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Test Equipment

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Testing GaAsFETs with a VOM

By Kent Britain, WA5VJB

When Al Ward, WB5LUA, first suggested the possibility of checking GaAsFETs with an ohmmeter, I gasped! The test works fine, however, if you take a few safety precautions. Use a regular VOM powered by a 1½-V battery (don't use a 120-V ac-powered DVM!) with a 10-kΩ resistor connected in series with the VOM's probe. Don't forget the usual static electricity precautions. A diagram of the test setup and typical measured values is shown in Fig 1.

The 0.01-mA current doesn't stress the transistor's gates, and 1½ V is well below the transistor's breakdown voltage. With different manufacturers using different gate designations, this test is useful in determining transistor pinouts.

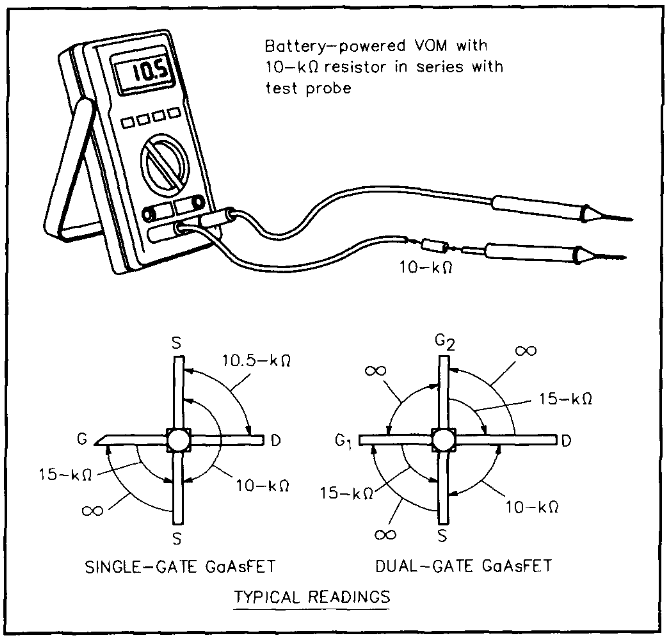


Fig 1—Test setup for testing GaAsFETs with a VOM. Typical resistance values are shown.

Microwave Absorber Notes

By Henry Burger, K7PSS

When the uses of microwave absorber materials are discussed, it's often stated that these materials are too expensive for hams to use. That's true—if the materials are purchased from a manufacturer. They are expensive indeed for a professional, building a quiet anechoic chamber for exacting tests, but they need not be so for the experimenter.

First, samples may be available from manufacturers, some of whom are hams themselves. Second, some industries may have scraps of old, unserviceable pieces that will be fine for amateur applications. Any ham that is interested in microwaves in the first place can probably find such a source. Third, the first absorbers were made in garage shops using hardware-store materials, and while not very good by today's standards, they can still be used for loads, and to dampen reflections in equipment boxes and in local areas.

The first absorbers were animal hair and wood shavings, coated with carbon-filled rubber. Flat black latex paint is a similar material. It is easy to get and handle, and its conductivity can be increased by mixing lampblack, graphite, or some other conductive material into it. Black powder toner from the waste container of a copy machine is another possibility, as is iron filings or powder. Several coats of such a mixture poured over a sheet of open-fiber packing or upholstery material should make a reasonably effective absorber for large surfaces.

You can make another type of absorber at home by loading any kind of plastic polymer with carbon or conductive material, and casting it to shape in a mold. Candidate materials are automobile body plastic, fiberglass resin from a fiberglass repair kit, and potting resins that hobbyists use to encapsulate objects for display. Again, lampblack, graphite, copy machine toner and iron filings could supply the conductivity required. Mix in the loading material before adding the catalyst. After mixing the catalyst, pour it (or cram it) into the desired mold. After curing, most of these materials can be shaped further by cutting or filing. This procedure can be used to make loads for waveguides and small cavities where it is desirable to absorb microwave energy.

These home-made absorbers can be tested in two ways. First, if a waveguide-type SWR setup is available, simply place a piece of absorber material inside the waveguide and observe the result as compared to a short circuit. Sheet materials can be tested with an open-end waveguide or small horn (the presence of the absorber in front of the open end should not increase the SWR, and should actually reduce the SWR compared to a metallic reflector).

The second way to test absorbers is to use a microwave oven. (The primary user of the oven may have something to say about this, but be firm.) Put a piece of absorber in the oven with a cup of water, and heat them for about 30 seconds. The water protects the oven if the candidate absorber material is not very lossy. If, after 30 seconds, the absorber does not get warm, forget it—if it smokes, however, you have a winner. (Better monitor the test closely for signs of overheating. You don't want to start a fire.) The faster the material heats up, the better it is as an absorber. It ought to heat up faster than the water.

