QEX 34

December 1984

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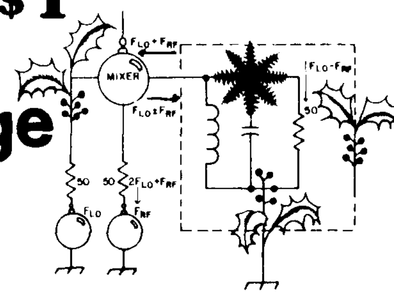
The ARRL Experimenters' Exchange



Quite a Weekend in Los Angeles

The main event during the weekend of November 9-11 at the Amfac Hotel in Los Angeles was the AMSAT technical symposium and annual meeting. But there were quite a few side meetings, as well, on a variety of subjects. Taken together, this was a gathering of hightechnomasochists comparing notes, and making plans for new projects that will finesse the state of the art, obliterate the wallet and possibly hasten divorce. I'll cover the different events of the weekend, from my vantage point, in chronological order.

On Friday evening, I had dinner with representatives of the Southern California Repeater and Remote Base Association (SCRRBA). This was mostly a get-acquainted session to discuss the UHF bands, which are more active there than in the rest of the U.S. SCRRBA representatives stressed the need for greater ARRL Hq involvement in spectrum-management issues, both technically and representationally (for example, attending meetings of professional associations of other spectrum users). Readers of the fine print in the December issue of QST will see that spectrum-management issues were a topic of discussion at the ARRL October Board meeting. It is safe to predict that spectrum management will be a hot topic for quite some time to come because of saturation of the 2-meter band in some areas and because the other VHF/UHF bands are considered increasingly important resources to be properly managed.

Later that evening, I attended a meeting of SCRRBA and ARRL to try to resolve differences in two 23-cm band plans. To backtrack a bit, at the October Board meeting, the VHF/UHF Advisory Committee presented two band plans for Board approval. The Board approved the 33-cm band plan on an interim basis. However, the 23-cm band plan was referred back to the VUAC to try to reconcile differences with SCRRBA's band plan already in being for Southern California. What was the main issue? The VUAC band plan called for a 24-MHz offset between FM repeater inputs and outputs, while SCRRBA was already using 12 MHz. After discussion which lasted to about 2 A.M., VUAC Chairman, Dick Jansson, WD4FAB, was persuaded that a 12-MHz offset would be acceptable. Assuming VUAC and VRAC ratification, the new joint 23-cm band plan will go to the Board in January 1985.

It was a little rough getting up 'n at 'em the next morning, but I got to the AMSAT registration when it opened at 8 A.M. The daytime activies were billed as the Second Annual Amateur Radio Satellite Symposium, and there were talks on satellite subjects throughout the day. There were talks on an advanced-gateway concept (Al Dayton, KA4JFO), JAS-1 (Harry Yoneda, JA1ANG), computers and the satellites (Bob Diersing, N5AHD), solar-sail project (Mark Bergham, World Space Foundation), PACSAT (Harold Price, NK6K; Wally Linstruth, WA6JPR; Rick Fleeter, WA8VGK: and Phil Karn, KA9Q), and Phase IV and future projects (Vern Riportella, WA2LQQ).

I spoke on amplitude compandored single sideband (ACSB®) via OSCAR 10. An unexpected assist was given by Jim Eagleson, WB6JNN, who brought some transparencies and an audio tape that helped to bring home some of the details of ACSB to the audience. ACSB transceivers donated to ARRL/AMSAT were turned over to Rip and Jim for on-the-air tests with W1AW in the near future. (ACSB fans: Stay tuned to QEX for more ACSB news.)

Saturday evening was the AMSAT annual meeting. There was much talk about future AMSAT projects. PACSAT and Phase III-C (another satellite like OSCAR 10 in elliptical orbit) seem practical in the next couple of years with continued financial support. Phase IV (synchronous orbit) satellites, which many think is the way to go, will cost something like 10 dB more than Phase III. Where's the money coming from?

Sunday, all day, was a meeting of the ARRL Ad Hoc Committee on Amateur Radio Digital Communications. Dick Jansson was there and gave the Committee an explanation of the new proposed 23-cm band plan. The Committee agreed that the frequency slots proposed for digital communications would provide adequate space for the development of packet radio (for example, space for some intercity trunks operating at [say] 1.544 Mbit/s). Several people reported on progress with networking experiments. Doug Lockhart, VE7APU, reported progress on standard procedures for interfacing terminals to terminal-node controllers. Doug was named the focal point for development of an "AX.3/AX.28" protocol. Transceiver turnaround times (for CW, AMTOR, and packet radio) and other topics were also discussed. - W4RI

Correspondence

The Continuing Battle on Interference

Amateur Radio operation within urban areas in this country has become increasingly difficult because of power-line interference and conduction. More importantly, the lack of concern displayed by management is equally devastating. To survive, a change in attitude must be made by the companies involved. Leakage is a wasteful nonrevenue productive effort in which radio communications suffer. Unless disaster or governmental systems are involved, almost no attention is paid the complainer.

ARRL is a nationwide organization that represents the radio amateur. They should be able to address a complementary group representing the power industry regarding this outrageous problem. A starting point might be the Edison Institute. There **must** be construction standards in which interference maximums, even after some years of service, are defined. Repair crews answering service calls simply do not understand the situation and predictably limit effort to the obvious, short term repair. Noise sources such as microwave ovens, appliance noise, computer radiation and even direct connect TV set power supplies are ignored. It is a remarkable day indeed when a repairman has more than a battery powered broadcast set in hand when infrared and ultrasonic detection devices are a minimum tool in the hands of a skilled operator.

