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COLUMNS

ABOUT THE COVER

Reception of VISSR signals—the highest-resolution weather-satellite images presently available—requires stable, high-gain, low-noise preamplifiers. Until recently, such preamps were available only commercially, at high cost. Now you can build your own VISSR low noise amplifier. See the article beginning on page 3.

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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.

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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*.

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Empirically Speaking . . .

STA Extension for SKIPNET Autoforwarding Network

On January 5, 1989, Executive Vice President David Sumner filed a letter with the Federal Communications Commission seeking an indefinite extension of the special temporary authority to allow automatic control of certain packet stations on HF. The letter stated that the League is now working on language for a permanent rule change that would permit this type of operation. The ARRL Committee on Amateur Radio Digital Communication has studied the technical and operational performance of SKIPNET while operating under the STA and concluded that the following lessons were learned:

•The system works, moves traffic and, with careful frequency selection, can provide a public service without undue interference to other amateur activities.

•Network management and control are necessary.

•Accountability for traffic must be with the station introducing it into the network; accountability at relay points is not practicable.

•Packet is not compatible with other modes and needs separate frequencies; carrier sense is not adequate to protect against interfering with other modes on HF owing to transmission impairments, hidden station effects, etc.

•Frequency stability needs to be on the order of 10 Hz.

•Protocols need improvement, and new capabilities are needed.

•Modems need improvement.

•Watchdog timers (to disable the transmitter automatically in the event of malfunction) are essential.

•Stations need to change frequencies in accordance with propagation conditions to improve efficiency, reduce retries, and free up frequencies for other users.

•While a 200-watt power output has proven adequate for many domestic paths, there is no justification for a blanket 200-watt limitation.

The above lessons will be considered in drafting new language by the appropriate ARRL Board committees. Upon completion of this process, the plan is to petition the FCC to update the rules regarding packet-radio operation.

Same-day telephone approval of continued operation of the STA came from FCC Private Radio Bureau Chief Ralph Haller. However, he said that the term of the STA extension cannot be indefinite, but would be for one year.

Meanwhile, in the trenches, SKIPNET stations continue to move the mail within North America and to/from countries with which the US and Canada have third-party-traffic agreements. This is a day-in-day-out effort on the part of PBBS sysops who have dedicated hundreds of dollars of their own resources to provide this public service. Dr. David Toth, VE3GYQ, is serving as the overall informal manager of North American SKIPNET operations. Special thanks go to Dave and others who are helping to coordinate SKIPNET activities. US stations now authorized to operate under the STA are:

Edward F. Adams, N6YN; Kenneth F. Araujo, KA1ZT; John A. Bennett, N4XI; William G. Bertrand, AA4TM; Danial R. Bollander, WB9TYT; David W. Borden, K8MMO; Robert Boyink, KJ8C; Kenneth D. Brodeur, KK4WR; John J. Burke, KB6GOZ; Jim L. Cantrell, KI4FL; Gregory A. Cerny, NØBEV; David H. Cheek, WA5MWD; Brian C. Churchill, N1BBT; Dr. Thomas A. Clark, W3IWI; Clement C. Constant Jr., W9CD; James R. Curran, KA4OJN; Eugene W. Danials, KD4EQ; Coy C. Day, N5OK; Joseph P. Demers, WA1WLV; Ken Dresser, KK4L; Samuel W. Drinkard, WA4PHY; Richard C. Duncan, WD5B; Lee Dusbabek, WB6KAJ; John J. Eigenbrode, KD6SQ; Dennis N. Ely, W5XO; Carl A. Estey, WAØCQG; Patrick J. Fagan Sr., WA5DVV; Frederick S. Fahrner, WB6IKS; Joseph A. Fisher, KC2TN; David C. Fitz, KIØQ; David A. C. Freeborn, KCØQJ; Joseph E. Galipeau Jr., WA1LRL; Ronnie L. Gamel, N5WX; David Garnier, WB9OWN; Karl H. Geng, N1DL; Kenneth L. Harwood, WA5QZI; Jack A. Hearn, KH6WY; Thomas J. Hogan, WB7DCH; James M. Homan Jr., W4DPH; Scott D. Hudler, KB3X; Theodore A. Huf, K4NTA; Harry W. Jackson, WD4PPF; Lew Jenkins, N6VV; Kenneth D. Johnson, N1CWP; John R. Jones, WB8CQV; Joseph P. Lagermasini, AG3F; C. A. Lee Jr., WA4EWV; Gregg R. Lengling, WD9DHI; Byron D. Lichtenwalner, W1HAB; John Lind, KD7XG; Shelton B. McAnelly, KD5SL; Perry D. McLean, WØXK; Donald J. S. Merten, K2AAA; William R. Michalson, AA2S; Gary A. Mitchell, WB9TPG; John Moran, W1ZLG; Wesley F. Morris, K7PYK; Terrance M. Neal, AA6TN; Donald W. O'Neil, WA4SZK; Henry N. Oredson, WØRLI; Vincent C. Ott, WD4NUN; James D. Page, WA7ARI; Regg K. Patterson, KR5S; Michael Payne, WØLVJ; Eugene R. Poole, AJ6F; Alva P. Ramsey Jr., WA4VMV; Herbert H. Salls, WB1DSW; James E. Scalf Jr., K4TKU; Fred R. Scalf, K4EID; Robert A. Schiers Jr., NØAN; Joseph Schiminel, W2HPM; William A. Schroeder, W9ZBD; Gary Sharp, WDØHEB; Dave L. Shavey, KØHOA; John C. Shew, N4QQ; John L. Sielke, N4JS/2; Donald Simon, NI6A; John T. Smith, KI4XO; Arthur R. Sprague, KH6GPI; Jonathan A. Starr, AH6GJ; Norman J. Sternberg, W2JUP; Donald R. Stiver, N6EEG; Robert D. Straughn, WN4IIV; Joesph T. Subich, AD8I; James N. Tanis, NXØR; Robert Thanisch, KN5D; Ronald M. Thomas, N4RT; Lynn Tory, NØHMF; John D. Vester, KE7CZ; David N. Wade, KC7CG; Charles P. Walker, NA2B; Bill O. Walton Jr., KJ6EO; Dennis Watters, WBØTAX; Aubrey E. Whitcher, WY5J; Harvey L. Williams, KK4CQ; Neal S. Wood Jr., KØKBY; Edward B. Wright, WØLJF; David L. Zeph, W9ZRX

—W4RI

A Low-Noise Preamp For Weather Satellite VISSR Reception

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Geostationary weather satellites, from which radio amateurs have long delighted in recovering earth images, actually provide two distinct products. WEFAX, or weather facsimile, is the more familiar of the two, and consists of preprocessed and enhanced frames, taken at visible or infrared wavelengths, relayed via a frequency modulated microwave carrier, in slow-scan format (four lines per second). Because of its relatively narrow bandwidth (30 kHz per channel) and respectable output power (on the order of five watts), the WEFAX signal is commonly received with little effort using relatively simple equipment.[1,2] But the data is *secondhand*. WEFAX images are analog retransmissions of raw satellite data that has been downlinked to a central data-collection facility for processing,[3] then returned to the satellite for distribution. In WEFAX mode, the satellite is thus serving as a repeater.

In its other operating mode, the familiar WEFAX satellite provides a wide-band digital signal variously called VISSR (for Visible and Infrared Spin-Scan Radiometer, the image sensor on the satellites), VAS (for VISSR Atmospheric Sounder), or HRPT (for High-Resolution Picture Transmission, which highlights the signal's advantage over WEFAX). Throughout this article, I shall use the term VISSR.

The ultimate microwave challenge is reception and display of raw satellite data prior to processing. This has—until recently—been an elusive goal, because in the digital mode, the satellite is transmitting at greater than 2 Mbits per second. Hence, a wide receiver bandwidth (on the order of 8 MHz) is required. Thus, though the VISSR transmitter power is on a par with that of the WEFAX transmission, the modulation sidebands are spread over a frequency spectrum about 300 times as wide. The resulting low spectral density makes it necessary to employ some rather sophisticated receiving equipment; government stations employ 60-foot dishes, parametric amplifiers and large mainframe computers for image processing.

[1]Notes appear on page 6.

Nevertheless, a few enterprising radio amateurs have succeeded in building homemade VISSR receiving stations,[4] defying the odds as did those first few homebuilders who recovered TV pictures from domestic communications satellites a decade earlier.[5,6] In truth, to date they have been successful in recovering only "stretched" VISSR data, which is the more narrowband of the two available digital weather-satellite services. For this, the Government uses only a 24-foot dish! For stretched VISSR and TVRO reception alike, success depends upon the development of low-noise, high-gain, stable receive preamplifiers, which is the subject of this article. The techniques presented here can, of course, be applied equally well to other services and other frequencies.

Performance Requirements

Actually, the satellite TV analogy is apt in that the spectral densities of the two transmissions are roughly equivalent. So, the antenna and preamp requirements for VISSR and TVRO reception should be about the same. The frequencies differ, of course, with TVRO operating near 4 GHz, and VISSR around 1.7 GHz. For a given parabolic-antenna diameter, gain varies inversely with the square of wavelength.[7] Therefore, you can expect a given TVRO dish to produce about 7 dB less gain in VISSR service. But, it's a lucky coincidence that free-space path loss varies *directly* with frequency, at precisely the same rate. Which means, for a given antenna, the two effects exactly cancel! In other words, if receiver noise figure is equivalent between the two services, a TVRO dish will perform just as well for VISSR as it did for satellite TV.

What does that leave us with for required preamp performance? Successful TVRO installations typically require a low-noise amplifier (LNA) with noise temperature on the order of 100 Kelvins (a 1.3-dB noise figure), and enough gain to overcome feed-line losses and mixer noise (typically 40 dB). These figures give us our design objectives for the VISSR LNA: 100-Kelvin noise temperature, with about 40 dB of gain, should suffice. Such a preamp will not only do a credible job receiving stretched VISSR on a 12-foot-diameter TVRO dish, but provides spectacular standard WEFAX reception using a dish made from a 2-foot-diameter snow sled!

LNA Topology

The active device of choice to establish the required low-noise performance is obviously the gallium-arsenide field-effect transistor, or GaAsFET. But, stability considerations dictate limiting preamplifier gain to about 15 dB per stage,[8] suggesting that this would have to be a three-stage preamplifier. Unfortunately, my success rate in producing stable and reliable three-stage GaAsFET preamps leaves much to be desired! The problem is that cascaded GaAsFETs are hard to match, squirrelly to tune, exhibit poor input SWR when tuned for minimum noise figure, are high-Q and narrow-band devices by nature, and love to oscillate! I'm sure I'm not the only one who's had those experiences.

Bipolar monolithic microwave integrated circuits (MMICs), on the other hand, are the most docile of devices. They have moderately low noise figures and acceptably high gain, are wideband by design, have 50-ohm inputs and outputs, are unconditionally stable for any combination of source and load impedances, and cost less than the discrete components needed to duplicate their function. I have used them successfully as the input stage of WEFAX receivers; unfortunately, they're about 2 dB too noisy for VISSR front ends.

A likely compromise, it would seem, is to cascade a single GaAsFET stage (to establish the required noise figure) with a couple of MMIC stages (to establish system gain). The result looks like the circuit shown in Fig 1, with the gain and noise data representative of the prototype I ultimately built. The overall performance of the cascade preamplifier, as calculated by Teledyne's computer program, RF Toolbox,[9] is shown in Table 1.

