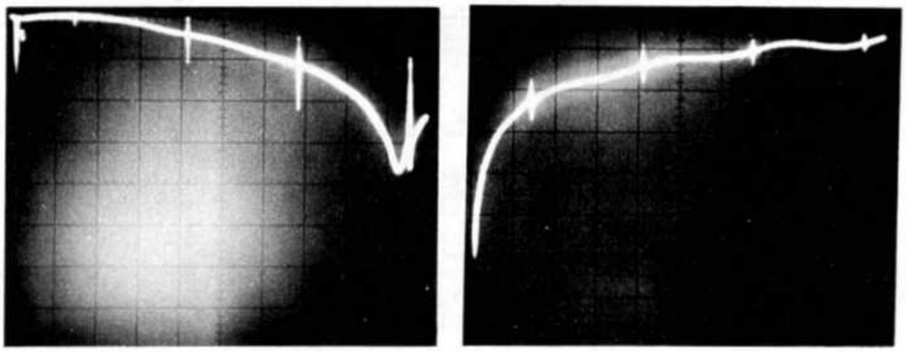
It's also important to build the circuit in a shielded box that isn't too small. (The inside dimensions of my latest version are 0.75 × 1.375 × 2.875 inches [HWD].) If the enclosure is too small, putting on the lid will detune the oscillator—possibly enough to stop oscillation. In particular, the air-core inductor, L1, requires space for its magnetic field. By placing the axis of the coil parallel to the ground plane, the "lid effect" seems to be minimized. Another technique that reduces the effect the lid's presence is to solder two 1-3/8-inch by 1/4-inch strips of copper-clad PC board across the top of the box about 3/4 inch from each edge (see the title photo). These were tapped to take the no. 4-40 screws used to hold down the lid. Placing the bodies of parts (such as capacitors) near the ground plane also helps performance by reducing energy radiation.

The best material for the oscillator box and ground plane seems to be unetched, double-sided, glass-epoxy circuit board. Bought at a hamfest, the stuff is pretty cheap. The fiberglass acts as a thermal insulator, aiding temperature stability. Stay away from phenolic board—it's too brittle. (If you don't have access to a metal shear, you can make straight cuts in fiberglass board material by scoring both sides and bending the board until it breaks.)

Resonating Inductor Important

Fig 2 illustrates the importance of the resonating inductor, L2. As Fig 2A shows, the inductor should resonate just below the desired frequency of oscillation. Without this inductor (Fig 2B), the stray capacitance across the crystal presents a relatively low impedance that may allow the oscillator to free run.

Retuning the oscillator for a different frequency can be done without sophisticated equipment by trial and error in conjunction with appropriate insight into what each circuit component does. For example, suppose you wanted a 5th-overtone oscillator, but your circuit would only tune to the 7th overtone—even though the tuning capacitors (C1 and C2) are at minimum capacitance. This would indicate that the inductance of the resonating inductor was too low to select the desired overtone.



A B

Fig 2—Series attenuation of a 192-MHz, 7th-overtone crystal with (A) and without (B) the resonating inductor. Horizontal scale: 20 MHz/div; vertical, 10 dB/div.

Circuit Adjustment

Tuning the circuit is quite simple. First, adjust R1 for a Q1 collector voltage of 2.5. Next, adjust C1 and C2 alternately for the best compromise between output frequency and power output. Unfortunately, the tuning range often isn't enough to hit the target frequency exactly—my oscillators are a few hundred hertz off with high-accuracy crystals, and a few kHz off with low-accuracy units. This isn't a problem when I use my Ten-Tec Argonaut as an IF; the Argo's frequency readout isn't any better. In my opinion, an oscillator with slight, *stable* frequency error is preferable to an oscillator that can be moved 20 or 30 kHz just by accidentally dropping it.

Fig 3 shows the output spectrum of the oscillator/amplifier module. Parasitic oscillations are at least 75 dB down. The presence of harmonics in this spectrogram may be of concern to some readers. Because this LO is designed to drive diode mixers, the ideal output waveform should be a square wave, which has *lots* of harmonics. An oversimplified way of looking at this is to consider the LO as a clock that turns the mixer on and off. Non-ideal performance occurs when a strong signal interferes with this switching. A little thought reveals that gradual switching (sine-wave LO) is more likely to be interfered with than rapid switching (square-wave LO). Besides, if you look at the LO port of the mixer with a scope, you'll see that the mixer diodes distort the LO signal no matter how well the LO is filtered! In short, ridding the output of an LO of harmonics is often a waste of time and effort.