Management is not swayed to take action unless attention is brought via customer marketing or by the Public Utilities Commission of the State. Little or no funding is provided unless an outage is suggested in which costs can be added to the rate base. At this point, limited action is provided in an effort to satisfy the PUC and little can be done to lessen the long term problems that remain. The utilities can report compliance to the Commission only when the surface difficulty has been cleared.

How can we aid the ARRL in acquiring the attention of a national representative of power distribution companies? What experience is effective in seeking improved local technical skills and facilities? How can we deal with the management that will not appreciate that power leakage is detramental to everyone, especially for the utility that is in a cost plus profit rate structure position? — Clifford Buttschardt, W6HDO, 950 Pacific St., Morro Bay, CA 93442.

L-Network Program for the TIMEX

Recently I had a desire to look at a number of L-network combinations. To ease the computations, I wrote a short program for my TIMEX 1000. Perhaps other TIMEX owners are interested. — R. H. Knaack, Jr., W7FGQ, 11415 - 28th S.W., Seattle, WA 98146.

```
10 REM "L NET"
20REMREF:SIMPLIFIEDDESIGNOFIMPEDANCE, MATCH-
   ING NETWORKS, G. GRAMMER, QST MAR 1957
30 REM ASSIGN SMALLER RESISTANCE TO SERIES ARM
   (RS)
40 REM ASSIGN LARGER RESISTANCE TO PARALLEL ARM
   (RP)
50 PRINT AT 1,8;"L NETWORK"
60 PRINT AT 3,1;"ENTER RS"
70 INPUT RS
80 PRINT AT 10,8;"RS=",RS
90 PRINT AT 5,1;"ENTER RP"
100 INPUT RP
110 PRINT AT 12,8;"RP=",RP
120 LET Q=SQR ((RP/RS)-1)
130 IF Q=0 THEN PRINT AT 15,8;"NO SOLUTION"
140 LET XS=Q*RS
150 LET XP=RP/Q
160 PRINT AT 7,1;"ENTER F,MHZ"
170 INPUT F
180 PRINT AT 14,8;"F=",F
190 LET L=.159*(XS/F)
200 LET C=159000/(F*XP)
210 PRINT AT 16,8;"L(UH)=",L
220 PRINT AT 18,8;"C(PF)=",C
230 PRINT AT 20,1;"TO CONTINUE, PRESS FUNCTION
    KEY <CONT>, AND <ENTER>
240 STOP
250 PRINT AT 5,1; "FOR ADDITIONAL FREQUENCIES,
    USING SAME RS AND RP"
260 GOTO 150
```

Switched-Capacitor Filter Technology

The switched-capacitor filter (SCF) is one device that is using digital electronic technology to its fullest. These SCFs provide the same filtering capability as its analog counterpart, the basic difference being the monolithic processing of the capacitors and the elimination of resistors in the production of SCFs. Mostly found in the telecommunications industry, they perform a rapid change of digital switching and transmission networks for telephone communications, designed to increase service capacity.

Richard Schellenbach, W1JF, and Frank Noble, W3MT, are two authors who researched this subject. If you are interested in learning more about SCFs, reference their articles in the March 1984 issue of QST, p. 19, and the July 1984 issue, p. 11.

If you are involved in working with SCFs, especially towards Amateur Radio applications, how about writing your experiences on paper and sending them in to QEX for review? We have had several requests from interested readers who would like to learn more about these digital filters. Have a happy holiday season! — KA1DYZ

Theory, Limitations and Adjustment of Reflectometers and Other SWR Meters

By Albert E. Weller, WD8KBW

The Reflectometer

The device shown schematically in Fig. 1A is called a reflectometer or monimatch. The ARRL Radio Amateur's Handbook states that this device cannot be analyzed in terms of lumped parameters, however, its inventor based the design on this, stressing that the dimensions should be small compared with the operating wavelength. [1,2]

The equivalent circuit of the reflectometer, in terms of lumped parameters, is shown in Fig. 1B. The physically short coupling line is a capacitance to the main line, C. The reactance of this capacitance, Xc, and the resistance, R, form a voltage divider. Provided the reactance is much greater than the resistance, the output voltage will be:

$$jE\omega RC$$

where E is the voltage on the main line.

The coupling line is also magnetically coupled with the main line, with a mutual inductance of M. As the magnetic circuit is unloaded, the induced voltage will be:

$$\pm jI\omega M,$$

where I is the current in the main line. The orientation of the coupling line with respect to the main line determines whether the sign is plus or minus. The voltage at the output terminal is the sum of the voltages developed by the voltage divider and the magnetic coupling:

$$e = jE\omega RC \pm jI\omega M.$$

The current, I, is simply EY, where Y is the admittance at the point on the main line where the reflectometer is located. Thus:

$$e = j\omega E(R \pm MY).$$
$$X_c$$

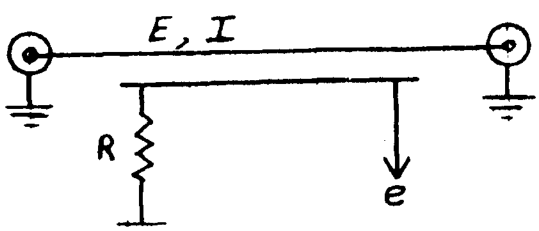


Figure 1a. Schematic Drawing of a Reflectometer.

