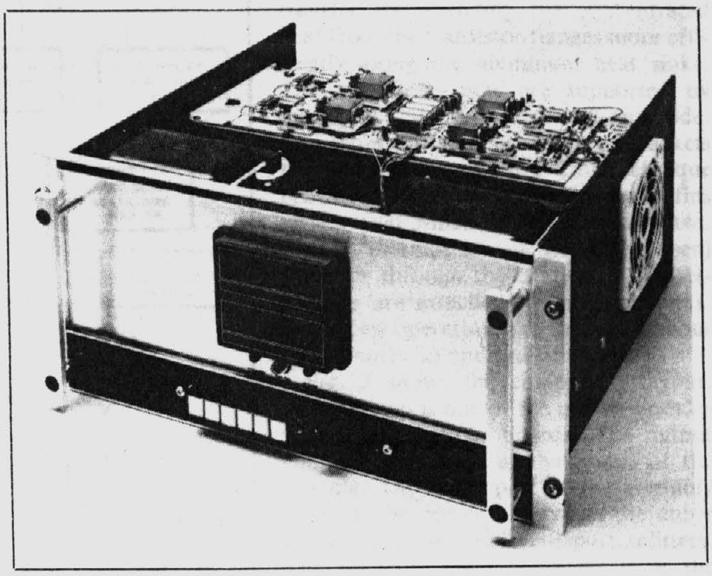
MOSFET RF Power: An Update

Part 1: Power FETs are now practical for 2-30 MHz broadband-amplifier use at the kilowatt level. How do they compare with vacuum tubes and bipolar transistors? This report by ARRL TA Granberg provides some interesting answers.



By Helge Granberg,* K7ES/OH2ZE

Interested in power-FET technology? This paper combines a status report with some useful application notes for MOS power devices. It is intended to document the progress that has been made recently in power-FET development and deployment, but it is not meant as a construction article. The data presented are, however, suitable for amateur designers who wish to develop their own power-FET amplifiers.

Early power FETs were relatively low-level devices, but now these transistors can produce power amounts of 100 to 150 W. In contrast, we now have bipolar devices that yield up to 250 W of rf power (500-600 W in special water-cooled packages from at least one manufacturer). If we compare watts against dollars, using an 8877 tube as an example, a watt from a solid-state amplifier is about twice as expensive. If we include the necessary harmonic filters for broadband amplifiers, the price will be even higher.

Narrow-band solid-state amplifiers are considered feasible only for single-band use. This is because the low impedance levels (typical) make band switching the passive elements impractical. Also, losses will result.

Push-pull solid-state amplifiers aren't difficult to design, and they are desirable because the impedances are higher. Also, rf ground loops are easily eliminated. Other advantages are that the powers of the two devices are combined and the even harmonics are suppressed. Broadband-transformer matching is suitable up to vhf. Other techniques, such as coaxial or other transmission lines (to provide a 180° phase shift) are practical up to the microwave spectrum.

The output harmonic filters of amplifiers usually are designed for 50 ohms, which makes them easy to switch. One low-pass filter usually covers less than an octave, but the frequency can be varied within the filter response without tuning or switching in a new filter. Because the amateur bands are one octave apart (except for 10, 18, 21 and 24 MHz), a separate filter is required for each band. More on this later.

Amplifier Specifics

The components for the circuit of Fig. 1 are not available from a single source. Many are engineering samples that were obtained from various manufacturers. As stated earlier, this treatment is conceptual rather than practical.

This amplifier provides a power output of 1600-to 1800-W PEP or cw, inclusive of the 0.3- to 0.4-dB filter losses, depending on the operating frequency. A nominal 40 W of drive is needed for full output. The input line contains an attenuator (selectable for 1, 2, 3 or 6 dB) to make the amplifier compatible with various commercial exciters, and to comply with FCC regulations. Over 2 kW of output power is possible with the 16 Motorola MRF150 MOS field-effect transistors used, but the power supply is rated only for 2800 W — the limiting factor.

The main power supply, shown in Fig. 2 (60 V no load, 48 V full load), consists of two smaller supplies. Each operates one of the large power modules. A regulated supply would dissipate some 400 to 500 W in the regulating process and would greatly increase the total weight of the system. A switching-mode supply would be rather complex for this power level, and would require RFI shielding.

Why FETs?

Power FETs have these definite advantages over bipolar transistors in this application:

1) More tolerant to load mismatch.

2) Simplified circuit design and biasing.

3) Lower high-order IMD (comparable to vacuum tubes).

4) Easier to make broadband because of higher input Z.

5) Gain can be controlled by varying the bias voltage. This can be used for alc shut-down instead of PIN-diode switches in the rf input. Linear alc can be had for ssb, but excessive bias reduction will deteriorate the IMD.

6) Higher power gain. The increase at 30 MHz can be 3 to 6 dB.

Industrial interest in power FETs probably relates to item 3. High-order IMD (9th order and up) causes adjacent-channel "splattering." This would also happen with an over-driven or mistuned tube amplifier. Low-order IMD (3rd and 5th) can be as high as −20 dB, and the signal will sound good if the high-order products are absent. Thus, the FCC specification is for only −25 dB on the 3rd-order product, and −60 dB or more for the 9th-order product and above (Marine).

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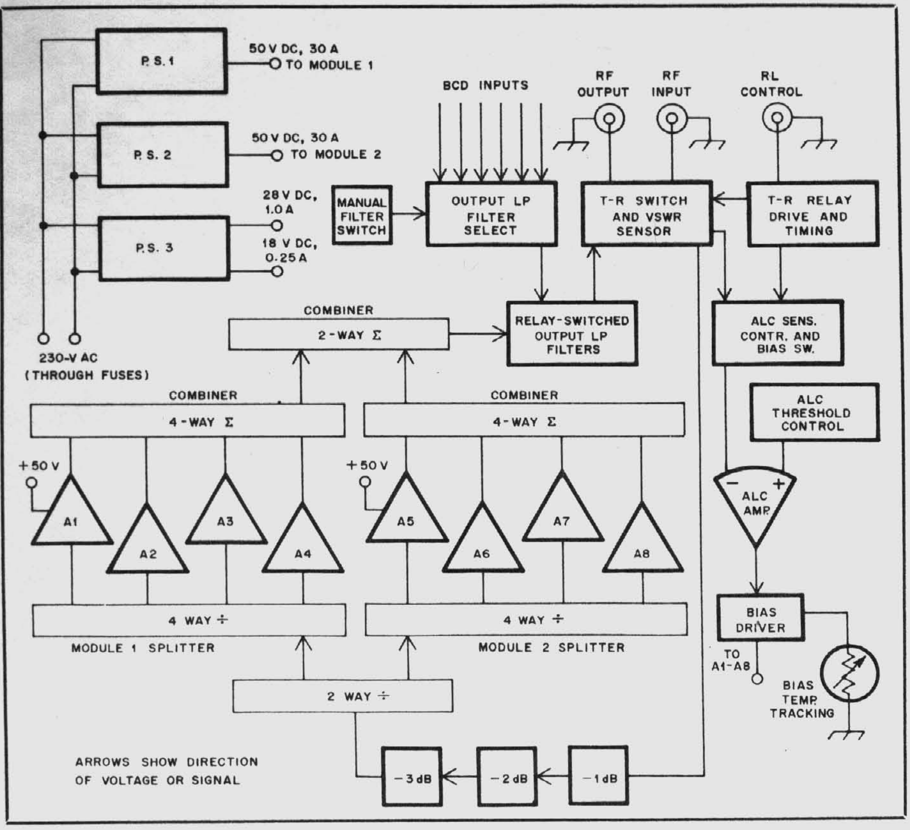