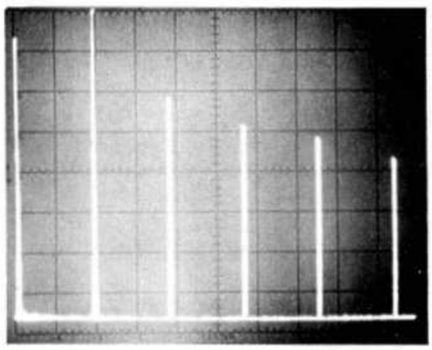


Fig 3—Spectrogram of the 192-MHz signal source. Horizontal scale: 100 MHz/div; vertical, 10dB/div. Parasitic oscillations are at least 75 dB down.

Troubleshooting

What if your version of the oscillator doesn't work? If possible, check for the presence of RF at the collector of Q1. Because 192 MHz is a rough go for the test equipment owned by most home builders, I won't specify a particular RF level for this measurement.

Even if the oscillator doesn't oscillate, the next step is to feed some sort of signal (preferably at a frequency between 50 and 250 MHz) through the buffer chain to see if *it* works. (It's much easier to debug an oscillator circuit if the buffer chain works; the buffer allows you to look at the oscillator output without upsetting things.) If a new design frequency is the goal, the biggest problem will probably be choosing a value for L1.

Conclusion

This oscillator design takes advantage of VHF crystals that have recently become available for frequencies between 100 and 200 MHz. By directly generating a signal at 192 MHz, this circuit eliminates a multiplier stage and the filtering necessary to clean up a multiplied signal. Result: a 192-MHz LO that seems to be as clean and stable as other designs. See you at 220.1!

Components

By Mark Forbes, KC9C
823 N 15th St
Canon City, CO 81212

One big problem we all run up against when building is where to obtain the parts that we want to use. Whenever possible, I'll list sources of small quantities of the parts that appear in the column. In many cases, I can only list the manufacturer; a call to them may be able to yield a source of small quantities in your area. In other cases, minimum orders may prevail. If you see a component that you'd like to purchase but can't find someone to sell you "onesies and twosies," drop me a line. If enough people are interested, perhaps we can pool our needs to meet the minimum.

Micrometals Toroidal Cores

Toroids are quite useful in many RF circuits, often being preferred to other core types. Micrometals, Inc (1190 N Hawk Circle, Anaheim, CA 92807) offers a wide selection of cores and mounts for radio frequencies ranging from 20 kHz to 250 MHz. Their catalog no. 3 *(Toroids for RF Applications)* lists hundreds of cores, and contains other useful data such as saturation curves, Q-v-frequency curves and a toroidal winding table.

Micrometals sells several engineering kits that will be of interest to experimenters. Each kit contains a selection of cores. Two kits are reasonably priced and contain enough cores to get an experimenter going.

Engineering kit no. 21 contains 15 different core designs (59 pieces in all) for the 20-kHz-to-3-MHz range, and the cost is $50. Engineering kit no. 23 is available for $40, and contains 61 pieces covering 250 kHz to 30 MHz. You might also be interested in Micrometals' other catalogs: *Iron Power Cores* (catalog no. 1), *Shielded Coil Forms* (catalog no. 2) and *EMI/Power Toroids and E Cores* (catalog no. 4).

XICOR E²POT

Xicor (851 Buckeye Ct, Milpitas, CA 95035-7493), one of the pioneers of nonvolatile semiconductor memories, has introduced an interesting product, the E²POT™. The part is a digitally programmed potentiometer that remembers its setting even after power is removed. This could be quite a handy little part.