If the admittance is a pure conductance equal to the characteristic admittance of the transmission line to which the reflectometer is attached:

$$e_i = j\omega E(R \pm MYo).$$
$$X_c$$

For the orientation to yield the negative sign, the output voltage, identified as e2, can be made zero. This is done by adjusting the parameters so that:

$$RC = MYo$$

With this choice, the two output voltages become:

$$e_i = j\omega ERC\left(1 \pm \frac{Y}{Yo}\right).$$

The ratio of the two voltages:

$$\frac{e_2}{e_1} = \frac{1 - Y/Yo}{1 + Y/Yo} = \rho$$

is the complex reflection coefficient. As:

$$SWR = \frac{1 + |\rho|}{1 - |\rho|},$$

$$SWR = \frac{1 + |e_2|/|e_1|}{1 - |e_2|/|e_1|}$$

In the following material, vi will be used to represent $|e_i|$, and v'_i to represent the measured output voltage of the coupler.

Adjustment of Reflectometers

Reflectometers typically use two coupling lines (couplers) to develop v1 and v2. This type

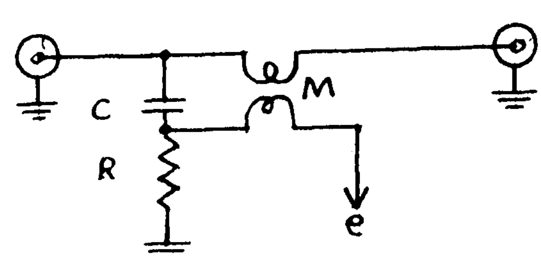


Figure 1b. Equivalent Circuit of a Reflectometer.

*505 King Ave., Columbus, OH 43201

of SWR indicator is easily adjusted to correctly indicate unity. [3] With a load equal to the characteristic admittance of the transmission line to be used, the reflected coupler, the one developing v2, is adjusted to give a zero output. Instructions generally specify adjusting the forward coupler, the one yielding v1, to give a zero output when the reflectometer is reversed in the main line. However, this has no bearing on the accuracy of reading the SWR of one correctly.

The adjustment is performed by moving the output tap along the coupling wire (changing the magnetic coupling) or bending a free end of the wire toward or away from the main line (adjusting the capacitive coupling). This procedure ensures that RiCi= MiYo. It does not ensure that R1C1 = R2C2, M1 = M2. A "calibration" factor may remain:

$$|\rho| = \frac{v_2}{v_1} * \frac{R_1C_1}{R_2C_2}$$

where R1C1/R2C2 is the calibration factor.

To have a unity calibration factor, the two couplers must be adjusted to produce the same voltages in their respective forward directions while producing zero voltages in their reflected directions when the reflectometer is connected to a load of Yo. When so adjusted, the reflectometer will correctly indicate an infinite SWR (e2 = e1) for both open- and short-circuit loads.

Keep in mind that the reflectometer and most other SWR meters do not "measure" SWR. Rather, they measure certain relations between line voltage and line current; the SWR is inferred from this measurement. It is the choice of RC = MYo that makes this inference possible. The reflectometer will indicate an SWR even when no transmission line is present and can be adjusted to indicate an SWR of one for a wide range of loads. If the load used in making the adjustments did not have an admittance of Yo = G = 1/Zo, or if the characteristic admittance of the transmission line differs from that assumed during adjustment, the reflectometer will not indicate a correct SWR.

With the main line open circuited, only the capacitive coupling is effective. When the main line is short circuited, only the magnetic coupling is effective. This is a helpful feature in performing the adjustments.

Design Considerations

Adjustment of the reflectometer is somewhat simplified if the two couplers are combined into a single line, as shown in Fig. 2. This ensures that R1C1 = R2C2 and only the magnetic coupling will need adjustment. Still a simpler adjustment is possible if a single coupler is used and the direction of the main-line current is reversed by external switching (Fig. 3). A momentary contact switch is not required if the selection switch(es) is make before break. An ordinary DPDT switch will momentarily remove the load from the transmitter, but it may or may not survive such switching. Above the HF bands, good coaxial switches or relays may be required.

Detector Diode Non-Linearities and Output Meter Limitations

The adjustments described above assume that the output of the couplers is read by a voltmeter capable of reading arbitrarily small voltages. If, as is typical, the output voltages are converted to dc by a diode detector (Fig. 4), errors will occur in adjustment and use. Good microammeters can detect a voltage of about 0.2% of the full scale reading, smaller voltages being indicated as zero. Momentarily ignoring the behavior of the diode, the minimum voltage ratio v2/v1 that can be detected is 0.002. This corresponds to an SWR of 1.004 and is the smallest measurable. In fact, the diode detector will be in its square law region when measuring v2, so that if v'_2 is about 0.004 V, then v2 will be perhaps 0.026 V. [4]

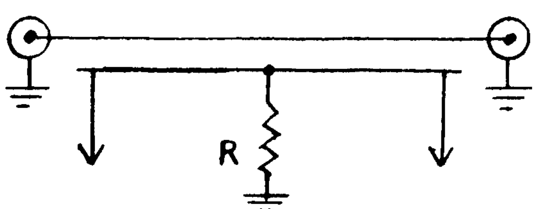


Figure 2. Two Couplers With a Common Voltage Divider.

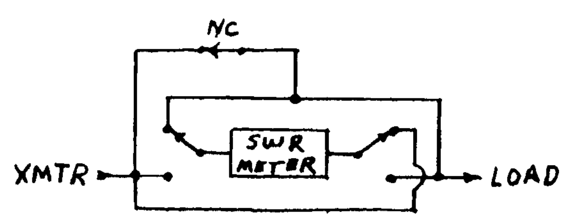


Figure 3. External Switching to Read Forward and Reflected Voltage With a Single Coupler.

v2/v1 will then be nearer 0.013 and the minimum detectable SWR will be 1.026. With a poor output indicator, detecting a minimum of 1% of full scale, and with square law operation of the diode reading the reflected wave, the minimum detectable SWR will be 1.085.