One use for solid absorbers is as a waveguide load. The object is to shape the load so that it absorbs all the energy incident upon it, with none reflected. Starting with a reason-

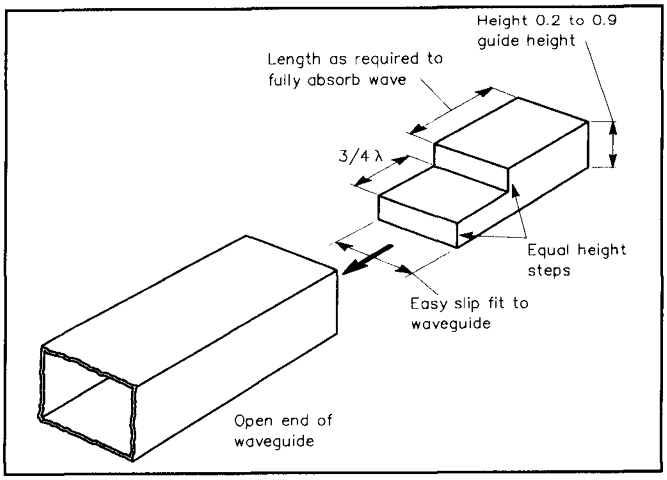


Fig 1—Construction details of a stepped-load for waveguides using microwave absorber materials. A "plug" of microwave absorber material is inserted into the open end of the waveguide. If the absorber material is formulated and shaped correctly, no microwave energy will escape. See text.

ably good absorber, shape the absorber according to Fig 1. The step is a one-quarter wavelength transformer to match the impedance of the waveguide to the impedance of the absorber. I have built waveguide loads this way using bar absorber stock.

Absorber can be applied on almost any surface where a bothersome reflection may occur, such as an unavoidable metal object near an antenna under test. For best results, the incident energy should be perpendicular to the absorber, but incidence at an angle will do. The absorber material should be several wavelengths thick, and more if the absorber is inefficient.

A dummy load for an antenna can be made this way by lining a box or a can with absorber and placing it over the antenna. (If everything is okay, the SWR on the antenna should not change when you cover it, and no signal should radiate outside the cover). This is a common practice in the industry for small antennas.

The solid absorbers can also be placed along the inside walls of circuit enclosures to absorb unwanted microwave energy. For receivers built in a can, this may be necessary for proper functioning, because all waveguide joints leak energy (Murphy's Law says this energy will always get in the way.) The enclosure for the circuit may resonate at a critical frequency, and the use of an absorber will spoil the Q of the resonance. This is also standard practice in the microwave industry.

Attenuators

By Bob Atkins, KA1GT

(From Jul 1988 *QST*)

Microwave experimenters have a wide range of applications for attenuators. Attenuators can be used in determining amplifier gain and noise performance, and they can be used to "force" an impedance match to interacting components such as diode multipliers, noise sources and load-sensitive oscillators. Attenuators can also be used in conjunction with small amplifiers to simulate isolators.

Although it is quite easy to make accurate, low-SWR attenuators for use at HF, it is considerably more difficult to do this for VHF and microwave frequencies. There are two reasons for this difficulty. First, the physical layout of the attenuator components is unlikely to provide a constant impedance through the attenuator. This causes a high SWR to exist on the transmission line to the attenuator. The magnitude of the SWR is a function of the impedance and electrical length of the mismatched section. The SWR is minimum at zero electrical length, rises to a maximum at ¼, and falls again to a minimum at ½. Thus, if the length of the mismatched attenuator is a very small fraction of the wavelength of operation (less than 0.1%), as will be the case at low frequencies, then the resultant SWR increase will be negligible.

The second reason for the difficulty in building attenuators for use at microwave frequencies is the presence of reactive components in addition to the desired resistance in resistors. Component leads have inductance whose reactance rises with frequency. Also, inter-resistor stray capacitance causes reactance that decreases with frequency. This can result in unwanted coupling. Despite these problems, it is possible to construct attenuators with reasonable performance up to the lower microwave bands. In addition, many of the measurement applications use attenuators in the IF section of a receiving system where the frequency is much lower, and attenuator performance is more predictable.

Two resistor configurations are commonly used to form attenuators, as shown in Fig 1: the T arrangement and the pi arrangement. A list of resistor values for given attenuation values can be found in Chapter 25 of recent editions of the ARRL *Handbook*. The resistors should be of carbon composition. (Don't use wire-wound resistors—they make great inductors at RF!) To provide the best impedance match to a microstrip circuit, use the technique shown in Fig 2 to connect an attenuator to a source. This arrangement uses the pi configuration of Fig 1, with the single resistors to ground replaced by pairs of paralleled resistors to minimize inductance, increase power-handling capability and create a symmetrical structure. A binary sequence of attenuation values (1 dB, 2 dB, 4 dB, 8 dB and 16 dB) will yield the largest range of possible attenuation levels with the minimum number of attenuators. To minimize stray coupling problems, no single attenuator should provide more than about 20 dB of attenuation.