The parts are specified with minimum resistance, maximum resistance and resistance step size. All available potentiometers have a minimum resistance of 40 ohms. The 10-kΩ potentiometer has steps of 101 ohms, the 50-kΩ potentiometer increments in 505-ohm steps and the 100-kΩ version can be set in 1010-ohm steps. A 1-kΩ version (10-ohm steps) will be introduced shortly. All are packaged in 8-pin plastic miniDIPs. Price: about $3.

Signetics CMOS Frequency Synthesizer

Signetics (811 E Arques Ave, Sunnyvale, CA 94088), has introduced a number of RF-related parts in the past year, many of which are aimed at the lucrative cellular-telephone market. One such part, the NE602 mixer/oscillator, appeared in an interesting receiver design in *QST*.[1] Another part, the TDD1742, may be of equal interest. The TDD1742 is a single-chip CMOS frequency synthesizer. This part is primarily targeted for use in small VHF and UHF transceivers, although it's capable of operating from HF through UHF. Dc supply voltage range is 7 to 15, and the maximum reference frequency is 8.5 MHz.

The synthesizer can be programmed with a microprocessor or an external ROM/EPROM. This adds the flexibility of using the synthesizer in a circuit that does not contain a processor.

RF Power Amplifiers

If you think that the only source of high-power solid-state RF amplifiers is Motorola, take a look at what some other companies offer, notably TRW, Thomson-CSF and Philips, among others.

For example, TRW (RF Devices Division, 14520 Aviation Blvd, Lawndale, CA 90260) has recently introduced a 2-watt hybrid amplifier that has 18.5 dB gain. The CA2885 covers 40 to 500 MHz, making it an ideal general-purpose VHF and UHF driver for a final amplifier. The amplifiers operate from 24 V dc, and the price is around $50 each for quantities in the hundreds.

GE Surge Suppressors

One of the biggest gremlins in Amateur Radio has always been ac-line voltage surges that can destroy power supplies in expensive equipment. Lightning- and wind-induced faults, both of which occur during storms, are the most frequent culprits. Protection against such transients is available in rather inexpensive form with MOV™ and Surgector™ products from GE.

MOVs (metal-oxide varistors) are generally used across the ac line or on the low-voltage side of the supply. They're chosen to clamp at a voltage that will not disrupt normal circuit function but will not allow large surge voltages to propagate. The devices respond in microseconds. MOVs are available in voltages ranging from 5 to 3500 and can handle up to 70,000 A peak (and 10,000 joules of energy!).

A Surgector is a combination of a Zener diode and an SCR. The Surgector is in the off state at low voltages; in this state, it has a leakage current of only 50 nA. When the Surgector's clamping voltage is exceeded, it turns on and the Zener conducts. A few nanoseconds later, the SCR begins to conduct, shunting heavy currents to ground. When the transient has passed and the line returns to normal, the devices shut off.

For information on MOVs and Surgectors, call General Electric Solid State at 800-443-7364.

[1]John Dillon, "The Neophyte Receiver," *QST*, Feb 1988, pp 14-18.

Bits

MONITOR 1-232

A new dedicated RS-232-C interface, known as the MONITOR 1-232, adapts up to 8 discrete and 3 window inputs to an RS-232-C output format. It operates from a single power supply in the range of +5 to +30 V dc, or with a furnished ac adaptor. The MONITOR 1-232 is inexpensive and easy to install and use. For $99.95, you'll receive the MONITOR 1-232, an ac power adaptor, software and instructions. Contact CEI of Florida Enterprises, Inc, PO Box 16804, St Petersburg, FL 33733, tel 813-822-3001.
—*Maureen Thompson, KA1DYZ*

By Bill Olson, W3HQT
Box 2310, RR 1
Troy, ME 04987

Inexpensive Low-Power Microwave Terminations

This is going to be a short column. Preparing for (and going to) Dayton took most of my time this month. The Dayton trip was worth it, though, and I enjoyed meeting a bunch of you there.

This month's project comes from my good friend Dave Mascaro, WA3JUF. He has some ideas for inexpensive low-power microwave terminations that work fine up to 2.3 GHz. For those who don't have access to a lab full of test equipment, these loads are just the ticket for tuning up new gear.