Designing the GaAsFET First Stage

I selected the Avantek ATF-10235 low-noise gallium-arsenide FET for the input stage, for a number of reasons. At 2 GHz (the lowest frequency at which the data sheet fully characterizes the device, but

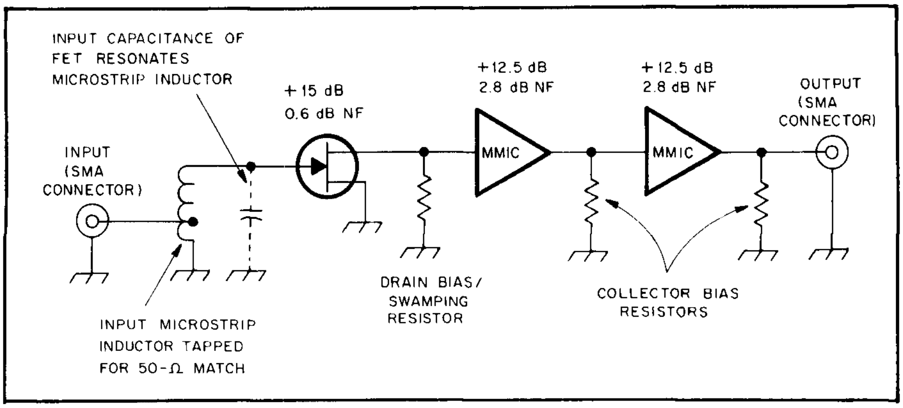


Fig 1—Cascaded GaAsFET and bipolar MMIC stages provide high-gain, low-noise performance in the 1680- to 1700-MHz weather satellite band. This RF equivalent circuit doesn't show the dc-bias components and bypass and coupling capacitors for simplicity of analysis.

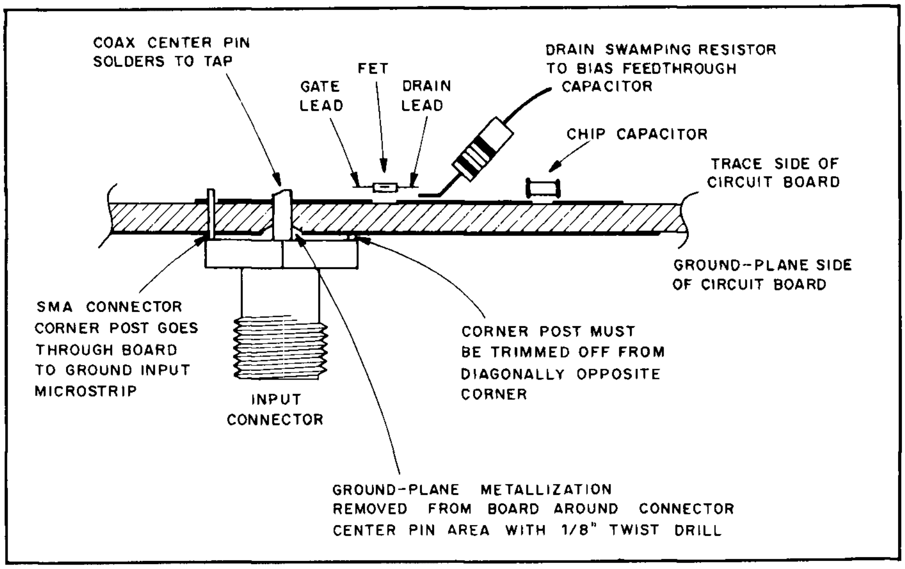


Fig 2—Detail of input microstrip grounding, tap and FET mounting.

near enough to the desired operating frequency to be useful), typical noise figure is a claimed 0.6 dB, with an associated gain of more than 15 dB. The recommended low-noise bias point of 2 V V_S and 20 mA I_{DS} affords an output 1-dB compression point near +10 dBm, hence, a wide spurious-free dynamic range can be expected. The cost of the device is low (on the order of $12). And, most important, Γ_o, the desired source-reflection coefficient for optimum noise figure, is very nearly the *exact complex conjugate* of S_{11}, the input voltage reflection coefficient, over a wide range of frequencies! This means that, unlike many other GaAsFETS, for the 10235, input conjugate match and optimum noise match will very nearly coincide.

A happy coincidence allows a rather simplistic approach to designing the input circuit. Scattering-parameter analysis[10] indicates that the input impedance at the desired operating frequency is near 200 ohms, and is somewhat capacitive. An inductor of suitable reactance shunting the gate lead will resonate the capacitive input component, and attaching the input coax connector to a tap halfway up the inductor affords a reasonable match of the resulting 200-ohm resistive component to 50 ohms. (Remember, an autotransformer with a 2:1 turns ratio affords a 4:1 impedance ratio). This inductor also provides a dc return to ground for the gate, enabling source self-biasing to be used, and is implemented in a 50-ohm microstrip, conveniently shorted to ground at the "bottom" by the corner post of the selected coaxial input connector, as illustrated in Fig 2.

If the input-matching scheme seems crude, the output circuit is even more so. Scattering parameters indicate the output impedance of the FET is also near 200 ohms; shunting the output with a 68-ohm resistor (which happens to double as the drain bias resistor) just happens to bring the output impedance rather near 50 ohms. This resistive swamping of the output also serves to stabilize an otherwise unstable active device. There is, of course, a gain penalty in resistive swamping, but this FET has more gain than we need in the input stage, anyway.

The input FET is a common-source stage, hence, the two source leads need to be at RF ground. To enable source self-biasing, they will be run to ground through bypass capacitors. To keep Q high and losses to a minimum, resonant (quarter-wave) capacitive stubs are used, as seen in the photographs and PC artwork. The resistors from these stubs to ground fix the FET's quiescent current by biasing the gate halfway to pinch-off.

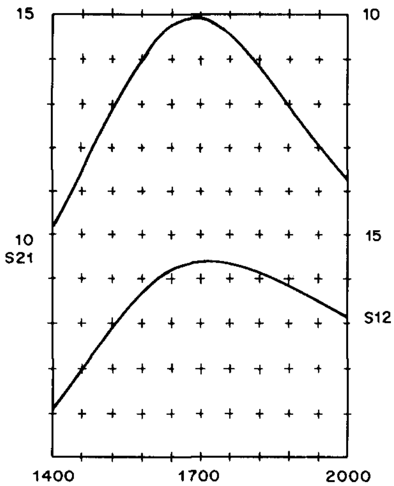
Computer analysis of the proposed input stage was performed using Randall Rhea's program, SuperStar.[11] Table 2 is the circuit file; expected performance is shown in Table 3. Note that gain peaks near 15 dB at 1680 MHz, that input SWR is nearly perfect, output SWR is a rather poor 3:1, and the device is only marginally stable, as indicated by the Rollet Stability Factor (K) hovering around 1. The relatively high output SWR is the result of totally ignoring the drain-circuit reactance when "matching" the output, and gain drops a few tenths of a decibel.

Swept response is shown graphically in Fig 3. Input (Fig 4) and output (Fig 5) stability circles confirm what K told us in the data table: The stage is only conditionally stable, with the instability regions just grazing the outer edge of the Smith Charts in both cases.

Selecting the MMIC Output Stages

The two MMIC stages following the GaAsFET must do far more than simply add gain to the preamp. They must so terminate the first stage as to render it unconditionally stable regardless of input match. In addition, the noise contribution of these stages must be negligible. As a rule of thumb, the first stage of a cascade establishes overall noise performance only if its gain exceeds the noise figure of the following stages by about 10 dB.[12] Because the GaAsFET stage is giving us a gain of 15 dB, this means the noise figure of the stages that follow must be under 5 dB. The selected MMICs more than meet this requirement.

The use of MMICs as gain blocks is well covered in the Mini-Circuits MAR guide,[13], and this design is based on considerations outlined therein. Because the selected MMIC has input and output impedances rather close to 50 ohms at the frequency of interest, no attempt is



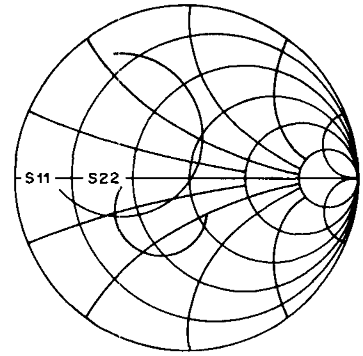


Fig 3—Swept response of the proposed input stage.

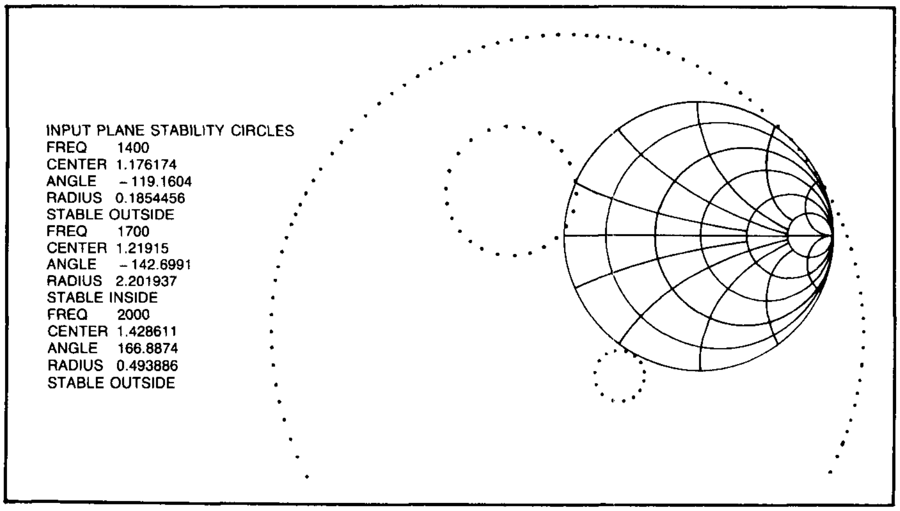


Fig 4—Input-plane stability circles for the proposed input stage.

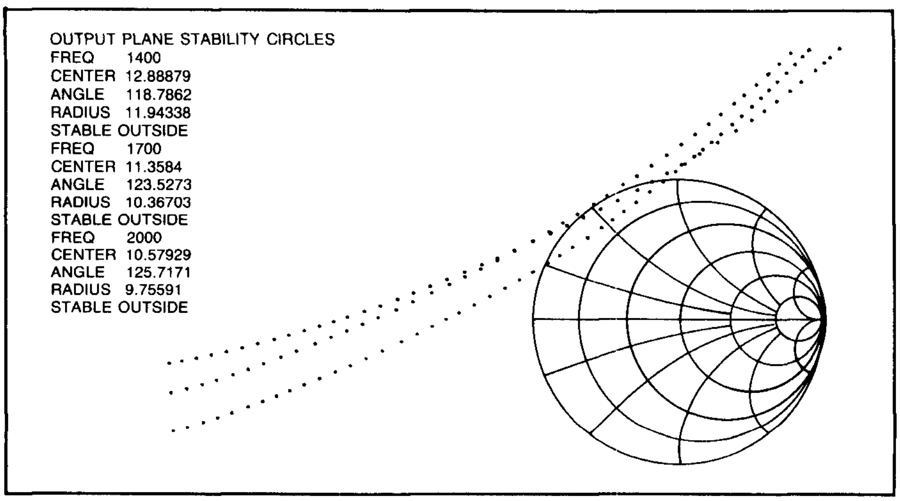


Fig 5—Output-plane stability circles for the proposed input stage.

made to provide further matching. The observed gain correlates well with the device data sheets.

Optimizing the Cascade

The true power of computer-aided design rests in the ability to do repetitive "what if" analyses. By adding the two MMICs (and their shunt collector bias resistors) to the SuperStar data file, it's possible to discern the effect of the latter stages on the input FET, and optimize as required. The expanded circuit file, seen here as Table 4, shows the length of the input microstripline, and the value of the drain bias resistor, preceded by a question mark [?]. This indicates that these values are available for manipulation when "tuning" the circuit in computer analysis, and in fact the final values (selected for optimized performance of the overall amplifier) are slightly different from those shown in Table 2.