The minimum detectable voltage ratio, v2/v1, determines the directivity of the couplers. In decibels it is:

$$D = 20 \log \frac{v_2}{v_1} \text{ min.}$$

The minimum voltage ratio may be established by the minimum detectable ratio or by failure to adjust the couplers correctly. Table 1 shows the directivity achievable for different minimum detectable voltage ratios assuming a correctly measured v1 and v2 measured by a linear voltmeter or a square law voltmeter with a linear scale.

Failure to correctly adjust the couplers to give a zero output with a matched load will result in increasingly large errors as the actual SWR increases. The situation can be analyzed by considering the coupler output voltages to be:

$$e_i = j\omega \, ERC(1 \pm k_i \frac{Y}{Yo})$$

where (ki - 1) or (1 - ki) is the voltage ratio detection limit, or ki is the error in setting RiCi = MiYo.

Table 2 shows the limiting values of SWR for a 5% error and the minimum detectable SWR for various directivities. As Y/Yo is not constant on a transmission line, the error not only increases with the SWR, but varies with the position of the reflectometer in the line. This appears to be one source of error causing the incorrect readings reported by Orr. [5] The error will be a maximum at either a voltage node or a current node. Table 2 is based on the error at a voltage node assuming k1 = 1.000.

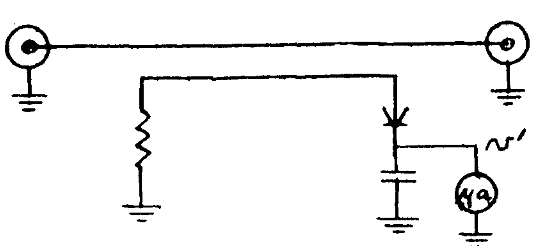


Figure 4. Typical Reflectometer Output Metering.

As both couplers are likely to have errors, the required directivity for a limiting SWR error should be increased by 3 dB for reflectometers with two couplers. For reflectometers using a single coupler so that effectively k1 = k2, partial cancellation of the error makes the limiting values of Table 2 conservative.

It is evident that any use calling for accurate measurement of high SWR requires very careful adjustment of the directivity using an output meter with excellent resolution. An example is determining transmission line loss by measuring the SWR of a shorted- or open-circuited line.

Diode detectors are nonlinear except at relatively high-input voltages and with light loads. [6] This nonlinearity will contribute errors to the SWR measurement even if the couplers are correctly adjusted. If the detector is a lightly loaded 1N34A and e1 and e2 are 1.0 and 0.5 V (SWR = 3), the output meter will read 0.905 and 0.425 V, respectively, leading to an indicated SWR of 2.77. The error depends on the absolute levels of e1 and e2 and is larger as e1 and e2 become smaller. It is smaller as e2 approaches e1 (SWR becomes larger). At the limit of very small voltages where both diodes operate in the square law region, an SWR of 3 will be indicated as 1.67.

Frequency Dependence and Transformer Couplers

As shown by the equations, the output voltages of a reflectometer are directly proportional to frequency. Where this is a disadvantage, as in using the reflectometer to measure power, the response can be flattened over an appreciable frequency range by loading the couplers with an appropriate RC combination. Reference 7 reports a flat response over a 2:1 frequency band. (While the preceding analysis of the reflectometer assumed unloaded couplers, Reference 2 indicates that the relation $|\rho| = e'_2/e'_1$ holds true regardless of the loading. Here, e'_1 is the voltage developed by the loaded coupler.) I suspect this technique is the approach used in the Bird coupler elements.

Frequency dependence can be eliminated over a very large frequency range by heavily loading the output of the magnetic coupling so that the output is effectively a current, independent of frequency and proportional to the line current, in phase or 180° out of phase with the line current, rather than a voltage having a phase angle of ±90° with the line current. The line voltage is sampled with a resistive- or capacitive-voltage divider to give an output voltage independent of frequency and in phase with the line voltage. Magnetic coupling is usually provided by a toroidal-core transformer. [8]

Stray-capacitive coupling between the primary and secondary windings of the transformer must be small. If not, there will be a frequency dependent error. In the best practice, the secondary is electrostatically shielded from the primary. [9] Also, the resistance of the transformer load must be small compared with the magnitizing impedance and large compared with the leakage impedance.

Adjustment of the transformer-coupled SWR meter is similar to that for the reflectometer. However, the magnetic coupling, if not correct, is not easily adjusted. It is necessary to adjust the transformer load(s) to achieve equal outputs from the two couplers and adjust the line-voltage dividers accordingly.

Ways to Increase Accuracy

Accurate SWR measurements with the reflectometer or transformer-coupled SWR meter requires precise adjustment of the couplers and an accurate sensitive indicating meter. Errors tend to be smaller if both adjustment and measurements are made at the highest practical power level.

The most straight forward approach to improved accuracy is to do away with the sensitivity adjustment and calibrate the indicating meter (preferably a DMM or VTVM) and detector diode as an ac voltmeter. This can be done at low frequencies where accurate, sensitive ac voltmeters can be obtained for comparison. The smoothing capacitor of the diode detector should be temporarily increased to provide a time constant of at least 10^3 periods in connection with the resistance of the output indicating meter. A 1-μF capacitor is appropriate at 60 Hz for an output meter resistance of 22MΩ. With this approach, the SWR must be calculated from the two measured voltages and the calibration.

If an attenuator is placed in the main line, the forward reading can be made equal to the reflected reading by attenuating the main-line signal. The SWR is then calculated from the attenuator setting(s) and is independent of the response law of the output indicator. The attenuator must have a power rating suitable for the operating power level. Attenuator step size must be small to accurately read high SWR — about 0.1 dB to read an SWR of 6. One dB steps are adequate for SWRs near one.