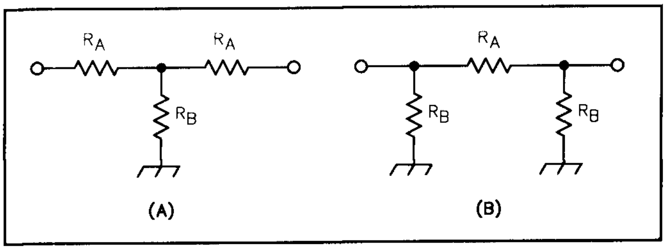


Fig 1—T (A) and pi (B) attenuator configurations.

Resistor values for various attenuation levels are given in Table 1. The values in Table 1 do not correspond exactly to commonly obtainable resistor values. One way to deal with this problem is to connect resistors in series and/or parallel to approximate the desired resistance. This approach is not desirable if the attenuator is to be used above VHF, because increased physical attenuator size and reactances can cause problems. A second alternative is to select resistors that have

Table 1
Pi-network Attenuator Resistor Values

Attenuation (dB)	*R_A (ohms)*	*R_B (ohms)*
1	5.8	869.5
2	11.6	436.2
4	23.8	221.0
8	52.8	116.1
16	153.8	68.8

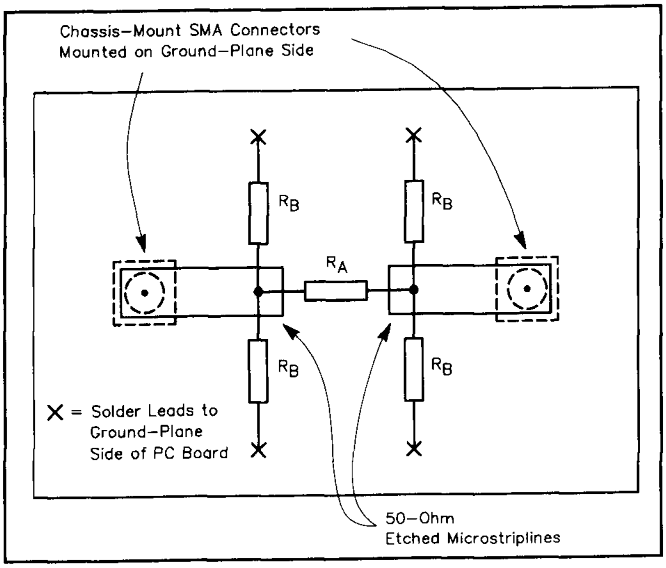


Fig 2—Circuit-board layout for a 50-Ω, pi-configuration microstrip attenuator for VHF use. Microstripline width should be 0.1 in. if double-sided, 1/16-in.-thick glass-epoxy PC-board material is used. Keep all resistor lead lengths as short as possible to minimize reactances at high frequencies.

slightly lower resistance than that required. A small round file can then be used to remove a small amount of the resistor, as shown in Fig 3. This will increase its resistance. After the resistor has been adjusted to the required value, a small dab of paint should be applied to cover the exposed carbon. If high power-handling capacity is not required, 1/8-W resistors may be used. The smaller the physical size of the resistors used, the larger the frequency range over which the attenuator will function as designed. Chip resistors are ideal for use in UHF attenuators, if the correct values are available.

Resistive attenuators built as described should work from dc up to the frequency at which impedance mismatch and/or reactances become significant. It follows that attenuators can be characterized at dc, and their RF performance can be inferred from this. To check the impedance match, the input and output dc resistance to ground should be 50Ω. Attenuation can be calculated as follows: Terminate one port of the attenuator with a 50-Ω load. Apply a small dc voltage to the unterminated attenuator port, and measure the dc voltage at the terminated port. Attenuation is given by:

$$A \text{ (dB)} = 20 \log (\text{Vout} / \text{Vin}) \qquad \text{(Eq 1)}$$

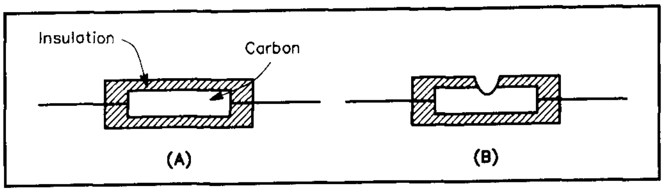


Fig 3—Cross-section of a carbon-composition resistor before (A) and after (B) filing to increase resistance. Be sure to seal the body of the resistor after filing (use a few drops of paint or nail polish) to prevent contamination and subsequent resistance-value drifting.

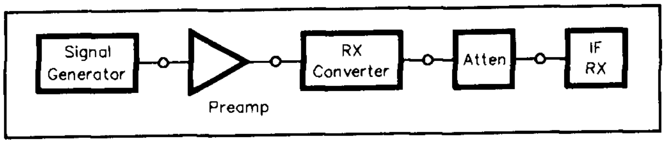


Fig 4—Preamplifier-gain measurement setup. See text.

Remember that this is only a dc check—how these values will change at RF depends on the factors described earlier. For accurate RF attenuation measurements, more complex testing is required. Alternatively, attenuators can be calibrated by comparison with attenuators of known accuracy.