The performance achieved after computer optimization is shown in Table 5, and graphically in Fig 6. Note that at the operating frequency, the overall amplifier has a gain of 40 dB, the input and output SWR is under 2:1, reverse isolation is on

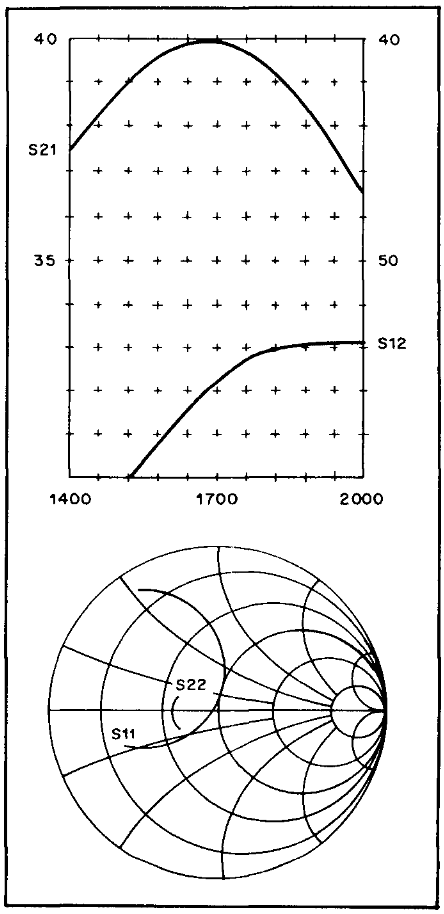


Fig 6—Swept response of the computer-optimized input stage (see Table 5).

the order of 56 dB, and the circuit is now unconditionally stable (the Rollet Stability Factor, K, is well over unity). The input (Fig 7) and output (Fig 8) instability regions now fall far off the edges of the Smith Chart, confirming that this amplifier will be stable for any combination of real source and load terminations.

Construction, Tune-Up and Test

The circuit schematic is shown in Fig 9 along with the parts list. Assembly details parallel my previous microstrip preamplifiers.[14,15] If the design approximations described earlier seem crude, the construction techniques I employed are even more so. This amplifier was built on a substrate of 1/16-inch fiberglass-epoxy PC-board stock, double clad, with one ounce of copper per square foot per side. PC-board artwork is shown in Fig 10, but I fabricated the prototype with what AMSAT stalwart Gordon Hardman calls an Approxo Knife—there's nothing exact about it! Actually, four straight cuts (two in parallel, spaced 0.1 inch apart for the microstrip; two in an X to form the bypass capacitors) plus a bit of "peel," do an amazing job of defining this amplifier! A view of each side of the assembled amplifier is shown in Fig 11. A parts-placement diagram is shown in Fig 12.

Tune-up? There is none! Just drop the parts (carefully) into the PC board, solder sparingly using a minimum amount of heat, test for proper dc bias with both ports terminated in 50-ohm loads, and you're done. Don't forget to observe proper antistatic precautions when working with this, or any, GaAs device.

The noise figure of the completed amplifier measures about 1 dB; nonideal components and substrate losses account for the excess noise. I say "about" because, at these low noise levels, measurement errors are on a par with the measured value itself. Fig 13 shows the swept forward gain of this preamplifier, as measured on a microwave network analyzer. Compare it to the predicted gain in Fig 6. Amazed? So was I!

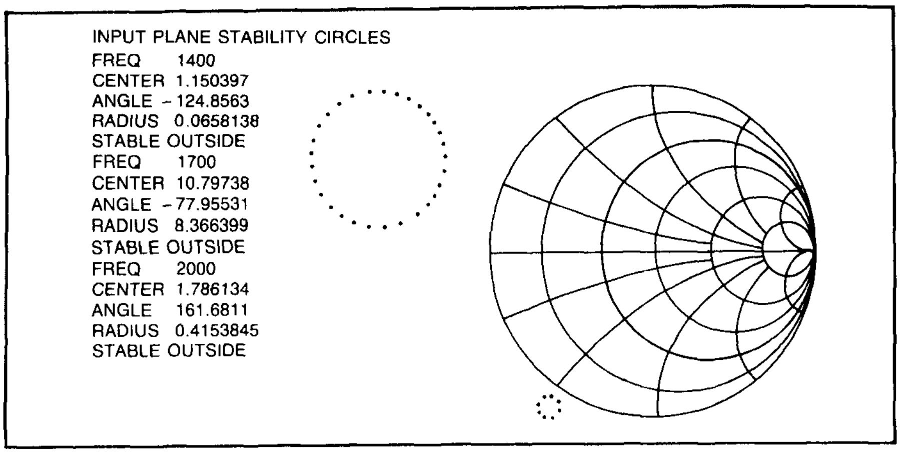


Fig 7—Input-plane stability circles for the computer-optimized input stage.

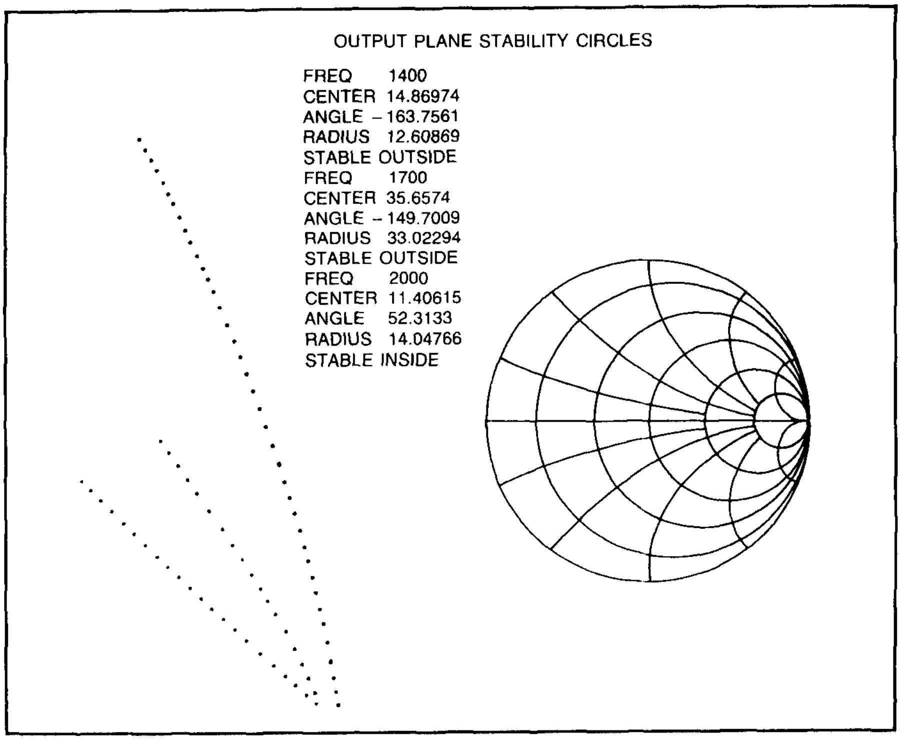


Fig 8—Output-plane stability circles for the computer-optimized input stage.

Notes

[1]H. P. Shuch, "A Cost-Effective Modular Downconverter for S-Band WEFAX Reception", *IEEE Transactions on Microwave Theory and Techniques*, Dec 1977, p 1127.

[2]H. P. Shuch, "A Weather Facsimile Display Board for the IBM PC", *QEX*, Sep 88, pp 3-7 and 15.

[3]For the GOES series of satellites operated by the US, this facility is located at Wallops Island, VA.

[4]J. L. DuBois, "A Low Cost GOES VAS Imaging System for PC/XT/AT Host Systems", *Journal of the Environmental Satellite Amateur Users Group*, First Quarter 1988, p 6.

[5]H. P. Shuch, "Vidiot's Guide to Microwave TV", *MicroWaves*, Jun 1979, p 40.

[6]H. P. Shuch, "Low-Cost Receiver for Satellite TV", *73*, Dec 1979, pp 38-43.

[7]H. P. Shuch, "Parabolic Paradox", *QEX*, Apr 1988, pp 5-6.

[8]H. P. Shuch, "Quiet! Preamp at Work", *ham radio*, Nov 1984, pp 14-20.

[9]RF Toolbox, a collection of useful programs for MS-DOS® computers, is available gratis to microwave and RF professionals. Address your letterhead request to: Teledyne Microelectronics, 12964 Panama Street, Los Angeles, CA 90066.

[10]H. P Shuch, "Solid-State Microwave Amplifier Design", *ham radio*, Oct 1976, pp 40-47.

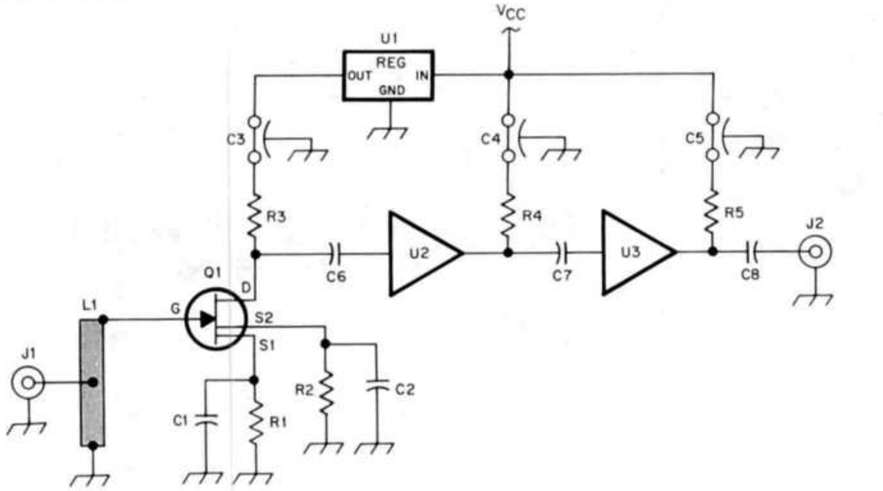
[11]SuperStar Version 3.2 is available for $595 from Circuit Busters, 1750 Mountain Glen, Stone Mountain, GA 30087. This highly sophisticated microwave circuit analysis and optimization program (the name derives from "S-parameter Two-port Analysis Routine") is somewhat slow when compared to its industry-standard counterparts, Super-Compact and Touchstone. On an 8-MHz 80286-based PC outfitted with an 80386 numeric data coprocessor, the program took several *minutes* to optimize this preamplifier. For occasional use (only as directed), I feel this is a small price to pay for a program selling for twenty times less than the competition. Please remember: No computer program is a substitute for sound engineering judgment.

[12]See note 8.

[13]"A Handy How-To-Use Guide for MAR Monolithic Drop-In Amplifiers", Mini-Circuits, PO Box 350166, Brooklyn, NY 11235, (16 pages).

[14]H. P. Shuch, "Microstripline Preamplifiers for 1296 MHz", *ham radio*, Apr 1975, pp 12-27.

[15]H. P. Shuch, "Low-Cost 1296-MHz Preamplifiers", *ham radio*, Oct 1975, pp 42-46.



C1, C2—Etched bypass capacitors (see PC artwork).
C3-C5—1000-pF feedthrough (Erie 2404-0-X5UO).
C6-C10—100-pF 10%-tolerance, ceramic chip (ATC 100B).
J1, J2—SMA connectors (Johnson 142-0298-001).
L1—Etched, tapped inductor (see PC artwork).
Q1—ATF-10235, Avantek GaAsFET.
U1—78L05 5-V regulator.
U2, U3—MAR-6 Mini-Circuits MMIC.
R1, R2—27 Ω.
R3—100 Ω.
R4, R5—510 Ω.
Misc: Die-cast, enameled enclosure (Pomona 2901).

Fig 9—Schematic diagram of the 1691-MHz VISSR LNA. All resistors are ¼-W, 5% carbon units.