Advantage can be taken of the fact that loading the couplers does not change the fundamental relationships of the device by placing the attenuator in the coupler output lines ahead of the output indicator. The required power rating of the attenuator is greatly reduced, but a very sensitive detector or output indicator will be required. The station receiver and S meter can be used for this purpose.

Checking an SWR Meter

A good idea of the performance and usable range of an SWR meter can be obtained with the following tests. Except where indicated otherwise, the internal indicating meter and sensitivity control should be disconnected and the coupler output read with a good high impedance voltmeter.

(1) For reflectometers only: Attach a design load equivalent to 50 ohms to the antenna terminal. Apply a known (approximate) power at the highest frequency of intended use and read the forward output voltage. This voltage should be equal or less than one fifth the line voltage, or:

$$v_1 \leq 0.2 \sqrt{PZo} \ .$$

This measurement establishes that Xc >> R. If the meter fails this test, you can determine the highest usable frequency by testing at lower frequencies until the criterion is satisfied.

(2) With a design load attached, apply the highest feasible power at the highest usable frequency and measure v2, the reflected output. The output should be zero. The directivity of the reflected coupler is:

$$D = 20 \log v_2 \quad \sqrt{PZo} \ .$$

If v'_2 is very small, assuming a diode detector,

$$v_2 \simeq \sqrt{v'_2/6} \ .$$

If v'_2 is zero, the directivity is equal to or greater than the value obtained assuming v'_2 is equal to the minimum detectable voltage of the voltmeter used. This test measures the equality of the capacitive and magnetic coupling of the reflected coupler.

(3) Reverse the input and load connections to the SWR meter and repeat test two, but measuring the output of the forward coupler. This test evaluates the directivity of the forward coupler and the equality of the capacitive and magnetic coupling.

(4) With a design load attached, apply moder-

ate power and read the forward coupler output, v'_1. Remove power, reverse the input and load connections, reapply the same power and read the output of the reflected coupler, v'_2. The two voltages should be equal. You may need to monitor the applied power with an RF probe, an RF ammeter or another SWR/power meter. In the absence of anything else, the transmitter-input current can be used to ensure that the same power was applied in each case. This test determines whether the two couplers provide equal coupling.

(5) With the output open circuited, apply a small amount of power and measure the forward and reflected outputs. These should be equal. This test compares the equality of the capacitive coupling of the two couplers.

(6) With the output short circuited, apply a small amount of power and measure the outputs of the two couplers. These should be equal. This test compares the equality of the magnetic coupling of the two couplers.

(7) With a load causing a moderate SWR of 2 or 3, apply maximum power and determine the indicated SWR. Reduce the applied power and again determine the indicated SWR. Continue until the lowest usable power is reached (3 or 4 steps). The indicated SWR should be constant. Reconnect the internal output meter and sensitivity control and repeat the measurements. Compare the results with those obtained using the test voltmeter. This test reveals nonlinearity in the detector diodes and determines whether the internal meter and sensitivity control excessively load the diodes.

References

[1] The Radio Amateur's Handbook, ARRL, 1983, pp. 16-10 to 16-11.

[2] Parzen, B. and Yalow, A., "Theory and Design of the Reflectometer," Electrical Communication V 24 1947, pp. 94-100.

[3] Reference 1, p. 16-33.

[4] Weller, A., "The Theory of Diode Voltmeters and Some Applications," QEX, Jan. 1984, pp. 7-10.

[5] Orr, W., "Ham Radio Techniques," Ham Radio, V 15, June 1982, pp. 76-79.

[6] Reference 4, p. 10, Fig. 2.

[7] Stryker, E., "A Miniature Reflectometer for Portable and Mobile Transmitters," IRE Convention Record, Part 8, pp. 29-33, 1955.

[8] Reference 1, pp. 16-32 to 16-34. Operation of the coupler is more easily understood if a grounded center tap is visualized for the secondary of T_1.

[9] Adams, R. and Horvath, A., "Broadband Reflectometers at High Frequencies," Electrical Communication, V 23, 1955, pp. 118-125.

TABLE 1. ACHIEVABLE COUPLER DIRECTIVITY

Minimum Detectable Voltage Ratio	Directivity, dB Linear Detector	Square Law Detector
0.01	40	28
0.003	50	33
0.001	60	38
0.0003	70	43
0.0001	80	48

TABLE 2. EFFECT OF COUPLER DIRECTIVITY ON SWR MEASUREMENTS

Directivity dB	Maximum SWR For 5% Error	Minimum Detectable SWR
30	2.1	1.07
35	4.4	1.04
40	8.6	1.02
45	16	1.01
50	29	1.006

A 2-Meter Preamplifier Using the TI S3030 Dual-Gate GaAsFET

By Kent Britain,* WA5VJB

This article describes a 2-meter preamplifier using the new Texas Instruments S3030 dual-gate GaAsFET device announced in this issue of QEX. The circuit presented here is adapted from a design I have used successfully with other dual-gate GaAsFET devices in Micro-X packages (such as the Motorola MRF-966 and the NEC NE411).

The circuit is straightforward. I started out with an 8-V supply like TI recommends, but I found the noise figure to be 0.1 to 0.2 dB better at 4.7 V. I measured the noise figure of this preamp at approximately 0.5 to 0.6 dB with a gain of 26 to 27 dB.

I built my preamp "dead bug" style on a piece of double-sided PC-board material. A suggested layout is shown below. It is important to use a shield between the input and output circuitry. The S3030 is a high-gain device, so careful construction is necessary for stability.

The chip caps used to bypass the gate 1 and source leads are soldered to the shield in the center of the box. The source and gate 1 leads are then soldered to the chip caps, suspending the GaAsFET above the shield. The drain lead protrudes through a 1/4-in hole drilled in the shield. Keep all leads as short as possible when you mount the components. The S3030 is static-sensitive so be careful when handling it. It is best to mount the transistor last, using a grounded or cordless soldering iron.