Fig. 1 — Block diagram of the 2-30 MHz power FET linear amplifier. A1-A8, inclusive, are individual FET amplifiers. See text for circuit description.

Bipolar transistors (depending on the type and internal structure) usually produce much more high-order IMD than is the case with FETs or tubes, unless the bipolar devices are biased to Class A and operated at reduced power.[1]

The power gain of a common-source FET amplifier can be varied 20 dB or more by adjusting the bias voltage. All present rf power FETs are enhancement-mode MOS types: If the gate and source are at the same potential, there is no current flow through the drain. The gate of an N-channel device must be positive with respect to the source in order to "turn on" the transistor. The gain-control range depends on the initial gate-threshold voltage (1 to 4 V typical) and the amplitude of the voltage swing at the gate. The transistor will be turned off completely and become an attenuator if reverse bias is applied.

In this amplifier, the bias is lowered only to near ground potential — a gain reduction of 8 to 10 dB. This can be sufficient in the event of a shorted or open load condition. Protective control voltage is obtained from a reflectometer at the amplifier output (Fig. 1). This amplifier also includes a self-contained linear alc system. Alc voltage is not fed back to the exciter as is done normally. The control range is limited to about 2 dB, beyond which it would degrade the IMD excessively. Bipolar-transistor amplifiers usually employ a variable voltage or current attenuator at the input port. PIN diodes are often used, but for linear alc an elaborate circuit is needed to prevent harmonics and distortion.

General Description

A block diagram of the total sytem (exclusive of the digital panel meters) is shown in Fig. 1. One meter reads 0-199.9 V and the other is for 0-199.9 mV. The latter one reads current and has 50-A and 50-mA shunts. The shunts (3) are shown in Fig. 2 as 0.001-ohm resistors. The shunts are used to read the individual currents of the rf modules, and one shunt is utilized to monitor the total current. The voltmeter monitors the nominal voltages of the main supplies, or about one half the voltage difference when switched to both supplies.[2,3] This metering system eliminates the need for heavy-duty wiring and prevents large currents from flowing through the switches.

There are two main rf modules (Fig. 1). Each contains four push-pull amplifiers. Combiners are used to produce a summed output of 1600 to 1800 W. Two pieces of Aavid Engineering heat-sink extrusion (no. 60140) support the amplifier boards, the combiners and other components of each module. Four copper heat spreaders (each 3/16 × 2 × 5-1/2 inches; mm = 25.4 × in.) are mounted individually on the flat surfaces of the heat sinks. The layout permits four MOSFETs (two boards) to be mounted on each of the heat spreaders. Copper is nearly twice as good as aluminum for heat conduction. Hence, it improves the instantaneous heat transfer by spreading the concentrated heat from the transistor flanges more efficiently along the aluminum heat sinks. The two heat sinks are supported by means of aluminum plates on each side. The plates also serve as mounting brackets for the whole structure. This technique provides a channel with the heat-sink fins inside. Two 5-inch fans (actuated by two 75° C thermostats — normally open) force air through the channel. The thermostats are attached to the heat sinks. During cw operation, the fans cycle about two minutes on and five minutes off.

Fig. 3 shows the component layout. Near the top is one of the power modules with four amplifier boards. The output combiner is shown at the center of the module. The lower part of the assembly contains the power supply. At the upper left are the two- and four-port splitters. The two-port main combiner is at the upper right.[4]

The transistor leads are pressed down (not soldered) against the pc boards and related contact areas. Teflon rings, then silicone-rubber rings, followed by aluminum rings, ensure firm contact when pressed in place by means of special standoffs. This method makes field service easier when replacing a transistor, since no soldering is required.

Filtering and T-R Circuit

The main combiner output is fed to a bank of low-pass filters. These are relay-switched for the desired band. The front-panel control switch operates when the BCD inputs to the filter-select circuit are open or high. When one or more of the BCD inputs are grounded, the manual switch is disabled and a light indicates the filter that has been selected by the code. This feature is useful for automatic band changing with transceivers that are designed for computer control.

Output from the filters goes to a T-R switch that consists of two Kilovac HC-1/530 vacuum relays. One is located at the amplifier input and the other is at the amplifier output. The relays are housed in separate shield enclosures (with BNC interconnect) to minimize unwanted crosstalk. T-R relay timing and drive signals must occur in a precise timing sequence to prevent "hot switching" in the amplifier output. Thus, the output relay must be energized first and released later than the input relay. For full QSK, the delays should be minimized. In this circuit the limit is about 8 ms, owing to the speed of the relays. Longer delays would hardly affect the QSK operation, but would shorten the marks for RTTY and cw, which would be apparent at high operating speeds. The control signal from the transceiver requires a key-to-ground

[1] Notes appear on page 16.

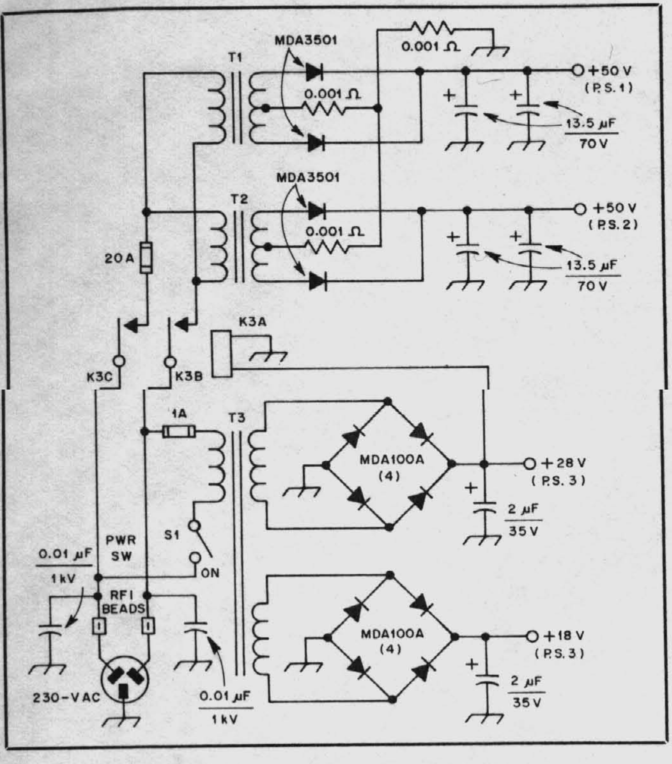


Fig. 2 — Schematic diagram of the FET amplifier power supply. Capacitors are disc ceramic, except those with polarity marked, which are electrolytic. Z1 and Z2 are ferrite beads for RFI suppression (Fair Rite no. 2673021801).