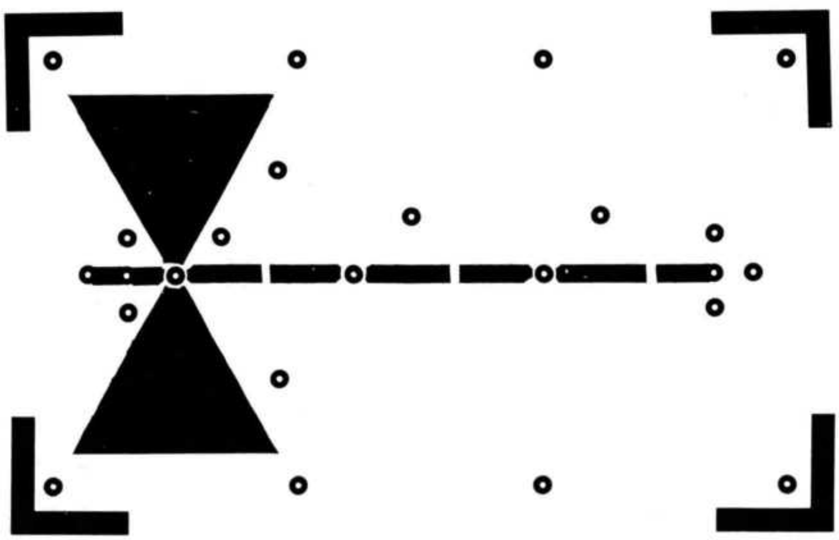
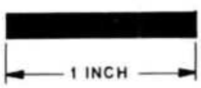


Fig 10—PC-board etching pattern for the low-noise amplifier PC-board. This pattern is etched onto one side of a $^1/_{16}$-inch-thick, double-sided, fiberglass-epoxy board. The non-component side remains fully clad, and serves as a ground plane.



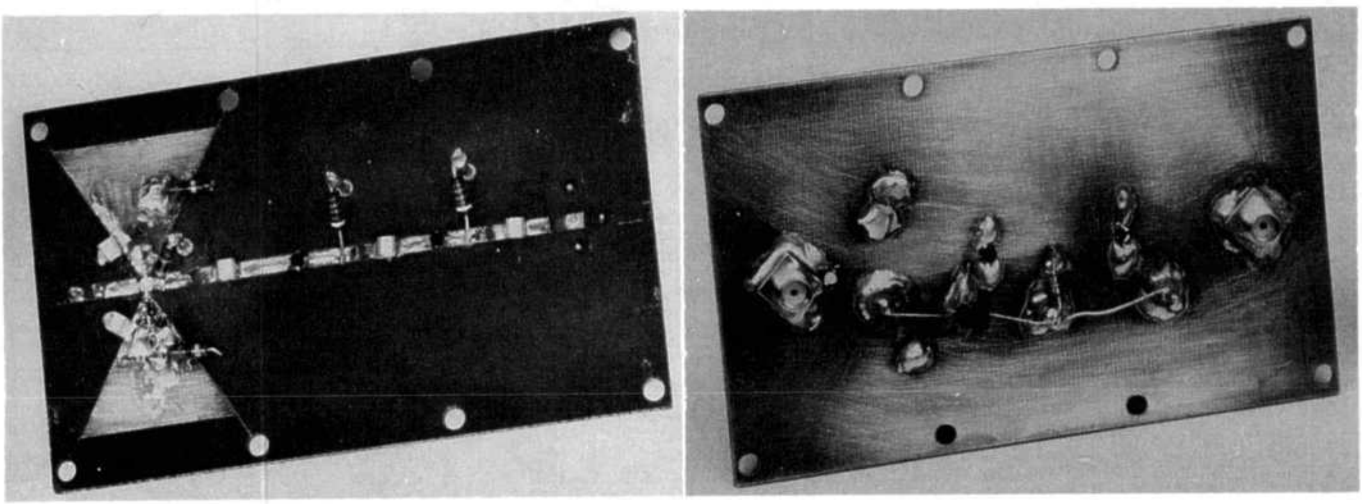


Fig 11—Views of both sides of the VISSR LNA PC board.

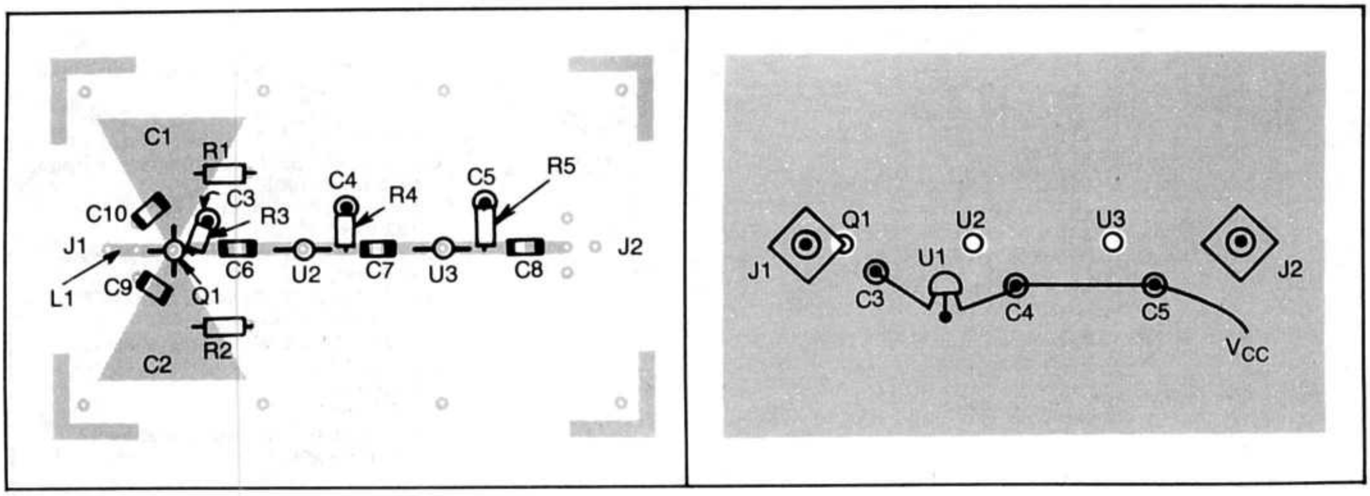


Fig 12—Parts-placement diagram for the VISSR LNA.

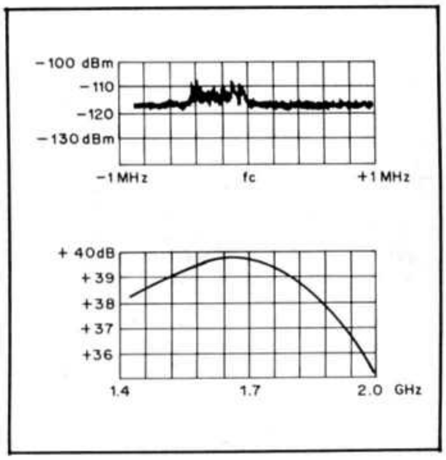


Fig 13—The swept forward gain of this preamplifier, as measured on a microwave network analyzer.

Table 1
Cascade Noise Figure and Gain

The following input data are expressed in decibels:

Stage number 1 Noise Figure:	+ 0.60
Stage number 1 Gain:	+15.00
Stage number 2 Noise Figure:	+ 2.80
Stage number 2 Gain:	+12.50
Stage number 3 Noise Figure:	+ 2.80
Stage number 3 Gain:	+12.50
Stage number 4 Noise Figure:	+ 0.00
Stage number 4 Gain:	+ 0.00
Stage number 5 Noise Figure:	+ 0.00
Stage number 5 Gain:	+ 0.00
Stage number 6 Noise Figure:	+ 0.00
Stage number 6 Gain:	+ 0.00

As a result of the above:

Composite noise figure = +0.71 dB or 1.18 as a ratio.
Total system gain = +40.00 dB or 10000.00 as a ratio.

Noise Figure Measurement

Given Information:		Calculated Result:
NF (dB)	0.71	T (eff) = 51.50573 K

Table 2
Circuit-Parameter Input to SuperStar

```
CIRCUIT VISSRFET.LNA
SST AA DG 50 20                 1691
TRL BB DG 50 ?41 1691
TWO CC SP 50 '/CIR-
CUITS/STAR/DATA/AT10235.220
RES DD PA ?68
CAX AA DD
OUTPUT
GPH AA S21        50        5 15
GPH AA S12 50 -30 -10
SMH AA S11 50
SMH AA S22 50
FREQ
SWP 1400 2000 31
```

Table 3
Run of VISSRFET.LNA Under SuperStar

Freq (MHz)	Input SWR	S21< ANG (dB)	S12 (dB)	Output SWR	K
1400	13.65	10.099< −156.55	−27.860	1.639	0.9409648
1420	11.64	10.567< −159.50	−27.259	1.704	0.9529018
1440	9.901	11.037< −162.68	−26.657	1.776	0.9635283
1460	8.391	11.507< −166.09	−26.058	1.857	0.9728609
1480	7.086	11.972< −169.76	−25.464	1.948	0.9809177
1500	5.962	12.428< −173.72	−24.882	2.049	0.9877161
1520	4.998	12.869< −177.97	−24.317	2.161	0.9932739
1540	4.175	13.289< 177.47	−23.775	2.283	0.9976093
1560	3.476	13.678< 172.61	−23.266	2.415	1.00074
1580	2.887	14.029< 167.46	−22.797	2.552	1.002685
1600	2.393	14.332< 162.05	−22.379	2.691	1.003463
1620	1.982	14.578< 156.42	−22.019	2.823	1.003094
1640	1.642	14.761< 150.63	−21.724	2.940	1.001596
1660	1.364	14.876< 144.76	−21.500	3.035	0.9989905
1680	1.140	14.922< 138.89	−21.347	3.100	0.9952967
1700	1.079	14.900< 133.10	−21.264	3.131	0.9905368
1720	1.270	14.816< 127.47	−21.245	3.128	0.9847315
1740	1.502	14.677< 122.06	−21.284	3.096	0.9779025
1760	1.767	14.491< 116.91	−21.372	3.040	0.9700729
1780	2.065	14.268< 112.04	−21.499	2.967	0.9612651
1800	2.397	14.017< 107.47	−21.657	2.884	0.9515032
1820	2.763	13.745< 103.20	−21.837	2.797	0.9408104
1840	3.164	13.460< 99.212	−22.033	2.709	0.9292111
1860	3.600	13.167< 95.498	−22.239	2.624	0.9167308
1880	4.071	12.871< 92.038	−22.450	2.542	0.9033945
1900	4.578	12.577< 88.810	−22.663	2.465	0.8892277
1920	5.120	12.286< 85.794	−22.874	2.394	0.8742574
1940	5.700	12.001< 82.971	−23.081	2.327	0.85851
1960	6.317	11.723< 80.322	−23.283	2.266	0.842012
1980	6.972	11.454< 77.831	−23.479	2.210	0.8247905
2000	7.668	11.194< 75.481	−23.668	2.158	0.8068737

Table 4
Expanded Circuit-Parameter Input To SuperStar

```
CIRCUIT VISSRCAS.LNA
SST AA DG 50 20          1691
TRL BB DG 50 ?38          1691
TWO CC SP 50 '/CIRCUITS/STAR/DATA/AT10235.220
RES DD PA ?75
TWO EE SP 50 '/CIRCUITS/STAR/DATA/MAR6.316
RES FF PA 520
TWO GG SP 50 '/CIRCUITS/STAR/DATA/MAR6.316
RES HH PA 520
CAX AA HH
OUTPUT
GPH AA S21 50 30 40
GPH AA S12 50 -60 -40
SMH AA S11 50
SMH AA S22 50
FREQ
SWP 1400 2000 31
```