Fig 1 shows a simple 1-W termination built on a four-hole SMA chassis-mount, flange-type connector. You can use a male or female SMA connector—your choice. In addition to the connector, you need four 220-Ω, ¼-W carbon-composition resistors and two brass caps from Johanson piston trimmer capacitors. Drill the brass caps as shown, and slide them over the resistor leads. Push the caps flush against the resistor bodies and solder the leads to the cap. Cut off excess lead length and solder the resistor assembly between the SMA connector body and center pin as shown. At 1.3 GHz, the measured return loss (RTL) for this termination is 28 dB (SWR = 1.08:1), and at 2.3 GHz the RTL is still 20 dB (SWR = 1.22:1).

A 3-W load, again built on an SMA flange-type connector, is shown in Fig 2. This load uses four 220-Ω, ½-W carbon-composition resistors. Instead of the brass caps, the resistors in this design are soldered to a couple of small pieces of 0.010-inch-thick brass sheet (thickness is not critical). Cut and drill the end pieces as shown, solder them to the resistors and trim the leads. Make the round end piece as small as possible to minimize capacitance between it and the connector flange. Again, it's very important to get the end pieces flush with the resistor bodies to minimize lead inductance. Solder one end of the resistor bundle to the SMA center pin. Next, form two side pieces from brass sheet and solder them between the other end of the resistor bundle and the body of the SMA connector.

At 1.3 GHz, the measured RTL for the 3-W termination is 22 dB (SWR = 1.17:1), and at 2.3 GHz, RTL is 17 dB (SWR = 1.4:1). The RTL can be improved by adjusting the spacing between the resistor bundle end piece and the connector flange. You can solder a heat sink to the brass top of the 3-W load to increase power dissipation.

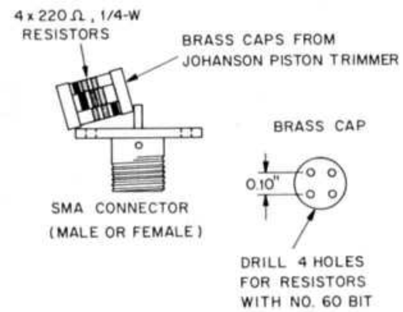


Fig 1—Construction details of the 1-W microwave termination.

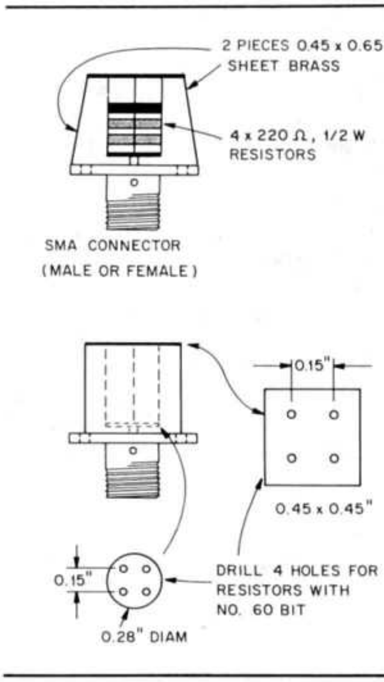


Fig 2—Construction details for the 3-W termination.

CONFERENCES

For those who like to travel, here are a couple of very important upcoming VHF/UHF events.

The 22nd annual Central States VHF Conference will be held July 21 to 24, 1988, at the Villager Motel in Lincoln, Nebraska. The tentative program includes the following talks: "The Challenge of Microwave EME" and "How to Build VHF/UHF Preamps" by Kent, WA5VJB; "How to Predict 144 MHz Sporadic E Openings" by Joe, WAØWRI; "Some Thoughts on PUFF—A Public Domain Microwave Circuit Analysis and Layout Program" and "Continuing Ideas on 903 MHz" by Jerry, N6JH; "Match vs Noise Figure Trade Offs in VHF Preamps" by Al, WB5LUA; "The Latest in Computer Aided Antenna Design" by Roger, WBØDGF; "Moon Data for 1988-1989" by Derwin, W5LUU; "Spatial Polarization and Faraday Rotation" by Tim, KL7WE; "Update on 6 Meter EME" by Ray, WA4NJP; "Simple Magnetometers, or How to Measure Your Own K Index" by Russ, W4WD; "Receiver Front End Protection" by Charles, AF8Z; "902 MHz Transverter, Power Amplifiers and Antennas" by Don, WØPW; "History of North American VHF Contesting" by Curt, K9AKS and Ed, WØOHU; "Additional Notes on June 14, 1987, Sporadic E Opening" by Mike, W9IP. For pre-registration information, write to Roger Cox, WBØDGF, 3451 Dudley, Lincoln, NE 68503-2034. *(Information courtesy WBØDGF's Midwest VHF Report)*