Attenuator Uses

A typical setup for measuring preamplifier gain using attenuators is shown in Fig 4. A signal source is fed into a converter and the output of the converter fed into an IF receiver, typically at 144 MHz or below. The signal level at the IF receiver (S-meter reading or audio output voltage) is then noted. The preamp is then put in line between the signal source and the converter and an attenuator is placed between the converter and IF receiver. The attenuator is then adjusted until the IF signal level duplicates that measured previously. At this point, the attenuator is canceling the preamplifier gain (assuming the converter is operating in its linear region). The preamp gain (in decibels) is then given by the total attenuation in decibels. Since the signal level at the IF receiver is the same in both cases, nonlinearities (such as AGC action) do not affect the result. An attenuator can also be placed ahead of a converter, if the attenuator is known to be accurate at a higher frequency.

Noise-figure (NF) comparisons can be made with the arrangement shown in Fig 5. A noise source is connected to a

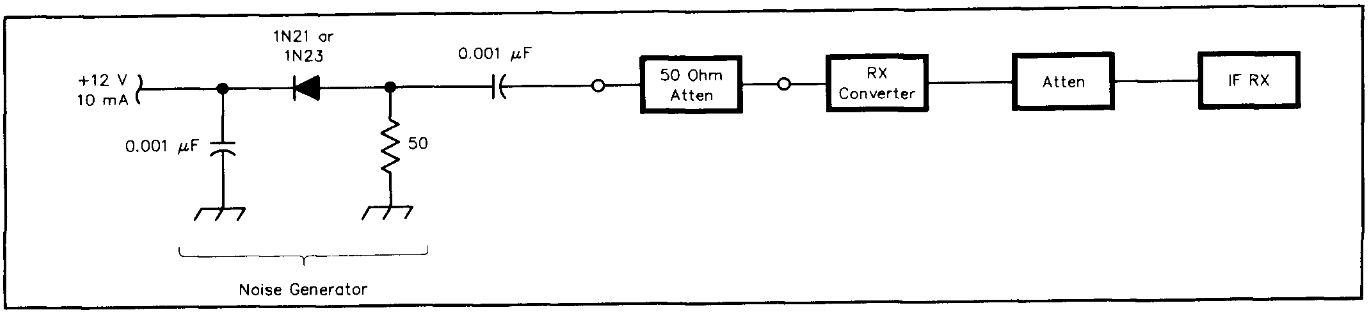


Fig 5—Receiver-noise measurement system. The attenuator immediately after the noise generator is for impedance-matching purposes. See text.

converter via an attenuator. The attenuator provides a 50-Ω match between the noise source and converter. The output of the converter is then fed to the IF receiver via a variable attenuator. The receiver output (S-meter reading or audio voltage) is noted with the generator off and the IF attenuator set to 0 dB. The noise generator is then turned on and the IF attenuator adjusted for the same reading as previously noted. The amount of attenuation required is related to the converter noise figure—the more attenuation required, the better the NF. The front end of the converter may then be tuned to yield the best noise figure, or two different converters may be compared.

Simple diode noise generators are useful for NF comparison tests, but are not reproducible enough to make accurate measurements. If a stable, calibrated noise source is available, this method may be used to determine actual NF.[1]

Note

[1] Schetgen, ed., *The ARRL Handbook for Radio Amateurs*, 71st ed. (Newington: ARRL, 1993), p 12-3.

Thermistor Power Metering

By Bob Atkins, KA1GT

(From *QST*, Aug 1988)

When building microwave equipment, it is often desirable to measure the relatively low output power of local oscillators or mixers. One cheap, simple and nearly foolproof way of doing this is by using a thermistor power meter. A thermistor is a device with resistance that depends on temperature. The power meter shown in Fig 1 uses a thermistor to sense the change in temperature of a 50-Ω, ⅛-W resistor used as a load for the power being measured. The 50-Ω load resistor dissipates applied RF in the form of heat, causing the resistor temperature to rise. The thermistor is in contact with the resistor, secured by a spot of epoxy glue. As the temperature of the thermistor rises, its resistance changes (usually, it decreases). When the resistance reaches a steady value, the reading is noted and the RF power removed. A dc voltage is then applied to the resistor. The voltage is adjusted so that the thermistor resistance stabilizes at the same value as that obtained when the RF power was applied to the resistor. Under these conditions the dc power and RF power applied to the resistor are equal and given by:

$$P = V2 / R \qquad \text{(Eq 1)}$$

A few points to note: The physically smaller the load resistor, the higher and more rapidly its temperature will rise for a given input power, hence, the greater the sensitivity of the measurement system. With a ⅛-W resistor, power levels of 10 mW can be easily measured. The smaller the thermistor bead and leads, the less heat it will absorb and the better it will track the resistor temperature.

There are, of course, some problems with this technique. It assumes that all of the power supplied by the generator is dissipated in the resistor. Any reflected power will not be measured, so a good match is required. Perhaps the best physical layout would be a 50-Ω microstrip terminated by a small 50-Ω chip resistor.