Table 5
Run of VISSRCAS.LNA Under SuperStar

Freq (MHz)	Input SWR	S21<ANG (dB)	S12 (dB)	Output SWR	K
1400	15.23	37.545< 26.323	−64.115	1.644	2.424124
1420	13.27	37.788< 21.637	−63.450	1.662	2.462655
1440	11.55	38.037< 16.767	−62.788	1.680	2.497414
1460	10.04	38.289< 11.704	−62.129	1.697	2.528298
1480	8.719	38.544< 6.4383	−61.476	1.712	2.55521
1500	7.557	38.800< 0.96308	−60.831	1.726	2.578061
1520	6.558	39.014< −3.5536	−60.202	1.733	2.604033
1540	5.681	39.219< −8.2785	−59.584	1.738	2.627557
1560	4.913	39.411< −13.216	−58.982	1.739	2.648608
1580	4.243	39.588< −18.369	−58.399	1.737	2.667161
1600	3.659	39.745< −23.733	−57.838	1.731	2.68319
1620	3.152	39.878< −29.302	−57.304	1.720	2.696666
1640	2.714	39.984< −35.062	−56.802	1.705	2.707556
1660	2.336	40.057< −40.993	−56.334	1.683	2.715832
1680	2.011	40.096< −47.069	−55.906	1.657	2.721457
1700	1.733	40.096< −53.260	−55.519	1.626	2.724402
1720	1.497	40.057< −59.528	−55.176	1.591	2.724632
1740	1.297	39.977< −65.835	−54.877	1.552	2.722115
1760	1.137	39.858< −72.144	−54.622	1.511	2.716819
1780	1.084	39.701< −78.417	−54.410	1.468	2.708714
1800	1.208	39.508< −84.621	−54.236	1.425	2.697773
1820	1.373	39.285< −90.728	−54.099	1.383	2.68397
1840	1.561	39.034< −96.718	−53.994	1.342	2.667281
1860	1.770	38.760< −102.57	−53.916	1.303	2.64769
1880	1.999	38.468< −108.29	−53.862	1.267	2.625179
1900	2.249	38.163< −113.85	−53.826	1.235	2.599741
1920	2.521	37.847< −119.27	−53.807	1.207	2.571372
1940	2.815	37.525< −124.55	−53.799	1.183	2.540075
1960	3.133	37.200< −129.69	−53.799	1.165	2.505859
1980	3.474	36.875< −134.70	−53.807	1.153	2.468742
2000	3.840	36.552< −139.59	−53.818	1.147	2.42875

RF Power FETs
Their Characteristics and Applications—Part 2

By H. O. Granberg, K7ES

At low frequencies, the gate of a MOSFET is essentially a high-Q capacitor (because of MOS capacitance distributed between the channel and the top of the die). At higher frequencies, the situation becomes more complex. Because of the Miller effect and the C_{RSS}, the gate-source capacitance no longer has a fixed value (it depends largely on the drain-source voltage swing). The gate-source capacitance can usually be determined from the Smith chart in the device's data sheet, but the value is usually given only at one fixed power output. Recent technology has reduced C_{RSS} values, making gate swamping resistors less necessary to achieve stable amplifier operation. Because an FET's input Q is an inverse function of its "broadbandability," this technique is still desirable (possibly in conjunction with negative feedback) for lowering Q in wideband amplifiers. For comparison purposes, a Smith-chart plot of a 150-W MOSFET and a bipolar device with the same basic geometry is shown in Fig 10. The gate of the FET has been shunted by a resistance of 20 ohms. Without the shunt resistance, the low-frequency input impedance is almost a pure capacitive reactance.

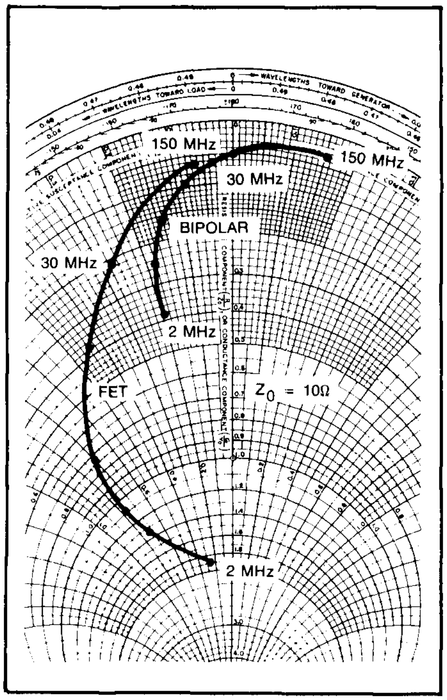


Fig 10—Smith-chart plots of the impedance of a 150-W MOSFET and a comparable BJT.

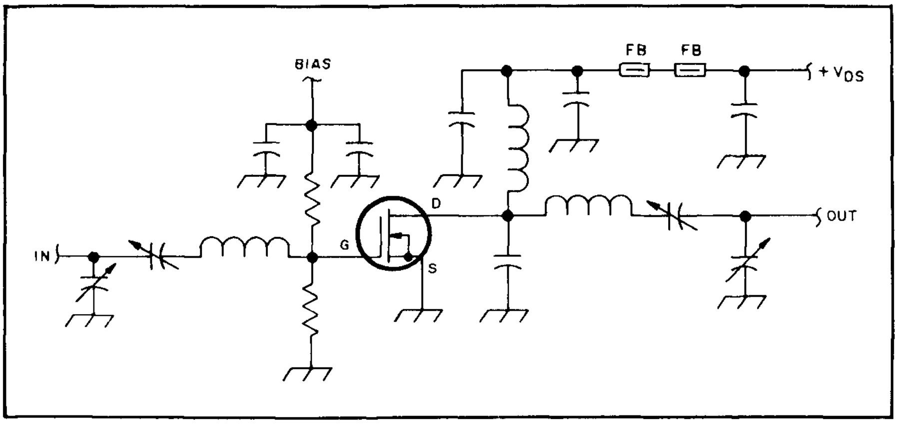


Fig 11—A typical common-source MOSFET power-amplifier circuit.

Common-source is the most widely used RF power FET circuit configuration (see Fig 11). It lends itself to both single-ended and push-pull designs. Negative feedback (cross neutralization) is easy to implement because of the phase reversal between the input and the output, which increases stability and makes circuit layout and shielding less critical. The input impedance is moderately high, even in single-ended class-A amplifiers, which are commonly employed as wideband, low-power drivers. Push-pull class-AB designs can be built to cover bandwidths of up to six octaves or more at power levels of several hundred watts (2 to 225 MHz), and five octaves at kilowatt levels (2 to 80 MHz). Power levels of 100 to 150 W at UHF are also state of the art. Narrow-band amplifiers (especially single-ended amplifiers) are frequently used in applications where the optimum performance of a device is required at a single frequency. Although MOSFETs can be operated without dc gate-bias voltage in a common-source configuration, power gain is considerably reduced, because the input voltage swing has to overcome the gate-threshold voltage to turn on the device. There is an improvement in drain efficiency by switching to class C or class D. In fact, controlling the gate-bias voltage can be used to implement an AGC function with a range of 20 to 30 dB, depending on the initial power gain of the device. AGC can be used to protect the device against mismatched loads. Because the FET's gate-source impedance changes only minimally with gate dc voltage, it is possible to design wideband amplifiers with extremely flat gain curves. Output impedance matching is similar to that used for bipolar devices. Output impedance mainly depends on the dc supply voltage, device saturation voltage and power output level. At low frequencies, the standard formulas can be applied, but at UHF, the drain-source capacitance and series inductance must be considered, as is the case with all solid-state power devices. Drain efficiency is comparable to the collector efficiency of bipolar circuits at high supply potentials (50 to 100 V), but at 28 V or less, the FET's inherently greater saturation voltage (V_{DS} [on]) results in drain efficiencies less than those of comparable bipolar devices.

The common-gate configuration, shown in Fig 12, has certain advantages, although it's not useful in applications requiring above-average linearity. The load impedance is reflected to the gate and, in effect, is in parallel with the source-to-ground impedance. The result-

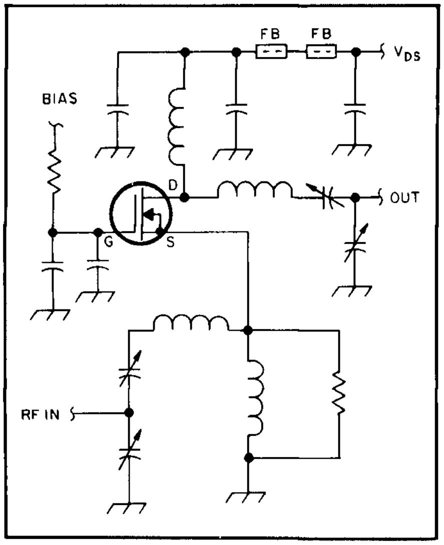


Fig 12—A typical common-gate MOSFET power-amplifier circuit.

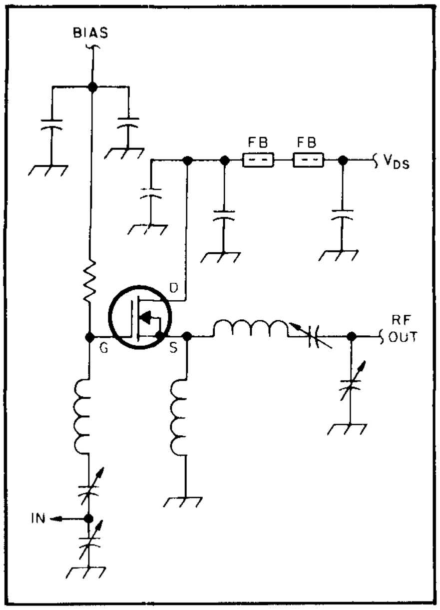


Fig 13—A typical common-drain, narrow-band MOSFET power-amplifier circuit.

ing input impedance remains fairly constant with respect to dc, but varies greatly with power output and supply voltage. As in comparable BJT configurations, the overall power gain is low, but the unity gain frequency is greater, making the common-gate circuit attractive for designs in the UHF and microwave regions. This configuration is also prone to parasitic oscillations because the input and output are in phase. "De-Qing" the input is not required in most cases, and would be possible only with shunt resistance, because negative feedback cannot be easily implemented. This circuit configuration also exhibits greater variations in power gain v gate-bias voltage, making it suitable for applications where high AGC dynamic range is desirable.

A common-drain circuit configuration is similar to an emitter follower in a BJT circuit (see Fig 13). In both cases, the input impedance is high and the load impedance is effectively in series with the input. The input capacitance (drain to gate or collector to base) is less than in common-source and common-gate circuits, and several times less for FETs (compared to BJTs of equivalent die size). This is because of the absence of the forward-biased, collector-base diode junction. A MOSFET source follower cannot be regarded as having current gain. Amplification takes place through impedance transformation. Because of the extremely high input impedance, which varies more with frequency than that in common-source and common-gate circuits, heavy resistive loading at the gate is necessary for broadband applications. Negative feedback is not necessary, nor is it easy to implement because the input and output signals are in phase. For these reasons, a common-source circuit exhibits exceptional stability, but excessive stray inductances can lead to low-frequency oscillations. Push-pull broadband circuits for 2 to 50 MHz have been designed for 200- to 300-W power levels. Their inherent characteristics are good linearity and a flat gain response without the use of leveling networks. High-power linear amplifiers are probably the most suitable application for this mode of operation. The AGC range is comparable to that of common-source configuration, but higher voltage swing is required. It must be noted that the gate rupture voltage can be easily exceeded in high-voltage operation: During the negative half cycle of the input signal, the gate voltage can approach V_{DS}.

Linearity Aspects

It is claimed in some literature that power MOSFETs have superior linearity when compared to bipolar transistors. This may be true, but only to the point where envelope distortion, instabilities or other factors are not present. As discussed earlier, MOSFETs require considerably larger die area than BJTs for comparable performance with respect to saturated power. The same holds true for linearity at a given power output level. Linearity is also a function of idle current (I_{DQ}). FETs usually require greater idle currents to achieve maximum linearity. For class AB, BJTs are biased only to the point where the base-emitter junction diode starts to conduct (forward conduction). Increasing the bias beyond that point helps little. Class A is an exception, but in class-A service, the device must be operated at approximately 25% of its rated class-AB level.

The main advantage of RF power MOSFETs is their superior high-order IMD performance (ballast resistors aren't required). In bipolar RF power transistors, nonlinear feedback is distributed to each emitter site through MOS capacitance from the collector. In devices using diffused silicon resistors, this effect is even greater because of the added nonlinear diode capacitance and the fact that the resistors themselves are nonlinear with with respect to current. High-order IMD (9th order and up) is proportional to the ballast-resistor values, which are optimized for an even power distribution along the die. Resistor values that are too low result in a fragile device; if the values are too high, high collector-emitter saturation voltage and low power gain, in addition to IMD problems, will result. Few bipolar devices with diffused silicon ballast resistors can meet today's stringent specifications for high-order IMD. Fig 14 compares the IMD values of typical BJT and FET amplifiers.