Microwave Update 88 will be held in Estes Park, Colorado, on August 25 to 28, 1988. This is a great conference for those of you interested in the bands at 902 MHz and above. The mountains around Estes Park are well worth the trip, too. Currently, 14 talks are planned, but the final program may be different. Presentations include: "Phase Locked Oscillators" and "Integrated Beacons" by Charles Osborne, WD4MBK; "Antenna Themes" by Margarete Ralston, KI4VE; "Microwave Cavity Amplifiers" by Ott Fiebel, W4WSR; "Computer Aided Design of Amplifiers" by Al Ward, WB5LUA; "Amateur Laser Work" by Dave Chase, KY7B; "Multiple Use Equipment" by Paul Finnell, W7EFQ; "2304 MHz No Tune Transverter" by Rick Campbell, KK7B; "10 GHz Transverters" by Bob Larkin, W7PUA; "Printed Filter Update" by Jim Davey, WA8NLC; "Smith Chart" and "Edge Coupled Filters" by Dave Haupt, W8NF; "Satellite Update" by Bill McCaa, KØRZ. For more information and registration information, contact Don Hilliard, WØPW, PO Box 563, Boulder, CO 80306. *(Information courtesy Don Hilliard, WØPW)*

Correspondence

Some time ago, I acquired 150-MHz transmit and receive ACSSB boards (along with a service manual) from STI. Does anyone know if there any active ACSSBers in the New York area? What frequency or frequencies are used by ACSSBers? Is there a source for TCXO modules? I'd greatly appreciate learning of any sources of information concerning ACSSB operation and equipment.—*Lynn M. Finch, W2MSJ, RD 2, Box 789, Rt 369, Port Crane, NY 13833.*

Diversity Technique Bibliography—Additional Material

Thanks are due to Domenic Malozzi (N1DM) for his useful bibliography on diversity techniques.[1] I have compiled some additional citations to help round out his list. Even taken together, the two bibliographies give only a small sampling of the considerable literature that has been published on this topic, particularly in the area of VHF/UHF mobile-radio diversity techniques (developed) in the last two decades or so.

Given the volume of literature available, and the fact that much of it is relatively ancient and sometimes difficult to find, books that contain compilations of the more important papers, or summarize the results of research in the area (such as the first three books listed), are particularly useful.—*Barry McLarnon, VE3JF, 2696 Regina St, Ottawa, Ontario, K2B 6Y1 CANADA*

Key to Citations

PROC IEE = *Proceedings of the Institution of Electrical Engineers* (published in London).

IRE Transactions on Communications Systems is a specialized journal that was published by the Institute of Radio Engineers. It is a separate publication from the *PROC IRE.*

Brayer, K., ed, "Data Communication via Fading Channels," IEEE Press, 1975.

Jakes, W. C., ed, "Microwave Mobile Communications," Wiley, 1974.

Schwartz, M., Bennett, W. R. and Stein S., "Communication Systems and Techniques," McGraw-Hill, 1966.

Allnatt, J. W., Jones, E. D. J., and Law, H. B., "Frequency Diversity in the Reception of Selectively Fading Binary Frequency-Modulated Signals," *PROC IEEE*, Vol 104, Part B, Mar 1957, pp 98-110.

Notes

[1]D. Malozzi, "Diversity Technique Bibliography," Correspondence, *QEX*, April 1988, pp 3-4.

Bronzi, G., "A Laboratory Comparison of Maximal-Ratio Combiners and Optimal Selectors for Fading HFSK Channels," *IRE Transactions Communications Systems*, Sep 1961, pp 320-321.

Gott, G. F. and Dutta, S., "Improved Diversity Combiners in HF Interference," *IEE Colloquium Digest* 1979/48 (Conference on Recent Advances in HF Communication Systems and Techniques), Feb 1979, pp 110-115.