The temperature reached by the resistor/thermistor combination depends not only on input power, but also on ambient temperature. If the circuit is in an enclosed box, some drifting of readings may occur if there is too much heat buildup in the box. Because of this, it is better to measure higher power levels (more than 50 mW) using calibrated attenuators ahead of the

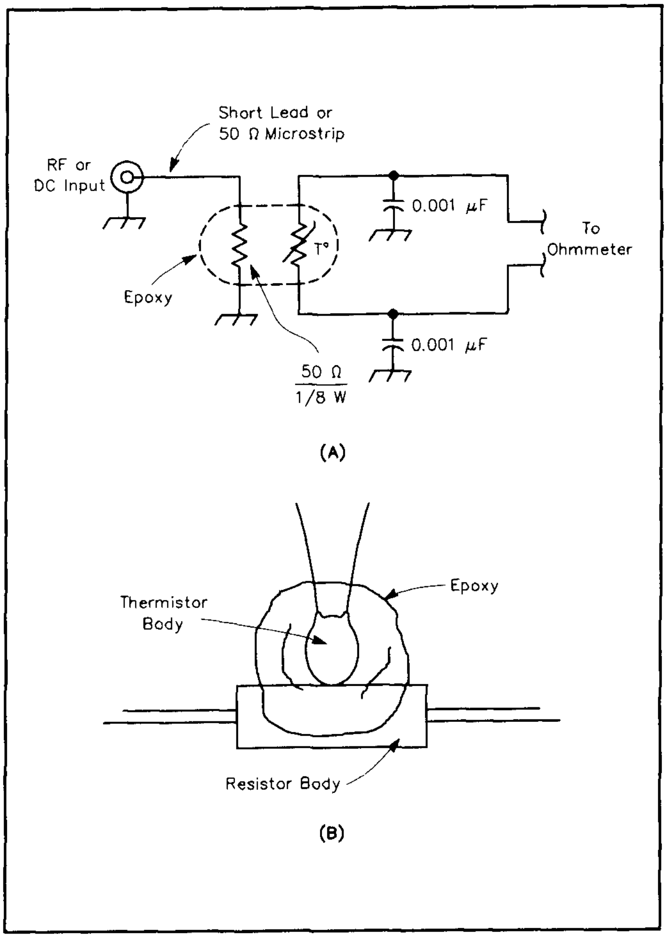


Fig 1—The schematic diagram of a thermistor power meter is shown at A. An ohmmeter is connected to the thermistor leads and resistance drop measured while power is applied to the load resistor, after the load temperature stabilizes. The resistance change of the thermistor from room temperature is then calculated. A dc signal that causes the same resistance change is then applied, and the power level calculated. The smaller the load resistor, the more precise the measurement can be. At B, the method of physical attachment of the thermistor to the load resistor is shown.

power meter. Finally, the system response time is slow (several tens of seconds, depending on the input power and thermal capacity of the system), so it cannot be used as an indicator to "peak" circuits. Nevertheless, I have found this type of instrument quite useful, and it can be constructed for only a few dollars.

Thermistor prices depend on the thermistor used (its stability, materials, construction and so on), but are typically in the $5 range. They may be obtained through local electronic parts distributors. One thermistor I have used is the Fenwell Electronics type 112-503JAJ-B01. Because it has a 0.040-inch bead with very thin lead wires, its thermal mass is small. Its resistance at room temperature is about 50 kΩ, and drops by about 7 kΩ when the device is in contact with a ⅛-W, 50-Ω resistor dissipating 10 mW. There are undoubtedly many other thermistor types and brands that would work equally well. A final point to remember is that this measurement technique measures the total power delivered to the resistor, including power at the desired frequency and that at all other frequencies (such as harmonics). This should not usually cause problems unless a particularly "dirty" source is being measured!

Zap Insurance—A Simple RF Detector for Operator Safety

By W. O. Troetschel, K6UQH

As higher levels of microwave power become available, amateurs must seriously consider a method of detecting dangerous levels of stray RF. The "Zap Insurance" detector (Fig 1), designed and constructed by Bill Troetschel, K6UQH, can help you locate sources of stray energy. When placed at head level, it provides a positive indication that your operating area is safe.

Circuit Details

The schematic diagram is shown in Fig 2. RF energy from a probe antenna is fed to a 50-ohm input circuit, consisting of R1 and R2. Germanium diode D1 is a simple RF detector. An RC filter follows, consisting of C1, C2, R3 and R4. The resulting dc is fed to M1 via potentiometer R5, which provides calibration adjustment. The frequency response of this simple circuit is extremely flat from 28 to 1296 MHz, and more than adequate up to 3456 MHz.