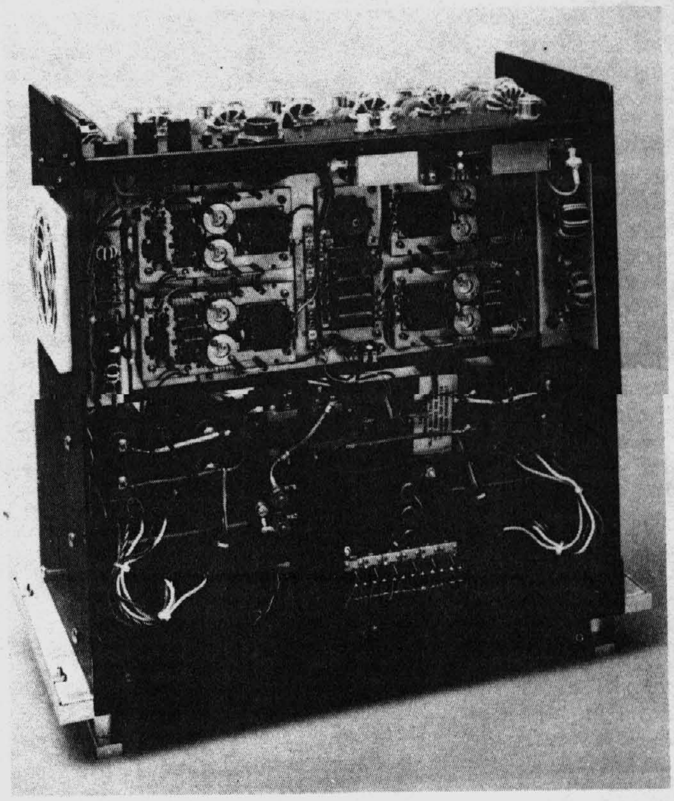


Fig. 3 — The inside of the solid-state amplifier. See text for discussion of module positioning.

polarity. The T-R, timing and drive-circuit module is visible at the upper right in Fig. 3 (just under the rear panel).

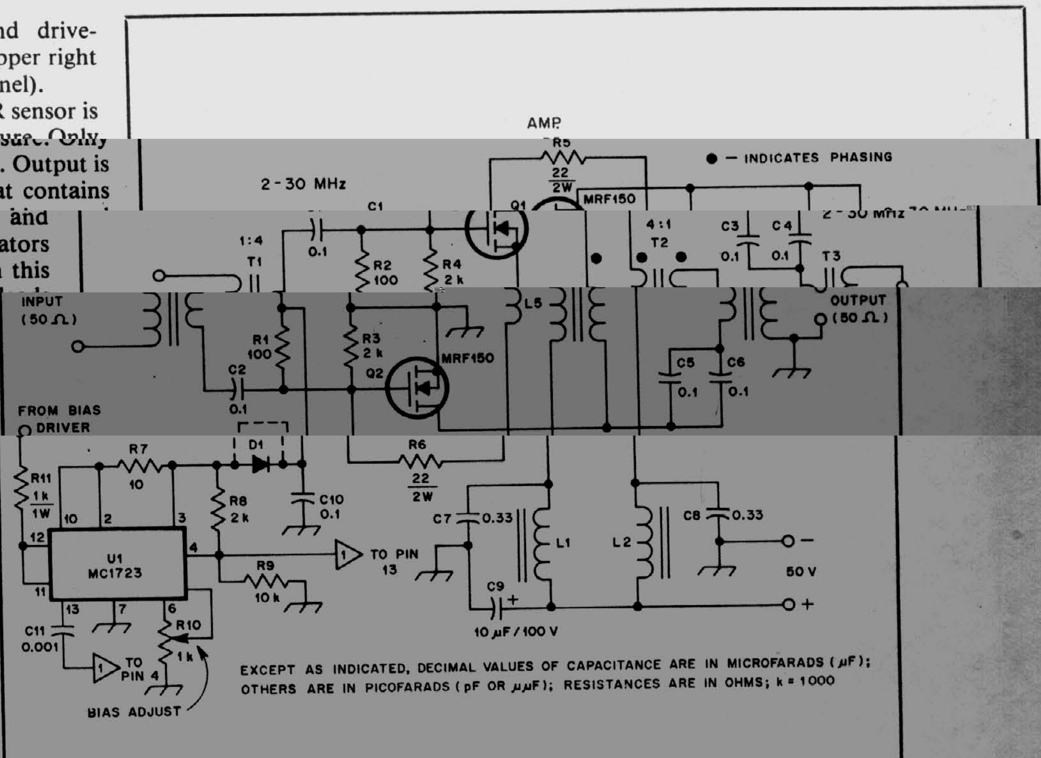
A reflectometer type of VSWR sensor is housed in the T-R circuit enclosure. Only the reflected power is measured. Output is routed to the circuit board that contains the alc, bias-temperature tracking and automatic filter-select circuits. Regulators for 12 and 24 V are also located on this ard. Ferrite beads are used on the leads at enter this board. This prevents rf om getting into the alc amplifier and MOS logic circuits.

The T-R output relay control voltage is uted also to the alc circuit. This turns f the transistor bias during standby, ereby preventing the 400-W standby ssipation (500 mA per transistor) of the devices.

The bias-temperature tracking feature eps the idling current constant with in-easing heat-sink temperature. Normally, would approximately double from 25° to 75° C. This function is handled by a ermistor that is coupled to one of the at sinks. Idle-current variation is 20% less.

ircuit Details

Fig. 4 contains a schematic diagram of ne of the push-pull amplifiers. The cir-uit is much simpler than one with bipolar ansistors. The external resistor values



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k = 1000

Fig. 4 — Schematic diagram of one of the power modules, inclusive of the bias regulator. C1, C2 and C10 are monolithic capacitors. C3-C8, inclusive, are ceramic chip capacitors. C9 is electrolytic and C11 is disc ceramic. Fixed-value resistors are carbon-composition, 1/2 W, unless noted otherwise. R10 is a Trimpot®. See text for discussion of the remaining components.

for the bias regulator permit regulation of the load but not the input voltage. The regulator serves three purposes: (1) Provides convenient bias adjustment (0.5 to 9 V) with R10; (2) provides a current sink for fast discharge of C10, for alc shutdown; (3) gives isolation between the bias circuits of each amplifer to the common-base driver. Hence, the adjustment of one amplifier does not affect the bias levels of the remaining amplifiers. D1 was used on the initial test board for protection of the regulator. It should be jumpered out for this application to prevent defeating the fast alc action. The regulator of Fig. 4 switches off in about 120 μs when the voltage from R11 drops from 24 to 2 or less. Considering all of the time constants and delays in the loop, the alc can still react in less than 0.5 ms.