Kawachika, N. M. and Villard, O. G., "Computer Simulation of HF Frequency-Selective Fading and Performance of the Mode-Averaging Diversity Combiner," *Radio Science*, Vol 8, No. 3, Mar 1973, pp 203-212.

Lee, W. C. Y. and Yeh, Y.S., "Polarizaion Diversity System for Mobile Radio," *IEEE Transactions Communication Technology*, Vol COM-20, No. 5, Oct 1972, pp 912-923.

Lee, W. C. Y., "Close-Spaced Diversity Antenna for HF Transmission," *IEEE Journal on Selected Areas in Communications*, Vol SAC-5, No. 2, Feb 1987, pp 270-273.

Lewin, L, "Diversity Reception and Automatic Phase Correction," *PROC IEEE*, Vol 109B, No. 46, Jul 1962, pp 295-304.

Lyons, W., "Error Protection via Time Diversity," *IEEE Transactions on Communication Technology*, Jun 1965, pp 234-237.

McManamon, P. and Janc, R., "An Experimental Comparison of Nondiversity and Dual Frequency Diversity for HF FSK Modulation," *IEEE Transactions on Communication Technology*, Dec 1968, pp 837-839.

McManamon, P., Janc, R. and Tsai, S., "Comparison of Diversity, Nondiversity, and Coding on a Parallel FSK HF Channel," *IEEE Transactions on Communication Technology*, Vol COM-17, Jun 1969, pp 355-367.

Parsons, J. D. et al, "Single-Receiver Diversity Systems," *IEEE Transactions on Communications Technology*, Vol COM-21, No. 11, Nov 1973, pp 1276-1280.

Parsons, J. D. et al, "Diversity Techniques for Mobile Radio Reception," *IEEE Transactions on Vehicular Technology*, Vol VT-25, No. 3, Aug 1976, pp 75-85.

Porter, G. C., "Measurement of Polarization Statistics of Signals Received Over a Short Range HF Path," *IEEE Transactions on Communications Technology*, Vol COM-14, No. 4, Aug 1966, pp 484-494.

Shortall, W. E., "A Switched Diversity Receiving System for Mobile Radio," *IEEE Transactions on Communications Technology*, Vol COM-21, No. 11, Nov 1973, pp 1269-1275.

Stein, S., "Fading Channel Issues in System Engineering," *IEEE Journal on Selected Areas in Communications*, Vol SAC-5, No. 2, Feb 1987, pp 68-69.

Vaughn, R. G. and Andersen, J. B., "Antenna Diversity in Mobile Communications," *IEEE Transactions on Vehicular Technology*, Vol VT-36, No. 4, Nov 1987, pp 149-172.

Vigants, A., "Space Diversity Engineering," *Bell Systems Technical Journal*, Vol 54, No. 1, Jan 1975, pp 103-142.

Villard, O. G., Lomasney, J. M., and Kawachika, N. M., "A Mode-Averaging Diversity Combiner," *IEEE Transactions on Antennas and Propagation*, Vol AP-20, No. 4, 1972, pp 463-469.

Feedback

Please refer to my article, "Extracting Stable Clock Signals From AM Broadcast Carriers For Amateur Spread Spectrum Applications," *QEX*, Oct 1987. There is an error in the schematic on page 7. The numbering of pins 14 and 15 of U4A should be reversed. The junction of R2 and the 12-pF capacitor connect to pin 15 (which should be labeled R_{ext}, C_{ext}), and the other capacitor lead connects to pin 14 (which should be labeled C_{ext}).—*Andre Kesteloot, N4ICK, 6800 Fleetwood Rd, McLean, VA 22101.*

☐ There is an error in the Bits item on p 15 of the May 1988 issue of *QEX*. Under "RSGB News—What's Happening Across the Pond?," the dates shown (July 22-23, 1988) are incorrect. The correct dates are July 29-31.—*Ron Broadbent, G3AAJ, Hon Sec/Treasurer, AMSAT-UK, 94 Herongate Rd, Wanstead Park, London E12 5EQ ENGLAND.*