The feedback capacitance (C_{RSS}/C_{RB}) also has an effect on IMD performance. It's a function of the die geometry and is usually less with devices having a high figure of merit, such as those made for UHF and microwave applications. An RF power MOSFET exhibits several times less feedback capacitance than a bipolar transistor with similar geometry. In a bipolar transistor, this capacitance partly consists of the collector-base junction, which is highly nonlinear with respect to voltage. This, together with the varying input impedance, generates nonlinear internal feedback, which produces high-order IMD. More noticeably, low-order IMD increases as drive levels are reduced (see Fig 15). This effect can be related to the different turn-on characteristics exhibited by the two device types. When a bipolar device is biased to class AB, the bias does not usually completely overcome the V_{BE} knee. Increasing the bias above the recommended class-AB values will help, and full class-A operation should eliminate the problem entirely.

Noise Performance

Much like high-order IMD in SSB equipment, broadband noise generated in the transmitter causes adjacent-channel interference. The noise can be generated by the signal source, mixers, or any stage in the amplifier chain. Noise generated in the low-level stages is amplified in all of the following stages, and is of great concern. In linear amplifiers using BJTs, where all stages are biased, the forward-biased diode junctions can generate sufficient shot noise to disturb a nearby receiver. Thermal noise (white noise) generated by the moving electrons in the semiconductor material and emitter ballast resistors is minimal under idle con-

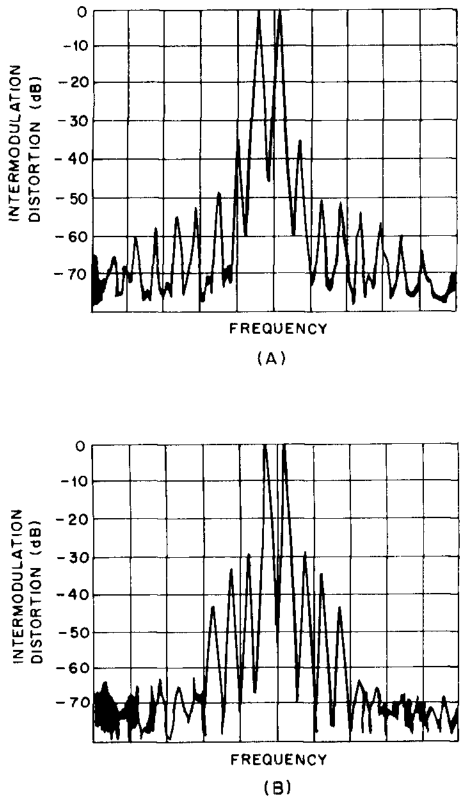


Fig 14—Typical IMD patterns for a bipolar RF transistor (A) and an RF power MOSFET (B). For the BJT, distortion products are visible up to the 15th order; for the MOSFET, 9th order and greater products are in the noise.

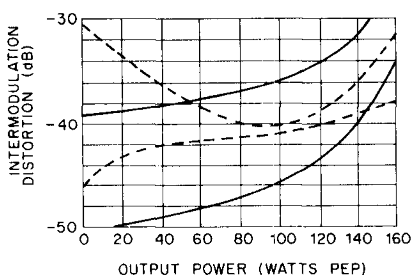


Fig 15—IMD as a function of power output. Solid lines represent a typical MOSFET; dashed lines represent a typical BJT. The upper two curves represent 3rd-order products; the lower two curves represent 5th-order products.

ditions, but gets worse at higher currents. Shot noise is mainly generated by diode junctions: these are absent in MOSFETs, leading to reduced shot noise levels in MOSFETs. Because the two types of noise have roughly equivalent spectra and are usually measured together, bipolar transistors typically exhibit noise figures that are two to three times higher than MOSFETs. The combined noise level is flat up to a point where the device gain begins to fall (upper-frequency noise corner), above which it increases approximately 6 dB per octave. There is a third type of noise commonly called surface noise or flicker noise (sometimes referred to as 1/f noise). It mostly plagues MOSFETs, which have their channels on the die surface. (Although the DMOSFET has a vertical-channel structure, it is still considered a surface-oriented device, in contrast to the JFET, which operates on a material's bulk principle.) A BJT exhibits some of both properties, depending on die geometry and wafer processing. With respect to 1/f noise, a coarse-geometry, low-frequency device performs better than one made for high-frequency use. 1/f noise increases inversely with frequency, indicating that a DMOSFET, especially one processed for RF applications, makes a poor audio amplifier. There are several other applications in which the 1/f noise figure is critical. These include high-level mixers, amplifiers for magnetic-resonance imaging systems, VCOs, amplifiers for frequency-hopping radios, and so on.

Several other factors influence the 1/f noise performance of a MOSFET. These include surface imperfections and material impurities, both of which may be difficult to control in volume production.

Switching-Mode Applications

Switching-mode RF amplifiers are becoming more and more popular in applications where high overall efficiency is desired. This is especially true since the development of the power MOSFET, which does not exhibit the carrier-storage problem inherent in bipolar devices. (The stored charge delays turn-off when a transistor is used as a switch. This situation can be corrected by not permitting the device to saturate, but in such cases the efficiency suffers.) The theoretical power output frequency limit for a power FET push-pull pair (switching-mode using today's technology) is approximately 2 kW at 1 MHz, falling to 5 to 10 W at 500 MHz.

There are several switching-mode amplifier schemes, ranging from class D to class S. Since class D represents the basic switching system, it will be the point of discussion here. Class D normally refers to a system in which the device(s) are switched on and off at the carrier frequency. Power is dissipated in the device(s) only during the switching transitions (the faster the signal rise and fall times, the greater the efficiency). The efficiency of a class-D amplifier is limited by the saturation voltage of the devices as well as their output capacitances. The theoretical maximum efficiency for class-D amplifiers is 100% (in contrast to 85% for class C, and 78.5% for class B). Practical values range from 80 to 90%, 50 to 70%, and 45 to 65%, respectively. An obvious advantage of a high-efficiency amplifier is reduced heat with respect to power output. This results in amplifiers with smaller heat sinks and more compact designs, leading to smaller devices and reduced cost. Another advantage is simplified circuit design, since interstage matching networks are not required, as they are with amplifiers operating in class A, B and C.

Class-D amplifiers are categorized as current-switching and voltage-switching types. The current-switching type is the most common; it uses a square-wave input signal. In a voltage-switching amplifier, the drive signal is usually a sine wave of sufficient intensity to saturate the output devices. An attractive aspect of the current-switching-amplifier is that the input signal can be a constant-amplitude square wave with constant pulse width, resulting in excellent gain flatness over a multioctave frequency range. From the low-level limiter to the final amplifier, the power gain can be as high as 40 to 50 dB, and the system bandwidth is limited only by the square-wave rise and fall times and the response of the output transformer or matching network. An RF MOSFET can switch almost as fast as its gate-to-source capacitance can be charged and discharged (the device output capacitance only directly influences efficiency). At low frequencies, the MOSFET gate presents a purely capacitive load to the driver, but the fast rise and fall times of a square-wave signal represent signal components that are much higher in frequency than the fundamental. The input capacitance of an FET switch is difficult to measure because it varies greatly between the open and saturated conditions, and during the switching transitions. Fig 16 shows how C_{ISS} varies with gate and drain voltages. At the left (zero gate voltage), the gate-source capacitance value is shown under the conditions under which the parameter is normally specified. At increased gate voltage, the capacitance decreases to its lowest value, just before reaching the threshold voltage. When the FET begins to draw drain current, there is a point where the device gain peaks. At that time, the drain voltage is also reduced, resulting in reduction in the depletion area and causing an overlap between the gate and the bulk material. This increases the value of the C_{RSS}, which is multiplied further by the gain reflected to the gate through the Miller effect. As a result, a sharp peak in the C_{ISS} occurs during the turn-on transition. When the FET saturates, the C_{ISS} settles to its normal value (under zero drain voltage and positive gate conditions). This peak value must be considered when designing the driver, which must have enough current-handling capability and speed to charge and discharge the C_{ISS} within a few nanoseconds. Although there is no definite rule, rise and fall times of not more than approximately 20% of the driving period are usually considered adequate. That means that at 50 MHz, the rise and fall times should not exceed 4 ns. C_{ISS} is lowest during the FET off cycle, and several times greater when the FET is in saturation. This means that the driver must supply a much higher current to turn the FET off than that required to turn it on. The turn-off cycle, during which driver dissipation is greatest, is most critical; any delays can result in both sides of a push-pull circuit drawing current simultaneously for a part of the switching cycle. The delay can be prevented or minimized by adjusting the driver-output voltage only to a level necessary to switch the output FETs to saturation. Any excess voltage increases the delay and also the dissipation of the driver. One of the most efficient and simple circuits for driving capacitive loads is a complementary bipolar emitter follower, a configuration that is increasingly feasible since the introduction of high-current, high-speed CRT drivers. Other circuits with all-NPN devices, such as those in the output buffers of MOS clock driver ICs, also offer excellent performance.

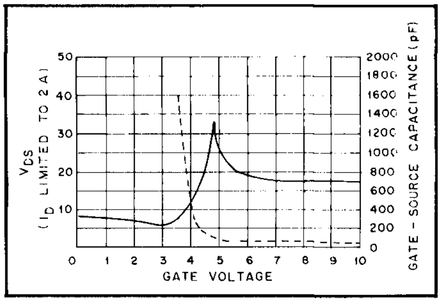


Fig 16—Typical 150-W RF MOSFET gate-source capacitance v gate and drain voltages. All power MOSFETs behave similarly, depending on processing techniques, geometry and C_{RSS}.

Good mechanical layout is important in class-D systems. Signal components of up to several hundred megahertz may be present in a class-D amplifier operating at 50 MHz, making such an amplifier comparable to UHF designs in this respect. A typical push-pull class-D PA and driver are shown in Fig 17. Using 150-W RF power FETs, four such systems have been combined to produce an output of more than 1 kW with an efficiency of 85% at 30 MHz.

Another switching-mode application for RF power FETs is in power converters or switching power supplies. Some years ago, switching-mode supplies operated at switching rates up to a few hundred kilohertz; present-day supplies operate at much higher frequencies (10 MHz and higher). The higher switching frequencies make the units more compact and lightweight, allowing the associated power transformers to be made smaller. In principle, these circuits are similar to amplifiers. Switching-mode power converters are used mainly in the computer industry and in certain military applications.

The State of the Art

Within the past few years, considerable advances have been made in RF-power-FET technology. These advances include improved manufacturing processes for higher wafer yields, finer die geometries, devices with higher figures of merit, and die structures to reduce feedback capacitance (C_{RSS}). Increased merit figures and reduced C_{RSS} directly effect the device's high-frequency and microwave performance. The C_{OSS} is critical in the sense that it affects the drain efficiency (it must be charged and discharged by the output voltage swing and can only be tuned out in narrow-band applications. The C_{ISS} primarily affects the device's input impedance and, thus, the broadband capabilities of the device (it's probably the least critical of the three capacitance parameters). In wafer processing, there is little that can be done to reduce C_{ISS} because its value mainly depends on the gate periphery. C_{OSS}, which consists mostly of MOS capacitance and the junction capacitance of the intrinsic diodes, is inherent to all MOSFETs. It can be reduced by decreasing the area of the wire bonding pads or reducing the junction size of the intrinsic diodes. C_{RSS} is more difficult to control because of its complex nature; it's affected by oxide thicknesses, Faraday shielding,

Continued on page 16.