The gates of Q1 and Q2 present an almost pure capacitance, respective to the sources, at 2-30 MHz. To ensure stability, the high input Q is lowered by means of R1, R2, C10 and negative feedback from T2. The source impedance and feedback are controlled by R5, R6 and L5. T2 is wound on a TV-antenna-balun style of core with a bifilar winding. L5 consists of one turn on the same core. The ferrite core should have a μ_e of at least 800 and a high Curie temperature for use down to 1.8 MHz. Teflon-insulated wire is recommended because T2 can reach temperatures in excess of 150° C. T1 and T2 are the common ferrite sleeve/metal tube style of transformers described in note 2. The transformation ratios are 4:1 and 1:4, respectively. A low-loss ferrite core (Stackpole C7/DB) is used for T3. It was chosen to prevent core overheating during extended periods of operation. It is available in a rectangular balun format (no. 55-7051) and has a larger cross-sectional area than the more common no. 57-3238 ferrite sleeves.

Most failures during initial testing occurred from overheating in the output blocking capacitors. Since they are ceramic chip capacitors, soldered rigidly to the pc board, cracks appeared in them. These capacitors must handle an average rf current of 4 to 5 A, but at relatively low voltage. Paralleled disc-ceramic capacitors were tried at C3-C6, inclusive. They worked fine, but were bulky for this board layout. They serve to dc-isolate T3 (unnecesary with this style of transformer) and to compensate for the frequency versus output-impedance slope of the transistors. Part 2 of this paper will appear in a subsequent issue of *QST*.

Notes

[1] "Power MOSFETs versus Bipolar Transistors," *AN-860,* Motorola Semiconductor Products, Inc.

[2] *The Radio Amateur's Handbook,* 59th edition (Newington: ARRL, 1982).

[3] *DC Power Supply Handbook* (Application Note 90A), Hewlett-Packard Co.

[4] "Broadband Transformers and Power Combining Techniques for RF," *AN-749,* Motorola Semiconductor Products, Inc.

SEASON'S GREETINGS FROM THE HAMS AT ARRL/IARU HQ.
(Listed in alphabetical order of call sign)

Joel Kleinman	N1BKE
Richard "Bones" Palm	K1CE
Naoki Akiyama	N1CIX/JH1VRQ
Jeannie DeMaw	W1CKK
Laird Campbell	W1CUT
George Grammer	W1DF
Elizabeth H. Karpiej	KA1DTU
Joan Merritt	KA1DTV
Maureen Thompson	KA1DYZ
Stephen C. Place	WB1EYI
Paul K. Pagel	N1FB
Doug DeMaw	W1FB
Hal Steinman	K1FHN
Marian Anderson	WB1FSB
Marge Tenney	WB1FSN
John Nelson	W1GNC
Bill Webb	WB1GOO
Bob Atkins	KA1GT
Ed Tilton	W1HDQ
Steffie Nelson	KA1IFB
Joan Becker	KA1IFO
Jean Peacor	K1IJV
Cheryl Sowers-Clift	KA1IXI
Andrew Tripp	KA1JGG
Brian Downey	WA1KSF
Dennis Lusis	W1LJ
Stan Horzepa	WA1LOU
Phil Accardi	AJ1N
Peter R. O'Dell	KB1N
Sally H. O'Dell	KB1O
Mike Kaczynski	W1OD
Bruce Kampe	WA1POI
George Woodward	W1RN
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Perry Williams	W1UED
George Collins	KC1V
Arline Bender	WA1VMC
Bill Jennings	K1WJ
Chuck Bender	W1WPR
Bob Halprin	K1XA
John Lindholm	W1XX
Sandy Gerli	AC1Y
Steve Pink	KF1Y
Ellen White	W1YL/4
David Sumner	K1ZZ
Edward C. Raso	WA2FTC
Carol L. Smith	AJ2I
Leo D. Kluger	WB2TRN
Mark J. Wilson	AA2Z
Christopher Imlay	N3AKD
Donald B. Search	W3AZD
W. Dale Clift	WA3NLO
Larry Wolfgang	WA3VIL
William A. Tynan	W3XO
Steve Ewald	WA4CMS
Gerry Hull	AK4L/VE1CER
Paul Rinaldo	W4RI
John Troster	W6ISQ
Wayne Yoshida	KA6KGU
Chuck Chadwick	K8AXL/WB8MOB
Chuck Hutchinson	K8CH
Jim Clary	WB9IHH
Bernard D. Glassmeyer	W9KDR
B. Robert Benson	VE2VW
Harry MacLean	VE3GRO
Maxim Memorial Station	W1AW
ARRL Hq. Station	W1INF

Strays

LISTEN UP

☐ According to Florida law, "No person shall operate a motor vehicle while wearing a headset, headphone, or other listening device, other than a hearing aid. . ." I gather this would include the single earpieces used by some hams to copy above road noise level. Some Florida hams are using this method, and I wonder if they know about the law. [Editor's Note: This law is not unique to Florida. Massachusetts, for example, has a similar law. It's a good idea to check it out with your state's Motor Vehicle Department.] *— Otto Freytag, K4QFM, Riviera Beach, Florida*

I would like to get in touch with . . .

☐ any New York amateurs who are interested in starting a Big Apple Novice net. Tony Sparacio, KA2HJP, 2 Stuyvesant Oval, New York, NY 10009.