The completed preamp may be mounted in a commercially-manufactured box. Alternatively, bottom and side plates made from PC-board material may be soldered to the assembly. Tune the variable capacitors for best noise figure on a noise-figure meter. If you do not have access to a noise-figure setup, tweak the capacitors while listening to a weak signal on the air.

This project will reward the careful builder with a good preamp at a modest cost. You may contact Texas Instruments for more information and the name of the nearest distributor. See the S3030 announcement in this issue of QEX.

*1626 Vineyard, Grand Prairie, TX 75052

(A top view of the S3030 preamp packaging can be found on page 11.)

2-meter preamp using the TI S3030 GaAsFET

S3030 N-CHANNEL GaAs-SCHOTTKY GATE FET TETRODE

electrical characteristics at 25 °C free-air temperature

PARAMETER	DESCRIPTION	TEST CONDITIONS	MIN	TYP	MAX	UNIT
V(BR)DS	DRAIN – SOURCE BREAKDOWN VOLTAGE	ID = 100 µA –VG1S = –VG2S = 4 V	12			V
V(BR)G1DO	GATE 1 DRAIN BREAKDOWN VOLTAGE	IG1 100 µA	15			V
–V(BR)G2DO	GATE 2 – DRAIN BREAKDOWN VOLTAGE	–IG2 = 100 µA	15			V
–V(BR)G1SS	GATE 1 – SOURCE BREAKDOWN VOLTAGE	–IG1 = 100 µA VG2S VDS 0	5			V
–V(BR)G2SS	GATE 2 SOURCE BREAKDOWN VOLTAGE	–IG2 100 µA VG1S VDS 0	5			V
–IG1SS	GATE 1 – SOURCE LEAKAGE CURRENT	–VG1S 2 V VG2S VDS 0			20	µA
–IG2SS	GATE 2 SOURCE LEAKAGE CURRENT	–VG2S 2 V VG1S VDS 0			20	µA
–IG1D0	GATE 1 – DRAIN LEAKAGE CURRENT	–VG1D = 10 V			60	µA
–IG2D0	GATE 2 – DRAIN LEAKAGE CURRENT	–VG2D = 10 V			60	µA
IG2S	GATE 2 – SOURCE LEAKAGE CURRENT	VG2S 2 V VDS = 8 V ID = 10 mA			100	µA
IDSS	DRAIN – SOURCE CURRENT (NOTE 1)	VG1S = 0 VG2S 2 V (NOTE 2) VDS 8 V S3030 S3030A S3030B S3030C	 20 20 30 50		 100 50 80 100	 mA mA mA mA
–VG1S(OFF)	GATE 1 – SOURCE PINCH-OFF VOLTAGE	VG2S 2 V (NOTE 2) VDS 8 V ID 1 mA		2	4.0	V
–VG2S(OFF)	GATE 2 – SOURCE PINCH-OFF VOLTAGE	VG1S 0 VDS = 8 V ID = 1 mA		1.8	3.5	V
|Y21S|	FORWARD TRANSFER ADMITTANCE	VG2S 2 V (NOTE 2) VDS 8 V ID 10 mA f = 1 kHz	18	24		mS
C11SS	INPUT CAPACITANCE	VG2S = 2 V (NOTE 2) VDS = 8 V ID 10 mA f 1 MHz		0.9		pF
C22SS	OUTPUT CAPACITANCE	VG2S 2 V (NOTE 2) VDS = 8 V ID = 10 mA f = 1 MHz		0.4		pF
GPS	POWER GAIN	VG2S 2 V (NOTE 2) VDS 8 V ID 10 mA f 1 GHz TEST CIRCUIT FIG. 1	15	22		dB
f	NOISE FIGURE	VG2S 2 V (NOTE 2) VDS = 8 V ID = 10 mA f = 1 GHz TEST CIRCUIT FIG. 1		1.1		dB

NOTE 1: IDSS MEASURED USING PULSE TECHNIQUES. TP = 300 µS
NOTE 2: VG2S APPLIED VIA SERIES RESISTOR RG2 2K2 OHMS
NOTE 3: GPSU (MAX) = MAXIMUM UNILATERAL POWER GAIN IN SOURCE CONFIGURATION
GPSU (MAX) IS CALCULATED FROM THE TERM

$$\frac{|S_{21}|^2}{(1-|S_{11}|^2)(1-|S_{22}|^2)}$$

Bits

TI Announces New GaAsFET

Arsenide Field Effect Transistors (GaAsFETs) are used by Amateur Radio enthusiasts to work and experiment in the VHF, UHF and microwave region. Back in 1978, low-noise microwave devices cost as much as $50 or more. Today, the cost has dropped considerably to the $5-10 price range. Texas Instruments recently announced a new line of GaAs-FETs costing even less — $3.50 each in single-lot quantities.

Housed in a Micro-X plastic case, the S3030 dual-gate GaAsFET features a noise figure of 1.1 dB at 1 GHz. The power gain at this frequency is typically 22 dB. Schottky-gate technology, affords superior low inter and cross modulation characteristics. The 2-micron gates of the S3030 offer a higher protection against voltage spikes and yield a higher breakdown voltage compared to a 1-micron device. For instance, the critical gate 1 and gate 2 to drain voltage is 15 V minimum.

Classified as an N-channel GaAs-Schottky Gate FET Tetrode, its applications include use in low-noise RF amplifiers up to 3 GHz. Possible commercial uses for the S3030 include cordless telephones, 900-MHz mobile radio, and cable TV amplifiers and mixers. Amateur applications include preamplifier and low-level transmitter stages for the VHF and UHF bands through 13 cm.