Construction

Fig 3 shows the parts layout. The entire circuit is built on a 1/16 × 1¾ × 3-inch piece of single- or double-sided glass-epoxy circuit board. A hole is drilled for the Type-N connector, which mounts on the underside of the circuit board (Fig 3C). A shim is placed between the connector flange and circuit board so that the ground "sleeve" on the connector is flush with the parts side of the board. Four screws are used to hold the connector in place, but a pair of thin brass shims must also be positioned through the circuit board. These should be soldered to both ground planes (or to one ground plane and the

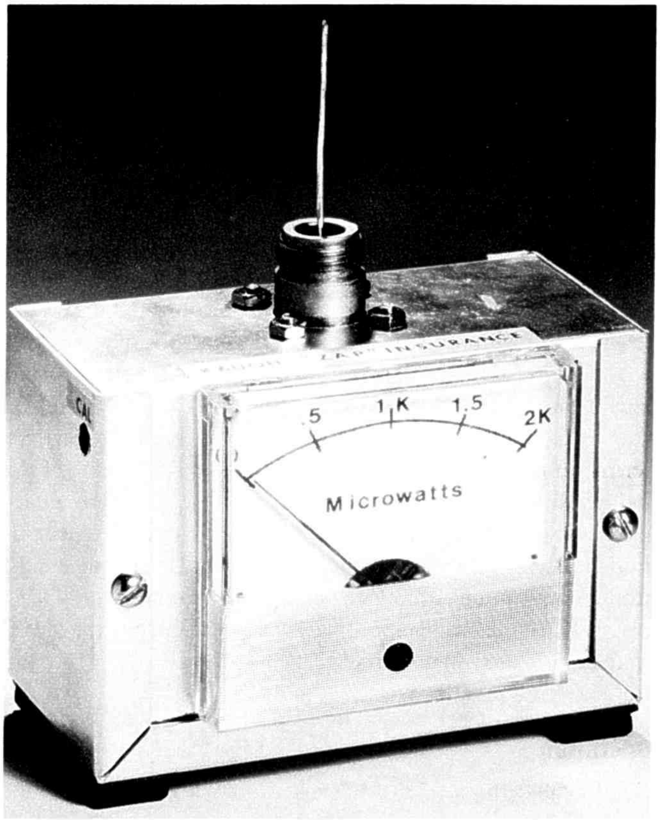


Fig 1—View of the K6UQH "Zap Insurance" detector, with a 1296-MHz probe antenna in place.

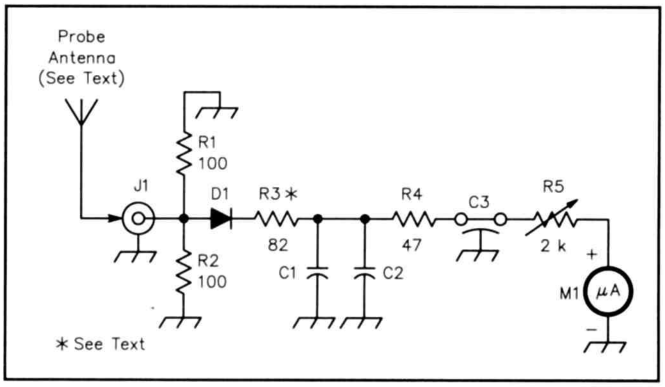


Fig 2—Schematic diagram of the zap detector. All resistors are ¼-W, carbon film.

C1—15- to 27-pF chip capacitor.
C2—560- to 910-pF chip capacitor.
C3—1000-pF feedthrough capacitor.
D1—1N82AG-C (International Rectifier Co).
J1—Chassis-mount female Type-N connector.
M1—0-200-μA panel meter.
R5—Miniature trimmer potentiometer.