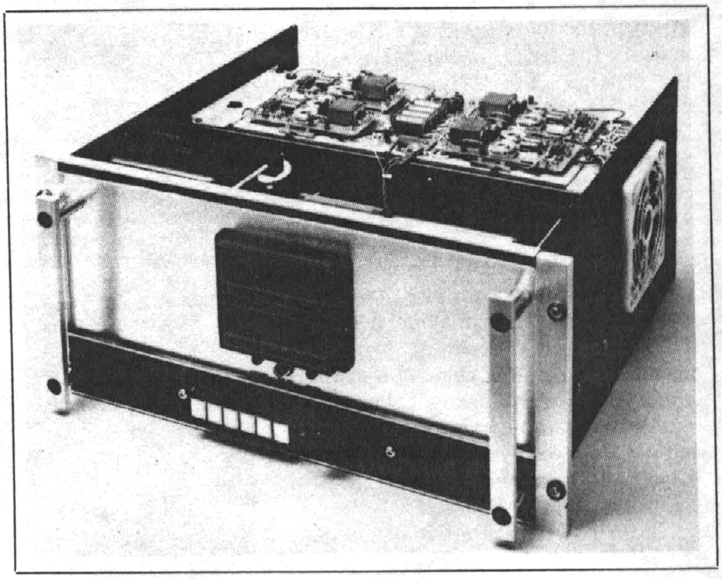
☐ hams who would like to form a national net for lovers of cw QRS for the purpose of formal and/or informal traffic and ragchewing at a speed not to exceed 16 wpm. Gerald Smith, KL7FX/4, P.O. Box 7592, Fort Gordon, GA 30905.

Next Month in QST

To begin the new year, which promises to be another exciting one for Amateur Radio, we'll bring you something old and something new, in the form of

- two well-respected antenna systems, and why they're just as useful in the 1980s as they were in years gone by.
- a spanking new vhf-uhf awards program based on grid squares. Who'll be first to qualify for VUCC?

MOSFET RF Power — An Update



Part 2: The general application information provided in this series offers hard proof that power FETs have graduated beyond infancy. This concluding installment describes filtering methods and other modern amplifier concepts.

By Helge Granberg,* K7ES/OH2ZE

In Part 1 of this article (December 1982 *QST),* the author described a number of design and performance features for his 2-30 MHz, broadband MOSFET linear amplifier. Part 2 provides a wrap-up and offers a variety of design principles that can be applied to other solid-state power amplifiers.

Output Low-Pass Filters

Chebyshev five-pole, constant-k low-pass filters are used in the amplifier output (Fig. 5). Shunt elements have been added to provide an elliptic function. These elements usually are designed to resonate with the filter inductances at the harmonic or other undesired frequencies. This provides notches that give much higher close-band attenuation than can be obtained from a simple five-pole filter. The filter skirt is also much steeper, but at the cost of degraded far-band attenuation. However, in an application such as this in which the filters are used at a multiplicity of frequencies, they can't be optimized in this respect. The worst case (−35 dB) for the third harmonic (which is of the most concern) is with the low-frequency filters. This probably is caused by the low Q of the inductors. MHz and below is approximately 50 dB or better, which meets the FCC requirement of 50 mW maximum for spurious emissions. The 1.8 to 30 MHz range can't be covered by the six filters without resultant gaps; a minimum of eight filters with sharper cutoff characteristics would be required for continuous coverage.

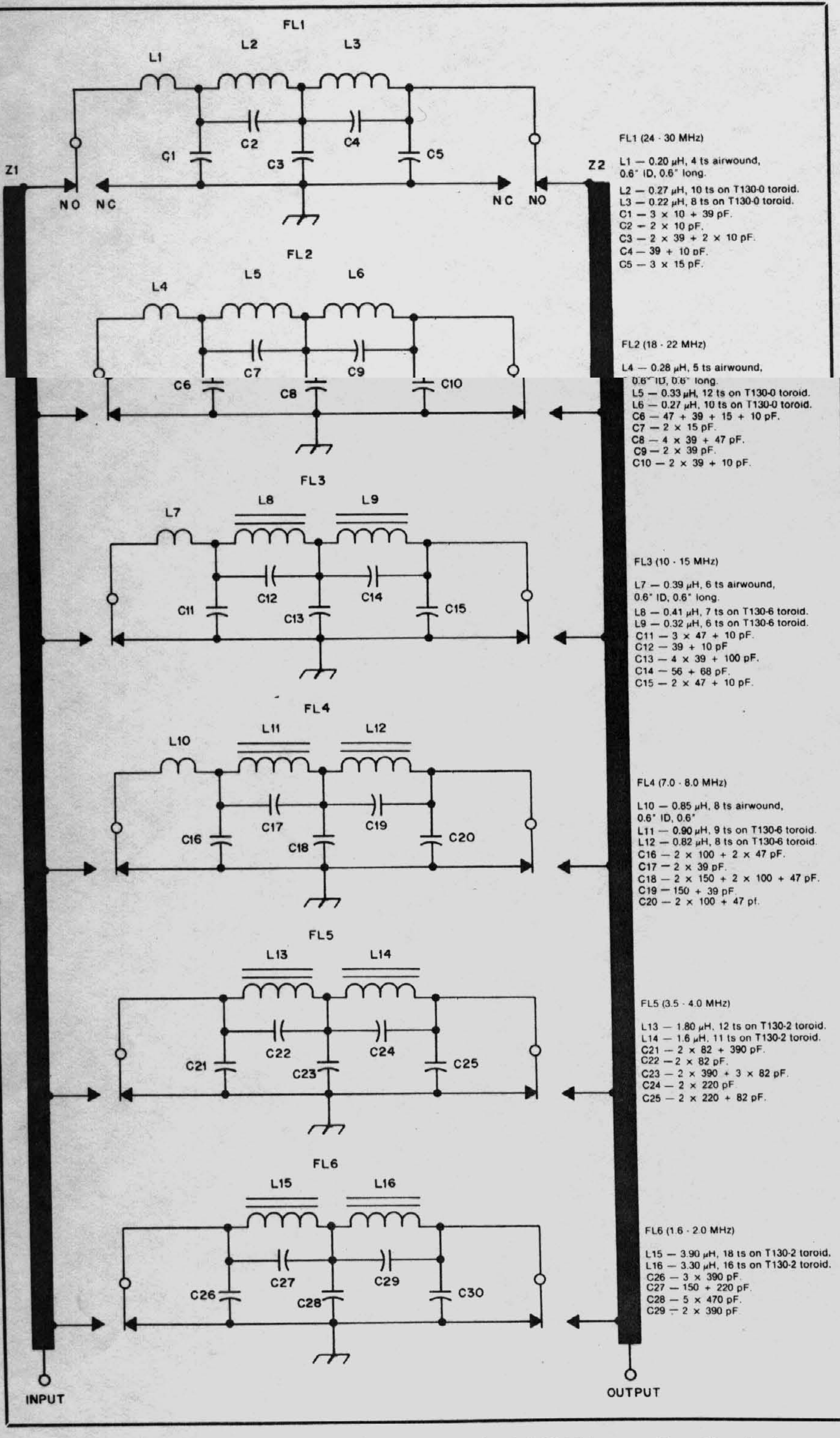
Filter Construction

Each filter is assembled on a separate pc board of similar layout. Only the component values are different. The most critical parts are the capacitors — often a stumbling block for practical high-power filter design. At these power levels, they must withstand peak rf voltages up to 800 (even higher in the event of a mismatch). The rf-voltage rating of ceramic capacitors is only some 30% of the nominal dc value. Hence, capacitors with ratings of 2000 to 3000 V are required. Also, they must be able to handle rf currents up to several amperes. These filters contain several inexpensive disc-ceramic capacitors in parallel. This provides current capacity and permits capacitor combinations that yield nonstandard values, as needed.