The S3030 is now in volume production and is available at the following prices:

Units	1	25	100	1000
$	3.50	3.03	2.80	2.33

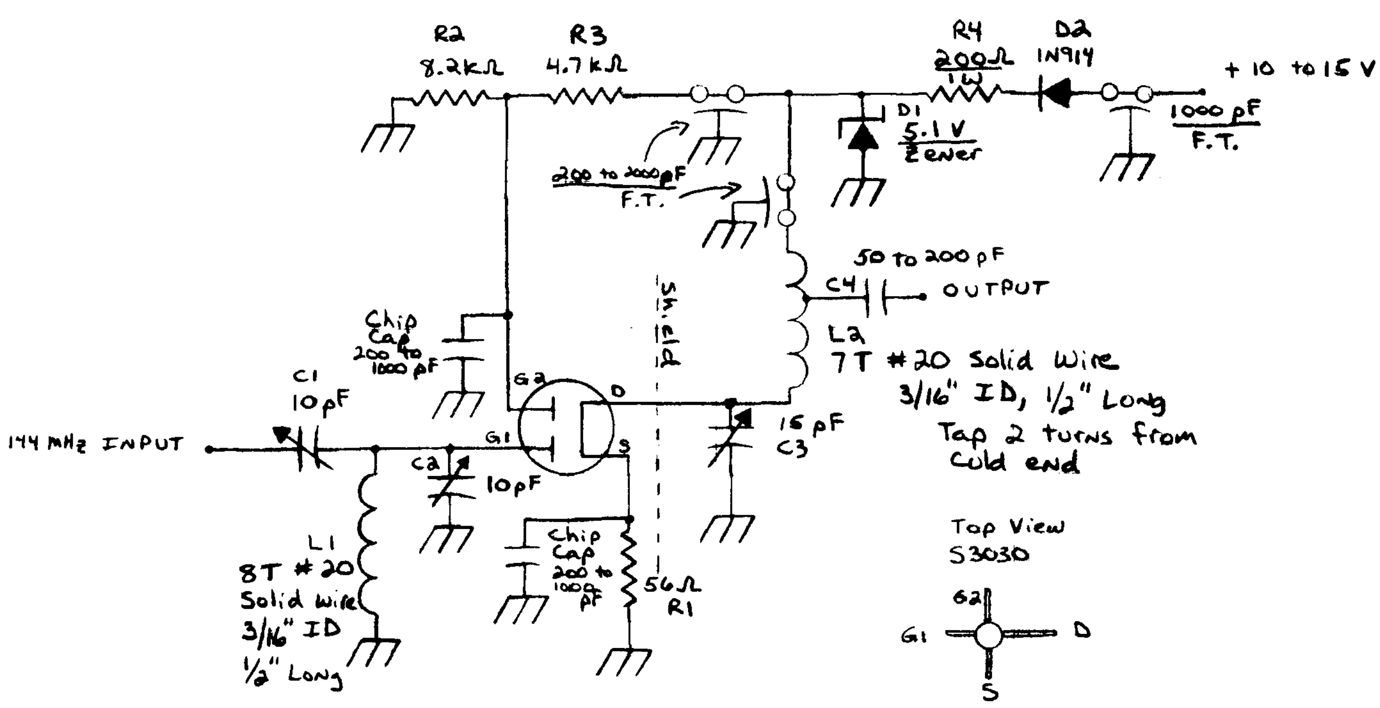
Specification sheets and other data relating to this new device are available from Texas Instruments, Inc., P. O. Box 225012, Dallas, TX 75265.

International Audio and Video Fair

Preparations are being made to host the International Audio and Video Fair in Berlin, Germany, 1985. This event will take place from August 30 to September 8. It will consist of a total of 27 halls, three pavilions, 40,000 square meters of open air grounds and the ICC Berlin will be fully booked. Hi fi, video, television equipment and the broad spectrum covered by the news media will be some of the main topics at this world's fair for consumer information and communication electronics.

This year's fair will be organized by the Messe-Veranstaltungs-Gesellschaft Unterhaltungsund Kommunickationselektronik (MVU)mbH, which was formed in June 1984. In order to display the most recent accomplishments in the industry, a major show such as this should occur at least every two years. So far, Berlin has been able to ensure such coverage.

A total of 423,390 visitors were recorded at the 1983 International Audio and Video Fair, with more than 12,000 from abroad.



VHF+ Technology

Conducted by Geoff Krauss, * WA2GFP

I recently reviewed a set of linear power amplifiers in the 50-MHz range using a number of VHF power FETs. Output powers of 100, 500, 1000 and 5000 watts were demonstrated. Unfortunately, the cost of the devices and the outrageous size of the 60-V, 50-A power supply indicates that 1500-watt output solid-state 6-m amplifiers will not be in use by amateurs for quite some time.

On Attaining High Frequency Stability

In the VHF+ frequency range, superior frequency stability is one of the hardest single characteristics to obtain. Most transmitter and receiver oscillator chains begin with a stage operating at a VHF frequency (between 30 and 100 MHz). For best stability, a crystal typically operating at the third or fifth overtone is used as the frequency-control element.

The stability of a crystal oscillator is affected by both the temperature at which that crystal is maintained and its impedance loading. A typical crystal might have a ±0.005% accuracy over some broad temperature range, even if the loading impedance is kept relatively steady. This is a difficult task because load impedance is set by active and passive components that also vary in impedance value with a change in temperature. For example, a 48 MHz, 0.005%, third-overtone crystal may have ±2.4 kHz frequency stability. This is ± 7.2kHz at 144 MHz,±21.6 kHz at 432 MHz, ± 64.8 kHz at 1296 MHz, and so on, after the appropriate frequency multiplications.