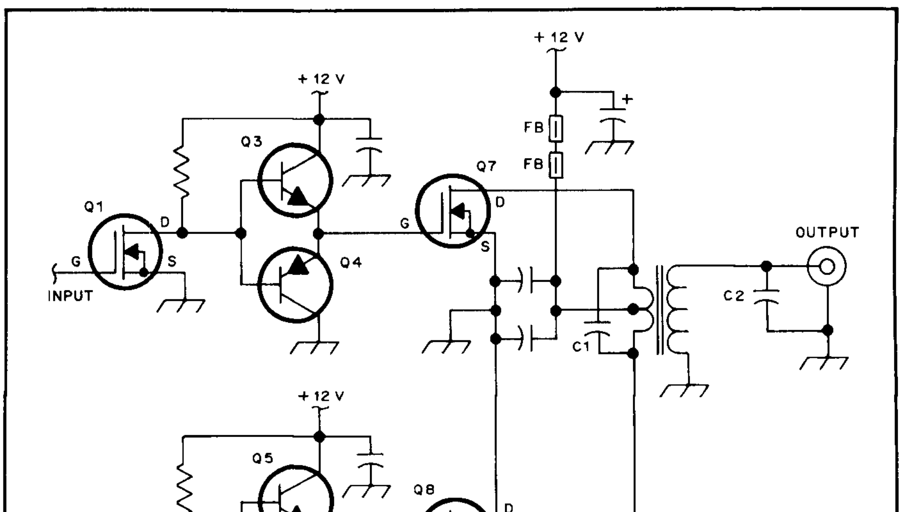


Fig 17—A typical class-D push-pull amplifier. ECL ICs are used to provide a 180° phase difference between the signals applied to Q1 and Q2.

Correspondence

A High-Stability Audio Oscillator

Here's a simple, but stable audio oscillator that uses a single quad op amp, eight resistors and four capacitors. Because of its excellent frequency stability, this oscillator can be used for many applications, such as an audio test oscillator, tone signaling or a DTMF tone encoder.

Fig 1 shows the oscillator configured as a DTMF tone encoder. The circuit resembles that of a state-variable band-pass filter shown in most active filter cookbooks. By reducing the negative feedback (with R5) and adding some positive feedback (with R6), the circuit can be made to oscillate at its nominal resonant frequency. This can be determined by equation 1:

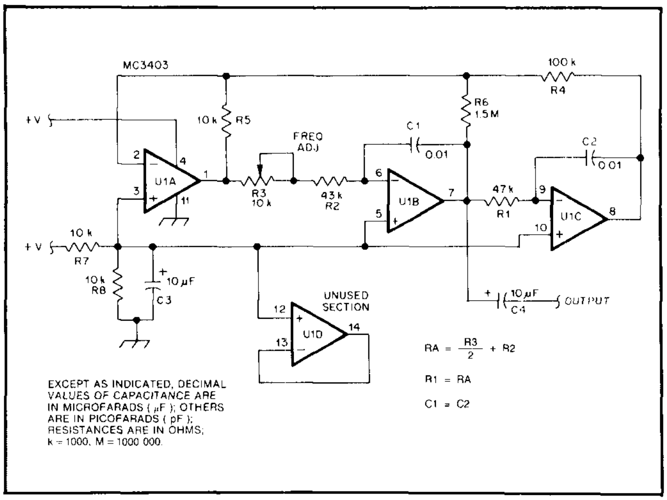


Fig 1—A single quad op amp is used to provide a high stability audio oscillator.

$$F = \left(\frac{1}{2\pi}\right)\left(\frac{1}{\sqrt{C1 \times C2 \times R1 \times RA}}\right)\left(\frac{\sqrt{R5}}{R4}\right)(0.985) \qquad \text{(Eq 1)}$$

where frequency is expressed in hertz, capacitance in microfarads and resistance in kilohms. The constant in the formula (0.985) represents the op amp open-loop bandwidth.

I prefer to select my capacitor values first because choice of values is more limited than that of resistors. As shown in Fig 1, the values of C1 and C2 are equal as are the values of R1 and RA. Working the the DTMF example, I can calculate the resistors needed for a DTMF tone by using equation 2:

$$R1 = \left(\frac{1}{2\pi F}\right)\left(\frac{1}{C1}\right)\left(\frac{\sqrt{R5}}{R4}\right)(0.985) \qquad \text{(Eq 2)}$$

Reasonable values for R1 and RA should be chosen, which may require calculation using several standard values of capacitors. Resistors with too low a value could cause op amp loading effects; too high a value could cause RFI and noise immunity problems. I try to obtain values between 5 kΩ and 150 kΩ for R1 and RA.

The following calculations are for a 107.2-Hz (1B) tone.

$$R1 = \left(\frac{1}{6.283 \times 107.2}\right)\left(\frac{1}{1 \times 10^{-8}}\right)\left(\frac{\sqrt{10{,}000}}{100{,}000}\right)(0.985)$$

$$R1 = (1.484 \times 10^{-3})(1 \times 10^{8})(0.316)(0.985)$$

$$R1 = 46.190 \text{ k}\Omega \qquad \text{(Eq 3)}$$

The nearest standard-value, 5%-tolerance resistor is used: 47 kΩ.

To calculate the value of R3, take 20% of R1 and find a trimmer potentiometer that closely approximates the calculated value.

$$R3 = R1 \times 0.20 = 9.4 \text{ k}\Omega \qquad \text{(Eq 4)}$$

The nearest standard value is 10 kΩ.

To calculate the value of R2, divide the value of R3 by two, and subtract the result from the value of R1:

$$R2 = R1 - \left(\frac{R3}{2}\right) = 42 \text{ k}\Omega \qquad \text{(Eq 5)}$$

The nearest standard-value, 5%-tolerance resistor is 43 kΩ. With a frequency counter, you can adjust R3 for the exact frequency of oscillation.

The quality of the components used in this circuit determine the oscillator's stability, particularly with varying temperatures. C1 and C2 are the most critical components. Polypropylene or polycarbonate capacitors with a 2% tolerance should be used in situations where varying temperatures are encountered. Metal-film, 1%-tolerance resistors should be used for best results. However, 5%-tolerance carbon-film resistors can also be used without experiencing too much temperature drift.

When designing an oscillator for higher frequencies, the values of C1 and C2 can become quite small. The capacitance of a PC-board layout can upset the calculations, and you'll end up with a lower oscillator operating frequency than expected. If this occurs, it may be necessary to decrease the values of R1 and RA.

The selection of a proper op amp depends on the oscillator frequency, and is important in order to obtain stability and lowest harmonic distortion. For frequencies from 1 Hz to 1 kHz, a 3403 (quad 741) will do fine. Above 1 kHz, an op amp with a greater open-loop bandwidth is necessary for proper oscillator operation. I've found that the Motorola 34079 quad op amp operates with the lowest noise and distortion, and best bandwidth up to 20 kHz. A 34079 oscillator designed for use at 1 kHz exhibited a total harmonic distortion of 0.2%. Op amps I prefer to use include the Harris HA4741, Raytheon RC4156, Motorola 34079 and, for frequencies below 1 kHz, generic 3403s.

The values of R4, R5 and R6 can remain the same as shown in Fig 1. R7 and R8 set up a voltage reference at one-half the supply voltage. C3 is the reference-supply filter capacitor, and C4 is the output dc-blocking capacitor.

Output-voltage swing is a function of supply voltage. Most op amps will operate in the circuit of Fig 1 with supply voltages ranging from 4 to 36 V. A simple three-terminal adjustable voltage regulator will provide a means of regulating the output-voltage swing. The oscillator frequency is not dependent on the supply voltage. —*Craig Carter, KA9OOP, 23860 W Rolf Rd, Plainfield, IL 60544*

VHF+ Technology

By Geoff Krauss, WA2GFP
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VHF+ conferences, at least those that I attended here in the northeastern US in 1988, seem to have matured in the sense that almost all topics presented were practical and of fairly high technical quality. The equipment and techniques covered were well within the domain of the nonengineer VHF+er, even though rigs for frequencies between 1.3 and 5.8 GHz seemed to predominate at the year's conferences. Unfortunately, a not inconsiderable group of VHF+ers seem to have other ideas, judging by the number of questions of the "I *generally* understand what to do, but I can't seem to find all the *specific* parts and subassemblies to do it with!" sort I receive. I find it especially interesting that I receive such comments most often while I am walking through a flea market where all sorts of VHF+ goodies (more so now than ever before) are available. At one very large flea market, several of the W2SZ/1 technical crew led a "tour" around the shopping area. A group of interested VHF+ers was with us, and everyone was exchanging information, pointing out various goodies and commenting on the cost, use and quality of what they saw.

At one table, we renewed our acquaintance with a gentlemen who is occasionally an excellent source of TWTs and other microwave tubes. He had a lot of the beasts out in view, and at pretty attractive prices, too. We immediately spotted a pair of Litton TWTs with N connectors. Okay—with scant data and a $35/each price tag, I can understand why most potential buyers were hesitant about buying the tubes. Taking a chance, Dick Frey, WA2AAU, and I, splurged and bought both tubes. It wasn't until three weeks later, when Dick finally found data on the tubes (they were capable of producing 10 W from 2 to 4 GHz), that we knew what we might be testing this winter, and what power supplies we would have to look for.

Next, we saw a shiny red Varian TWT with WG-90 flanges. It bore a number *known* to indicate a 10-GHz, 5-W TWT (we have several others like it). Its heater continuity tested A-OK, and its price ($20) was good. Interest in the tour group on this item was high as comments flew around about raising the piddling output (0.1 mW or less) of a well-filtered, but otherwise bare, SSB X-band mixer up to 5 W in *one* stage. Takers? None! No one knew if they could build the proper 1.5- to 2.5-kV power supplies—this response from VHF+ people who can assemble, almost in their sleep, 3- to 4-kV supplies for 1.5-kW-output linears!

We next looked at a group of TWTs intended for use from 20 to 26 GHz. Suitable for amateur work at 24.192 GHz, these tubes are about 12 inches long and about 2 inches square in cross-section, and include a built-in power supply! All you have to do is apply 120 V ac to two input pins, and ground a third pin. All of these connections are spelled out *right on the case.* The seller told us that, of a large number of units he had tested, some units had bad power assemblies and a few units had bad RF assemblies. Fortunately, the two assemblies can be separated from the case; by buying the lot, one could put together several working units. The price, even considering only 50% yield, was great if you consider that 24-GHz amplifiers are normally unavailable in *any* form. Number of takers? Two: Dick and me! Without a guarantee that the tubes worked, no one else was interested—this response from VHF+ people I've seen spend several hundred dollars for unguaranteed VHF power tubes (8877s) of questionable vintage!

Why do I again emphasize that you can't do VHF+ work that is close to *1960s* state of the art (much less, *present-day* SotA) without taking some chances and playing with lots of goodies of unknown pedigree and quality? Let's consider, as one example, the fate of a lot of millimeter-wave (>30 GHz) TWTs that this same seller told us about. He had previously sold them to a buyer in the northwestern part of the US. The buyer, after gaining access to test equipment, was able to test the TWTs and got at least a pair of them up and running. These small tubes are not 10-W units, nor are they capable of 1 W of output; typical outputs for these TWTs are in the region of 1 to 5 *milliwatts.* But with usable gains of 20 to 40 dB, the full output can be obtained if the TWT is driven by very low output power (say 1 to 10 μW, or −20 to −30 dBm) from a very inefficient transmit mixer. This extra 20 to 40 dB of system gain could be just the ticket to completion of a truly impressive QSO—like, maybe, if the millimeter-wave TWTs were used for 47.040025-GHz SSB signals.

Interestingly enough, Tom Hill, WA3RMX/7, and the Tektronix Employees' ARC, K7AUO, made a world-record 47-GHz QSO on August 6, 1988 (during the ARRL UHF Contest), using 3.5 mW of SSB (28.5-inch dish) on the 'RMX end of a 65.37-mi path and 4.3 mW (18-inch dish) on the 'AUO end. If my quick calculations are correct, a 2-ft-diam dish gives a gain of about 48 dB, and a beamwidth of about 0.6°. Achieving the required antenna aiming accuracy and stability is very difficult, but must be done, because path loss is high and, with attenuation from water vapor and other atmospheric sources, can be as great as 200 dB over a 65-mile path. Indeed, initial reports indicate that signals, S3 when contact was first established, greatly increased after clouds left the vicinity of the path!

By this time, probably all of my readers are aware of the first 10-GHz EME contacts between WA5VJB/KF5N (Kent Britain/Greg Raven) and WA7CJO/KY7B (Jim Vogler/Dave Chase) on August 27, 1988. Both stations were using high-power (50 to 200 W) TWTs on transmit, as well as low-noise (NF about 2 dB) GaAsFET preamps on receive. Antennas were a 12-ft dish at WA5VJB/KF5N and a 15.7-ft dish at WA7CJO/KY7B.