The distance from each filter input or filter.[5] Sections of 50-ohm strip line (Z1, Z2) are used to connect the T-R switch and the power amplifier to the filters. These are made of copper-clad pc-board material that is attached to the chassis below the filter boards. The lowest-frequency filter is located nearest to the input and output ends of the line, while the highest-frequency filter is at the far end. Thus, when the filters are switched, the remaining piece of unterminated line causes minimum VSWR in each case. Although the filters were tested individually with the amplifier, L1, L4, L7 and L10 had to be added to restore the filter characteristics in the composite circuit. This was necessary because the mechanical arrangement and the added capacitance from attaching the relay contacts to the 50-ohm lines affected the filters.

Filter-Relay Driver

Each relay field coil is driven by the circuit of Fig. 7. It is basically a programmable BCD-to-decimal decoder. The gates were added to accommodate the manual-switching feature. When the BCD input plug is disconnected, the manual switch is activated automatically.



FL1 (24 - 30 MHz)

L1 — 0.20 μH, 4 ts airwound, 0.6" ID, 0.6" long.
L2 — 0.27 μH, 10 ts on T130-0 toroid.
L3 — 0.22 μH, 8 ts on T130-0 toroid.
C1 — 3 × 10 + 39 pF.
C2 — 2 × 10 pF.
C3 — 2 × 39 + 2 × 10 pF.
C4 — 39 + 10 pF.
C5 — 3 × 15 pF.

FL2 (18 - 22 MHz)

L4 — 0.28 μH, 5 ts airwound, 0.6" ID, 0.6" long.
L5 — 0.33 μH, 12 ts on T130-0 toroid.
L6 — 0.27 μH, 10 ts on T130-0 toroid.
C6 — 47 + 39 + 15 + 10 pF.
C7 — 2 × 15 pF.
C8 — 4 × 39 + 47 pF.
C9 — 2 × 39 pF.
C10 — 2 × 39 + 10 pF.

FL3 (10 - 15 MHz)

L7 — 0.39 μH, 6 ts airwound, 0.6" ID, 0.6" long.
L8 — 0.41 μH, 7 ts on T130-6 toroid.
L9 — 0.32 μH, 6 ts on T130-6 toroid.
C11 — 3 × 47 + 10 pF.
C12 — 39 + 10 pF
C13 — 4 × 39 + 100 pF.
C14 — 56 + 68 pF.
C15 — 2 × 47 + 10 pF.

FL4 (7.0 - 8.0 MHz)

L10 — 0.85 μH, 8 ts airwound, 0.6" ID, 0.6"
L11 — 0.90 μH, 9 ts on T130-6 toroid.
L12 — 0.82 μH, 8 ts on T130-6 toroid.
C16 — 2 × 100 + 2 × 47 pF.
C17 — 2 × 39 pF.
C18 — 2 × 150 + 2 × 100 + 47 pF.
C19 — 150 + 39 pF.
C20 — 2 × 100 + 47 pf.

FL5 (3.5 - 4.0 MHz)

L13 — 1.80 μH, 12 ts on T130-2 toroid.
L14 — 1.6 μH, 11 ts on T130-2 toroid.
C21 — 2 × 82 + 390 pF.
C22 — 2 × 82 pF.
C23 — 2 × 390 + 3 × 82 pF.
C24 — 2 × 220 pF
C25 — 2 × 220 + 82 pF.

FL6 (1.6 - 2.0 MHz)

L15 — 3.90 μH, 18 ts on T130-2 toroid.
L16 — 3.30 μH, 16 ts on T130-2 toroid.
C26 — 3 × 390 pF.
C27 — 150 + 220 pF.
C28 — 5 × 470 pF.
C29 — 2 × 390 pF

Fig. 5 — Low-pass filter circuit and component values for 1.6-30 MHz operation. Standard capacitor values have been placed in parallel to obtain nonstandard values. Toroid cores are Amidon, Palomar or Micrometals Corp. powdered-iron units. The designator "×" means "times," E.g., 2 × 47 + 10 pF means two 47- and two 10-pF capacitors, all in parallel to provide 114 pF. Wire type for all inductors is no. 14 enameled. Capacitors are RMC 3-kV disc ceramic, except those units that are 390 pF and higher, which are RMC 2-kV units. Relays are Deltrol 20693-83/405 with 12-V coils.

are used because they can be operated directly from a + 12-V supply. Also, they have internal clamping diodes for suppressing inductive spikes.

T-R Relay and Timing Circuit

Fig. 8 shows the T-R relay driver and timing circuit. The input-output timing of the T-R relays must be precise. If the output relay is switched with full rf power, the relay lifetime will be very limited. Thus, the output relay must be switched on before the input relay is. Similarly, the output relay has to be switched off after the input one is deactivated. Simple R-C and diode networks are used to generate the delays.

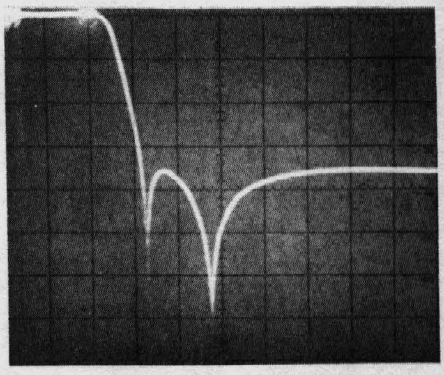


Fig. 6 — Typical response curve of an elliptic-function low-pass filter with two resonant shunt elements.

A CMOS hex inverter performs the amplifier pulse-shaping functions. The upper channel, which controls the input relay, delays the leading edge of the keying pluse and shortens it by a few milliseconds. The lower channel controls the output relay, and delays the trailing edge of the keying pulse, which is lengthened from the original as shown in the timing diagram. Both delays are adjustable from 2 to 12 ms by means of R2 and R4.

Other Circuitry

The circuit in Fig. 9 performs the following functions:

1) Output VSWR detection.
2) Linear alc control.
3) Alc shut down.
4) Bias switching between standby and operate.
5) Bias temperature tracking.

The VSWR reflectometer senses the reflected power caused by load changes (50 ohms nominal).[6,7] Capacitor C is approximately 1 pF in value and must be capable of handling high rf voltages. It can be made from a piece of coaxial cable, such as RG-58/U, with the braid stripped to a length of 0.4 inch. The section of center conductor can be used as the primary of T1 and for connections to the amplifier output and relay terminal.