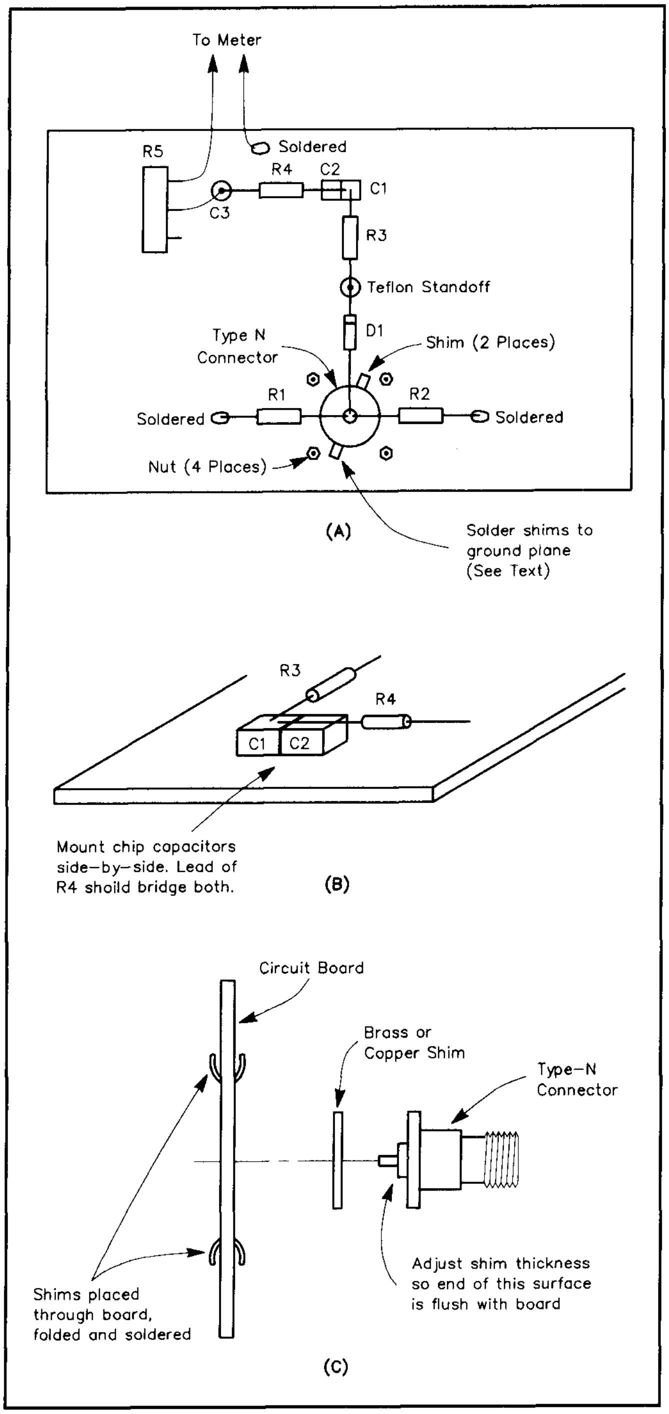


Fig 3—Top view of the circuit board (A). Part placement is not critical, with the exception of C1 and C2 (B). All leads should be kept as short as possible. Connector mounting detail is shown in C.

connector housing if single-sided board is used) to provide a long-lasting RF connection. Holes are also drilled for mounting the feedthrough capacitor and Teflon standoff insulator. The latter hole should be small enough so that the standoff is a press fit.

Keep all component leads as short as possible to prevent unwanted resonances. In addition, pay particular attention to the mounting of C1 and C2 (Fig 3B). The circuit-board assembly and meter are mounted together in a 4 × 2¾ × 2-inch LMB chassis box. A hole should be made in the box to provide access to the calibration trimmer potentiometer.

Calibration

The detector is calibrated by applying a 2-mW, 1296-MHz signal to the Type-N connector. Adjust R5 for a full-scale meter deflection. Reduce the drive level to 1 mW and note the reading. Using this point as a reference, apply a 1-mW, 28-MHz signal and check this reading against the reference. Should they coincide, no further adjustments are necessary for the range of 28 to 1296 MHz. If the 28-MHz reading is less than the 1296-MHz reading, increase the value of R3. If the 28-MHz reading is greater than the 1296-MHz reading, decrease the value of R3. The proper value of R3 is both diode and load dependent, and may vary from approximately 68 to 120 ohms. Whenever R3 is adjusted, always return to the beginning of the calibration procedure.

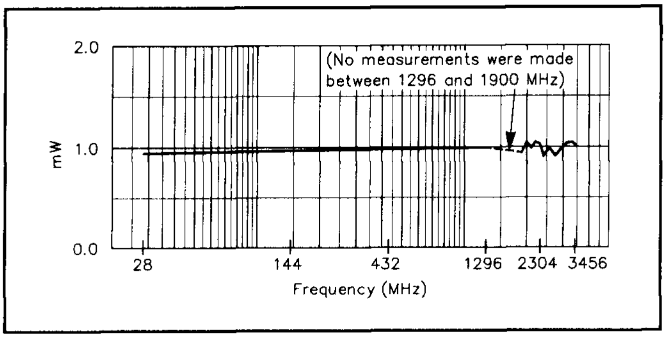


Fig 4—Relative response of the zap detector with 1 mW of RF applied.

Fig 4 shows the measured response of the unit from 28 to 3456 MHz. Notice the minor resonances occurring above 2000 MHz. These are minimized by a secondary calibration procedure. The object is to "tune" out stray inductance and capacitance by adjusting the height of D1 and R3 above the ground plane. With the proper equipment and a little patience, it's possible to extend the useful range of this instrument to well above 1296 MHz, as shown in the graph. The only possible exception is if a substitute diode is used for the 1N82AG-C. Many of the modern-day substitutes are silicon, rather than germanium. If a silicon diode is used, varying the value of R3 and R5 should allow operation up to 1296 MHz, but no guarantee of higher-frequency operation is given. Differences in diode junction capacitance and "catwhisker" inductance may produce unwanted resonances that can't be tuned out.

When calibration is completed, you can change the meter scale to indicate 2 mW full-scale. Alternatively, it can be calibrated in microwatts, as shown in Fig 1.

Operation

Because the zap detector is a simple device, you should not expect it to be as accurate as a true power-density meter. However, when checked against a commercial meter of this type, the accuracy was found to be surprisingly good. The zap detector's meter is a bit generous, which provides an additional, "built-in" safety factor.