Wow, that UHF Contest weekend (August 6-7, 1988) certainly brought a lot of true VHF+ers out. Among other action at about the same time, a new North American 3456-MHz DX record was set on August 7, with a 454-mi QSO between Lauren Libby, KXØO/Ø, Pikes Peak, Colorado (13.5 W and a 32-inch dish), and Dan Osborne, WB5AFY, Vernon, Texas (*250 W* and a 6-ft dish). WB5AFY used a TWT amplifier; I've received no word on the device used in the output stage of the KXØO/Ø transmitter, but my bet is a surplus TWT!

430-MHz State of Art: Conclusion

Having discussed what can be expected in the way of antennas and transmitter gear at 70 cm, I am left with the subtopic of 430- to 450-MHz receivers. Of course, if your interest is FM work, you are probably going to use a transceiver that has a reasonably broad IF bandwidth (15 kHz). Such equipment need not have a noise figure better than 2 to 3 dB to provide fine copy of any repeater that the transceiver transmitter (typically 1 to 10 W) can raise. If a better NF is required, many silicon transistors (Motorola MRF901, NE64535 from California Electronics Labs, BFR91 from various sources, and so on) can provide NFs down to about 0.8 dB. Some prebuilt medium-noise amplifiers (MNAs) are available commercially, although many VHF + ers now build their own; the broad bandwidths and device-current settings of these designs allow them to be tuned up with any of the weak-signal methods discussed in the *ARRL Handbook*. Of course, if you choose a low-noise amplifier (LNA) with a noise figure below 0.7 dB (and possibly as low as 0.1 dB) to improve the sensitivity of a separate receiving converter or transceiver front end, you must use a GaAsFET. While a number of dual-gate types are available (MRF966, NE41137, and so on), their noise figures are generally greater than 0.5 dB. This is fine for anything except EME and other extremely-weak-signal work, for which one of the numerous single-gate devices should be used. There is no point in my providing a list of these devices, because the selection changes constantly. Typical manufacturers of *single*-gate GaAsFETs include, but are not limited to: Avantek, Mitsubishi, NEC, HP, Sony and others. All types I have tested provide more than enough gain to assure that sensitivity is set by the preamplifier (assuming that the preamp is used ahead of a receiver with a reasonable [6- to 8-dB] NF). Prebuilt preamps are available from a number of sources, some of which sell only to the amateur market, or you can build your own. Be aware of several possible problems, though: Obtaining the best noise figure generally requires the use of some fairly expensive test equipment (a noise-figure meter [such as an AIL 70 or HP 8970] and a proper noise source [the HP 346, but the A rather than the B or C models]) and knowledge of how to use it properly, and, even more important, knowledge of what can go wrong: static discharge, which can blow a hole through the vary thin gate oxide of the GaAsFET and destroy it; and the somewhat-associated problems of wide input bandwidth and conditional stability exhibited by most LNAs—traits that can lead to self-oscillation and device destruction.

Fortunately, many knowledgeable VHF + ers have access to the proper test setups these days, and the static problem is not what it was a decade ago. Many 432-MHz operators are now trying to build narrow-band input circuits, usually using cavities, to keep out-of-band signals from damaging their pet LNAs. But—possibly because of the use of high-Q filters—more potentially unstable preamps (filtered and unfiltered) seem to be appearing as more and more designs are written up without being fully tested. The most critical test I have seen is fairly simple: Put a quarter-wave cavity, tuned to your center frequency of operation, in front of the preamp and put a wideband coupler and multiple-resonator filter after the preamp. Feed signal, sampled from the coupler, to a spectrum analyzer and watch the display as normal operating power is applied to the LNA. The highly reactive source and load impedances seen by the LNA in this setup—at least at frequencies other than the chosen center frequency—will make almost all LNAs oscillate! This is due, at least in part, to the high reflection coefficients these microwave-rated devices exhibit at lower frequencies.

Can this problem be solved? I *have* had an unconditionally stable, very-low-noise GaAsFET amplifier for the 430-MHz band pass through my office, but professional ethics require that I await the issuance of a pending patent before I can get the inventor's details into print. Stay tuned!

RF Power FETs

Continued from page 13.

mask alignment, and other factors.

Research is underway to improve MOSFET capacitance characteristics. Current state-of-the-art performances include 600 W at 100 MHz, 300 W at 200 MHz, 150 W at 500 MHz and 50 W at 1 GHz.

Most RF power FETs are made to operate at device voltages of 26 or higher, although some 12-V devices are on the market. Operated at low voltage and high current levels, power FETs are less efficient than their bipolar counterparts. Another disadvantage of low-voltage FETs is their high capacitances relative to those of higher-voltage devices. Thus, an FET processed for 12-V operation would have four times the C_{RSS} of its 50-V counterpart. This results from the lower resistivity of the starting material *and* the lower operating voltage. It can be imagined how much this would affect the FET's performance at UHF and higher frequencies.

To a lesser degree, the same processing constraints hold true for bipolar devices.

We can conclude that the RF power FETs and BJTs have specific applications in which the advantages of each can be utilized.

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Bits

NEC NE647 And NE648 High-Frequency Bipolar Transistors

California Eastern Labs recently introduced two new NEC silicon bipolar transistors capable of fundamental oscillation at over 20 GHz, over mil temperature ranges. CEL claims that designers can improve phase noise in oscillators by up to 24 dB compared to MESFET designs, using the new transistors. The devices are available in production quantities. For more information, contact Steve Morris, Product Marketing Manager, CEL, 3260 Jay St, Santa Clara, CA 95054, tel 408-988-3500. —*Rus Healy, NJ2L*

Components

By Mark Forbes, KC9C
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I've collected some interesting literature for a couple of projects that I'm getting started on, and want to pass it on to you. Therefore, most of this month's column covers components related to packet radio and digital repeater control.

AMD Am79C401 Packet Controller

Advanced Micro Devices (AMD) has a new packet-controller chip available. The Am79C401 integrates a data-link controller, dual-port memory controller, and a universal synchronous/asynchronous receiver/transmitter (USART) to form what AMD calls an integrated data protocol controller (IDPC). This single-chip part can perform the intelligent hardware interface for ISDN, X.25, SNA, or other LAN applications. Communication protocols supported include SDLC, HDLC, LAPB, LAPD, and DMI; maximum data rate is 2048 bit/s, making it ideal for VHF packet applications. The chip supports several of the OSI layer-2 functions.

The interface to the packet network is via the data link controller, and the terminal interface is via the USART. The FIFO buffers allow for DMA handshaking to remove the data shuffling burden from the microprocessor. When used with an 80188 microprocessor, the Am79C401 will operate at 12.5 MHz with no wait states. This rather large IC comes in a 68-pin PLCC or LCC packages. In 100 piece lots, the cost per part is about $21. In addition to the chip, AMD also offers low level driver software (Am79LLD401) to provide a common interface to higher levels of software.

Contact Advanced Micro Devices at PO Box 3453, Sunnyvale, CA 94088, tel 800-538-8450.

Siemens SAB82525 Data Controller

The Siemens SAB82525 has features that are strikingly similar to those of the AMD chip (above) and provides a pair of full-duplex HDLC channels. Like the AMD part, it supports several of the OSI layer-2 functions. Maximum data rate is 4 Mbit/s using X.25, LAPB, LAPD or SDLC protocols. This CMOS part is packaged in a 44-pin PLCC. Availability: probably this year.

EXAR High-Speed Data Communication Chip Set

EXAR, a company that has supplied interesting IC products for a number of years, has introduced the XR-T3588 and XR-T3589 interface chip set. This set conforms to the CCITT V.35 and Bell 306 data-communication specifications, and supports communication between data terminal equipment (DTE) and data circuit-terminating equipment (DCE) at data rates from 48 kbit/s to 10 Mbit/s.

The XR-T3588 (transmitter) contains three separate transmitters and a temperature-compensated voltage source. The XR-T3589 (receiver) contains three separate receivers. The 3588 is packaged in a 14-pin DIP; the 3589, in an 18-pin DIP. In large volumes, the chips cost less than $5/set. For more information, write or phone EXAR, 2222 Qume Dr, Box 49007, San Jose, CA 95161, tel 408-434-6400.

Ericsson/Rifa PBL3726/19 Telephone Speech-Circuit IC

This chip is intended primarily for use in telephone handsets, but could be a valuable component for an autopatch circuit. The IC contains "soft-clipping" circuitry that reduces distortion from high speech levels, and an ALC circuit. The impedance of the microphone input is adjustable with an external resistor, allowing it to be matched to the output of a radio receiver. In an 18-pin DIP, the PBL3726/19 sells for less than $10. Ericsson/Rifa can be reached at 403 International Pkwy, Richardson, TX 75081, tel 214-480-8300.

Silicon Systems DTMF Transceiver

Silicon Systems has produced a line of DTMF generators and receivers over the past few years, several of which have been covered in this column. Some of their products are even available through Radio Shack®; hopefully, Radio Shack will see fit to add this product to their line.

The SSI-75T2091 is a DTMF generator and receiver/decoder. In addition to the convenience of having both functions on a low-cost IC, the part is capable of call-progress detection—a feature that allows the chip to discern busy and ring signals, and dial tones. Only a 3.58-MHz color-burst crystal and a resistor are required to complete the circuit. The SSI-75T2091 looks like a very useful component for autopatch design. The price on this part (in a 28-pin plastic DIP) is less than $15. Get complete specifications from Silicon Systems, 14351 Myford Rd, Tustin, CA 92680, tel 714-731-7110.

Crystal Semiconductor Digital Audio Converter

The CSZ5126 16-bit, stereo A/D audio converter is another part that could have interesting applications in repeater controllers, remote-base stations, and other uses. This part has many potential applications because it is optimized for speech digitization.

The CSZ5126 accepts analog input, digitizes it and then outputs the data in serial form. The chip is a two-channel device, allowing for stereo at sampling rates up to 48 kHz. Of more interest to hams is the 2 × monaural mode that allows sampling from a single channel at 96 kHz. Although I doubt that you'll ever need that kind of fidelity for communications-quality audio, the CSZ5126 is an interesting component.

Crystal also makes an evaluation board for the 5126. The evaluation board contains the digital audio converter, plus serial-to-parallel conversion for bus interfacing, voltage-reference, timing and other signals. The chip itself is expected to cost in the neighborhood of $30 to $50 in small quantities. To obtain more information about this interesting part, get in touch with Crystal Semiconductor Corp, PO Box 17847, Austin, TX 78760, tel 512-445-7222.

US Sertek PS/2 Chip Set

A PS/2® for $88? Well, not quite, but almost with this chip set. The set provides the functions necessary to build a compatible PS/2 Model 25 or 30, including system control, graphics and peripheral control. The graphics controller and peripheral controllers can be used separately for PC/XT® or PC/AT® applications. Although you need an expensive license to produce actual PS/2 computers, you might find other application for this set of controllers.

The individual chips in the set include an M1201 system controller (for the 8086 and V30), M1203 buffer, M2107 I/O controller, M2201A bidirectional printer interface, M3113 MCGA and M3115 Hercules graphics controllers, M3205 256- × -18 50-MHz color palette, and an M5103 floppy disk controller. The chip set is called the PC86; a sample set sells for $88! Interested? Contact US Sertek, 926 Thompson Pl, Sunnyvale, CA 94086, tel 408-733-3147.

Anadigics APS30010 180° Active Splitter

The APS30010 is capable of splitting a given signal into components that are 180° out of phase. In addition to such applications as two-phase clock drivers, pulse inverters, and complementary data generators, the ideal application for this product is in a balanced mixer. The APS30010's transmission delay is only 200 ps, and its noise figure is 12 dB. This data is taken from an advance flyer, so I don't have a price. You can get more information from Anadigics, Inc, 35 Technology Dr, Warren, NJ 07060, tel 201-668-5000.