The rf meter (normally seen in place of D4) is replaced with a diode for rectifying the rf energy. The resultant voltage is filtered by means of C6 and R31. The level of 0.2-0.3 V for a 50-ohm load. It increases to about 1.5 V for 3:1 VSWR. This voltage is used for the linear alc function and is fed to the alc amplifier, an LM307. Controls R26 and R30 can be set, for example, so that an output VSWR of 2 will have no effect, but a 3:1 condition will reduce the power output by 3 dB or more. The alc shutdown operates also from the dc developed in the VSWR bridge. During a complete mismatch, such as an open load, the instantaneous voltage is greater than 10, thereby saturating the

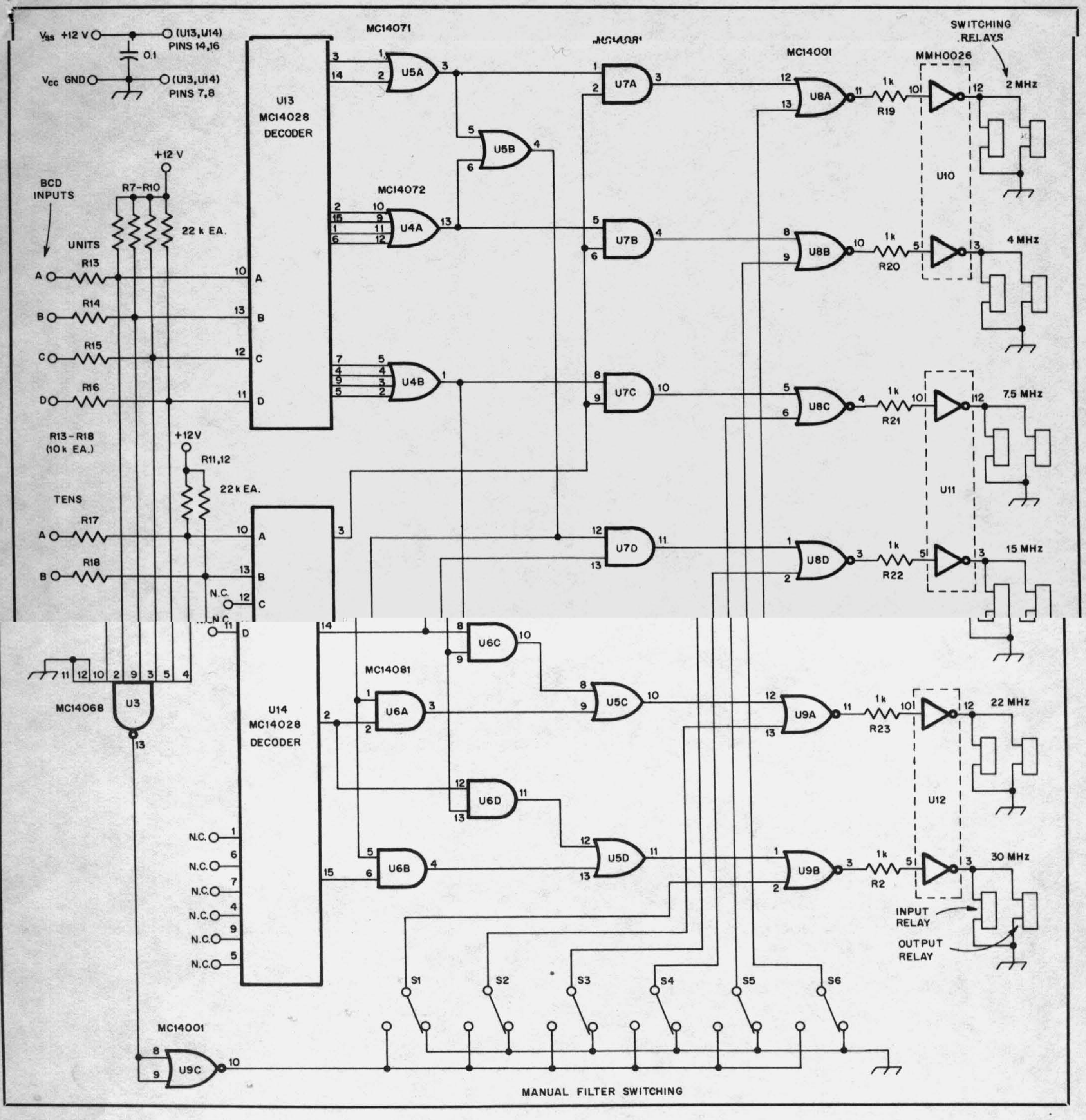
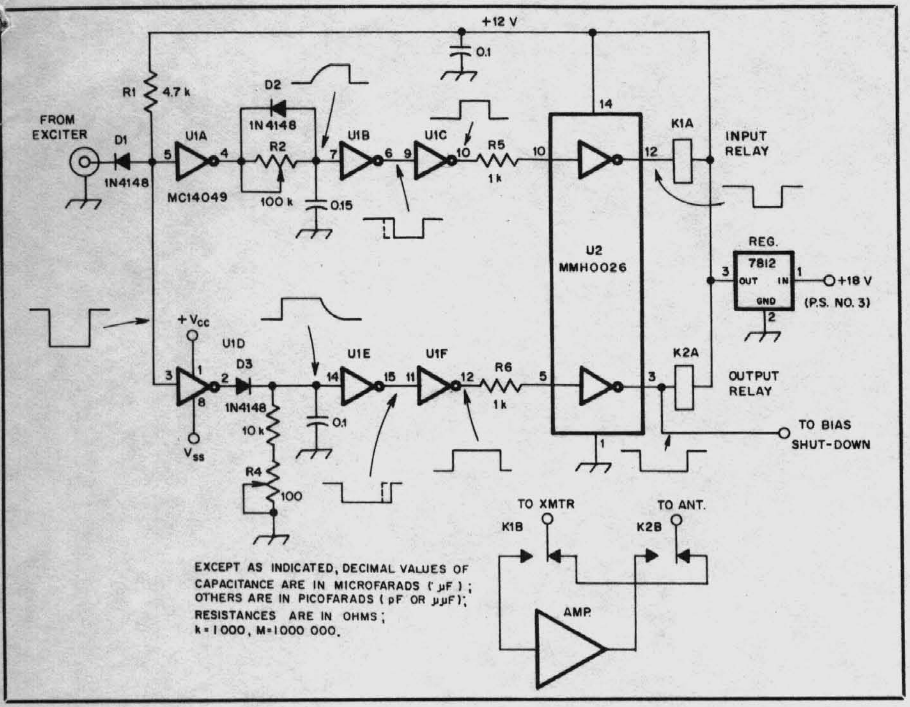


Fig. 7 — Output filter select and drive circuit indicator lights are connected in parallel with the relay coils, which are at the far right of the diagram.