These frequency deviations are probably unacceptable, even if they are measures of long-term drift. Try arranging a weak-signal 1296 schedule and listing a frequency as being between 1295.935 and 1296.065! Your signal may never be found. Of course, most VHF+ers know about limiting temperature variations and oscillator circuit load impedance changes. A good TCXO (temperature-compensated crystal oscillator) can have ±0.0001% (one part per million, or 1 ppm) stability and a very good TCXO may have a 0.1 ppm stability. This translates to respective frequency stabilities of: ± 144 or ±14 Hz at 2 meters,±432 or ±43 Hz at 70 cm, ±1.3 kHz or ±130 Hz at 1296 MHz, and so forth. Now we're talking easier frequency locatability, even for some EME-type operating modes!

How can a VHF+er achieve such frequency stability? There are TCXOs available with 0.1- to 1-ppm long-term stability at a cost between $50 and $800+. Not only are these costs somewhat beyond the means of most hams, but also the TCXO output frequencies (1, 2, 5 or 10 MHz) would require rather elaborate oscillator chains and synthesizers to provide the basic 30- to 100-MHz signal needed at the equipment multiplier-chain input. The only realistic solution is to build a properly designed TCXO or buy from surplus.

*16 Riviera Dr., Latham, NY 12110

In the last several years, I've purchased a number of TCXOs including a surplus unit for 55- to 99-MHz overtone crystals. It is sufficiently good to employ as a model for general VHF+ use. This unit was, and may still be, available as part number OSC99828 from Fair Radio Sales Co., Inc., P. O. Box 1105, Lima, OH 45802; typical cost was in the $8-10 range over several years. Even if the surplus units are no longer available, the mechanical/electronic techniques used in this unit are worth discussing.

Any high-stability oscillator must be designed to maintain the crystal and its associated oscillator-buffer electronics at a constant temperature. This can only be accomplished if that constant temperature is greater than any expected external temperature. An internal temperature of 75°C was chosen, as it is not only within the controllable temperature range of the oscillator unit, but also because high-stability crystals, at this temperature are available from a number of suppliers.

It is important to have a crystal cut for operation at a high temperature. Crystals at room temperature will be off frequency and will not have a "concave" region of their S-shaped temperature vs. frequency curve centered on the actual operating temperature. (The concave region helps to maintain a low frequency change even if the unit's internal temperature wanders over a few tenths °C).

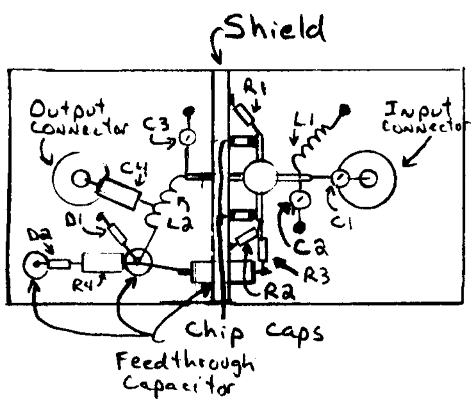
The oscillator and buffer devices should have a higher power dissipation and a larger case (2N3866) to provide a better safety factor at the elevated temperature. The actual circuit in the surplus oscillator is not necessarily the best, from a minimized load impedance change viewpoint, but it has not been replaced because of its small size. It also requires a minimum amount of operating power (+15-V dc at less than 100 mA), and has an output power monitor that can be used to tune the stages. This oscillator-buffer board comes mounted in the center of a metal box and has a pair of flat heating elements on the top and bottom covers, and a temperature-controller circuit. The power transistor of the temperature controller is mounted to the metal box. This enables any control device dissipated power to return to the basic volume and be heated to 75°C — this minimizes warmup and control response times.

The three temperature-controller thermistors, (they look like little black beads in this unit), were removed from the controller board. One is now situated near the crystal and each of the other two is positioned near an associated one of the flat heaters. The top and bottom covers and the four sidewalls of the box are frilled to add 12 to 16 self-tapping no. 6 screws in addition to the 6 to 8 screws already there. This increases thermal conduction between the heated top and bottom covers and the box walls.

The box is supplied with two complementary half-shells of a hard thermal-insulating foam; the box-shell unit is placed within a second metal box. The box-shell-outer box structure arrives in a soft foam packing box. This was kept and utilized as an outer thermal barrier. I have found the heaters to work best when the temperature controller circuit is supplied from an external +21-V dc regulator (it needs up to three-quarters of an ampere from a cold start). The resulting structure, having only a small groove running through the soft foam covers to bring out the power wiring and the RF output cable, was tested with a relatively sophisticated temperature meter. The recorded temperature, in the vicinity of the crystal, was 75 ±0.09°C over an 8-hour period, after a 4-hour warmup.

In one experiment, I measured the frequency of a 404 MHz-LO chain (for a 432-MHz receiver with a 28-MHz IF). It uses a 50.5-MHz crystal in one of the "bricks;" my frequency counter showed a maximum, and cyclical, deviation of about ±100 Hz (about ± 0.25 ppm). Not bad. It suddenly occurred to me that this deviation was also the rated stability of my frequency counter TCXO. Upon checking with a laboratory counter (having 10^{-8} accuracy, ±0.01 ppm), I found an actual overall stability of ±20 Hz (±0.05 ppm). That would be ± 65 Hz at 1296 MHz, ±115 Hz at 2304 MHz, and so on; not bad for a $10 piece of surplus equipment and a few hours of easy thermal modifications. The basic idea can be used with any homebrew oscillator-buffer unit!

(S3030 GaAsFET continued from page 8)



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Editor

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