LM307. After the alc loop has reacted and the power output has been reduced by a factor of 10, the voltage settles to 5 or 6 and the alc amplifier remains saturated. This holds the output power at the reduced level.

Automatic bias switching is initiated when a ground to +12-V signal is brought from the T-R relay driver to the same (−) input of the LM307. This overrides the control signal coming from the VSWR sensor and, when positive, turns the bias completely off. D5 makes this independent of the alc function.

Each of the rf amplifiers (Fig. 4, Part 1) and their associated circuits draw approximately 15 mA of bias current. The combined current is 120 mA, which is too much for the alc amplifier to handle. Therefore, a TIP31 has been added for use as a buffer/driver. Its input voltage is also controlled by a thermistor (R34), which is connected thermally to one of the main heat sinks. R34 controls the idling current of the power FETs by lowering the bias voltage during periods of elevated temperature, and vice versa. With bipolar transistors, this is normally done with forward-biased diodes, in which the diode voltage drop versus temperature closely follows that of the base-emitter junction. Both types of transistors have a positive temperature coefficient (for a constant bias current or voltxge, the idling current increases with temperature). Although the g_m (transconductance) of a MOSFET



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000.

Fig. 8 — T-R switch drive and timing circuit. The approximate wave forms for the input and output functions are shown near the related ICs. Fig. 9 provides details of the relay connections.

decreases under these conditions the gate threshold voltage goes down, and usually has a dominant effect. There is a reversing point at higher drain currents. For the MRF150, this is around 5 A. The exact value depends on the g_m and the initial temperature. This is why power MOSFETs do not go into thermal runaway, provided the dissipation ratings are not exceeded.[8]

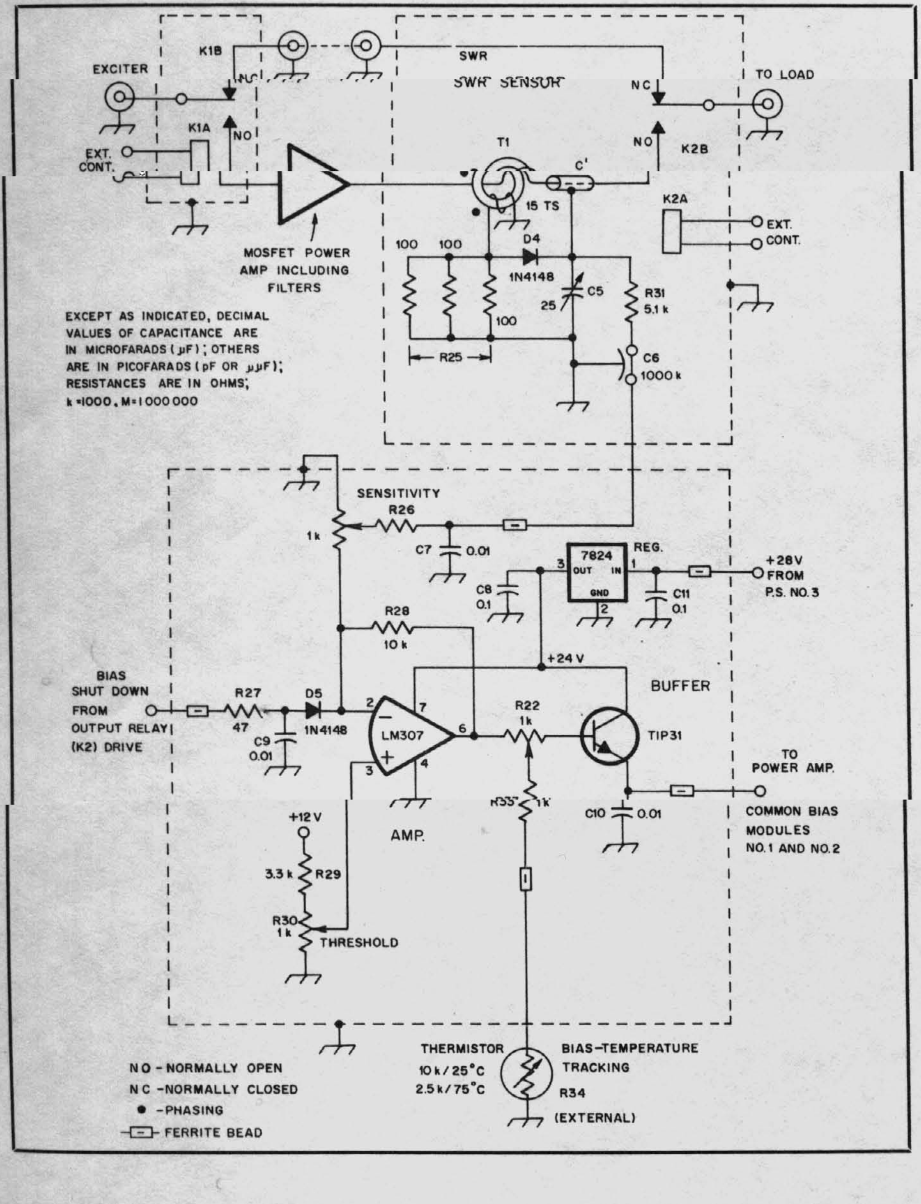
Conclusion

Although solid-state, high-power rf amplifiers are not in competition with vacuum-tube units, an increasing number of them appear on the commercial market regularly. Their feasibility versus output power is a question to be answered, at least until high-voltage rf transistors are available.

Power FETs are high-voltage, low-current devices, and some switching transistors for use up to several hundred volts are now on the market. In rf applications, however, several problems exist. Because of the internal structure of a transistor, rf arcing occurs (internally) easier at high voltages because the impedance levels are high. Transistor packaging techniques must be improved for this reason, and also for thermal considerations. The vertical-channel power FET (VFET) technology is fairly new and will certainly undergo many advances in the years to come.

This amplifier has been field-tested since late 1981, at the author's home. Switching from band to band is an absolute delight. Since no tuning is required, it should be a contest-operator's dream! Thus far, the only failures have been a jammed thermal switch that operates the fans, plus a burned low-pass filter board, caused by a loose piece of solder.

If you haven't worked with power FETs thus far, perhaps it's time you became involved, but maybe on not so grand a scale as is represented by this paper. Certainly, there are many advantages in the use of power FETs over bipolar transistors. QST



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000

Fig. 9 — VSWR sensor, alc and bias circuits. Note ferrite-bead chokes in all alc input and output lines. They help prevent rf energy from getting into the alc amplifier.

Notes

[5] mm = in. × 25.4

[6] *The Radio Amateur's Handbook*, 59th edition (Newington: ARRL, 1982).

[7] W. Orr, *Radio Handbook*, 18th edition (Indianapolis: Howard W. Sams & Co.).

[8] MRF150 data sheet, plus appropriate linear IC and CMOS data sheets.