Operation is simple. A piece of heavy bus wire is cut to a quarter wavelength on the band you wish to monitor, and inserted in the Type-N connector. A quick check can be made by inserting a 2304-MHz antenna and placing the unit near the door of an operating microwave oven. In most cases, the meter should read about 500 microwatts. (If it reads higher, have your oven checked for radiation leaks!) The detector should be placed near head level in the shack as an RF "sniffer" during transmission.

A Milliwattmeter for HF to 1296 MHz

By W. O. Troetschel, K6UQH

One of the most exasperating problems many hams face is accurately measuring power levels from HF to 1296 MHz, and knowing how the power readings taken at different frequencies relate to one another.

This device gets around those problems, and thus provides a method of measuring filter insertion loss, amplifier stage gain, oscillator/multiplier output power, and of course, serves as a field strength meter, etc.

The version shown in Fig 1 measures power in two valuable ranges: 0 to 10 mW and 0 to 100 mW. It also can be constructed with a single power scale, and either version can be easily extended to higher power ranges by appropriate external attenuators, directional couplers, or a combination of the two.

Different types of power meters have their advantages and disadvantages. The Milliwattmeter offers portability, quick power measurements, and ease of construction. It also measures everything from HF to 1296 MHz and up! Thus, accurate measurements require clean signals.

This problem shows up when measuring the output from an oscillator/multiplier chain. You might read 100 mW initially, and 40 mW after you install a filter on the output. Ignoring filter insertion loss, the difference represents power at other "unwanted" frequencies! However, once you know this, it becomes a manageable problem, and is in itself quite revealing.

In Fig 2, the detector's frequency response is compared to that of a commercial HP-423A detector head (for use up to 12.4 GHz). If you can get your hands on one of these—or its equivalent—to calibrate your meter scale, you'll really have a piece of equipment!

To build an accurate Milliwattmeter you must be willing to make your own meter scales. By doing this, you ease the problem of finding a suitable meter, and increase the accuracy of the unit (equal to your calibration standard). Buy or borrow some drafting pens and India ink, locate some white paper or cardboard that doesn't soak up the ink like a blotter, and "have at it." It's really not that difficult after the first dozen tries! Rub-on transfers complete the job.

In the dc and digital world, diodes are commonly viewed as simple-minded on/off switches. In the RF world, diodes are

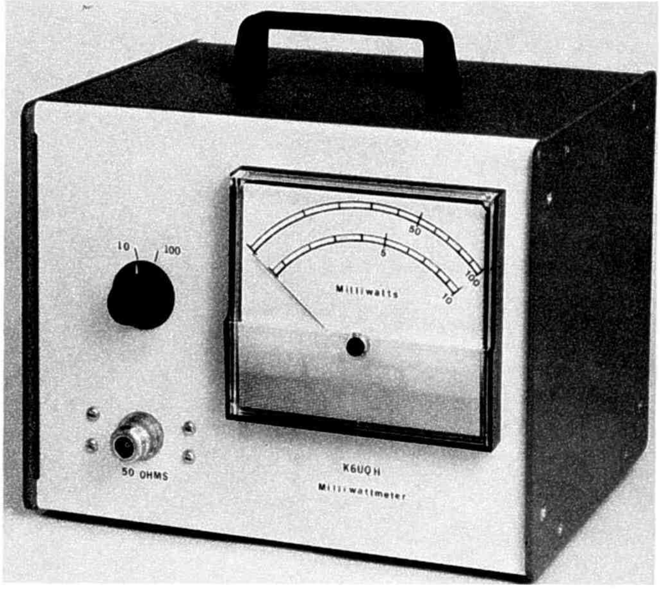


Fig 1—The K6UQH Milliwattmeter shown here is accurate and easy to build. Note the dual-range home-made meter face.

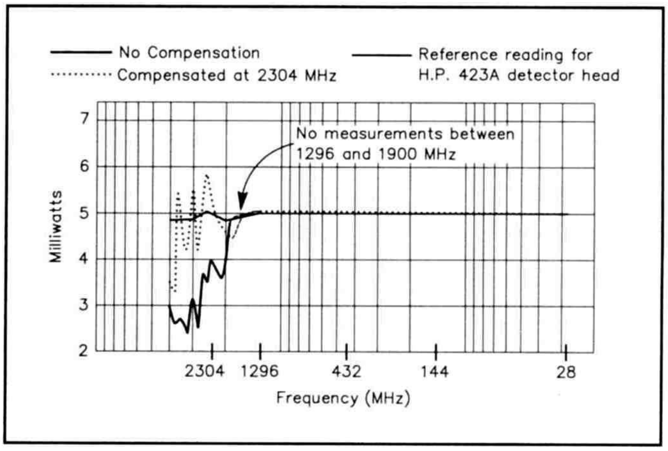


Fig 2—A comparison of the K6UQH Milliwattmeter versus the Hewlett-Packard HP423A. The response of both units is similar up to 1296 MHz.

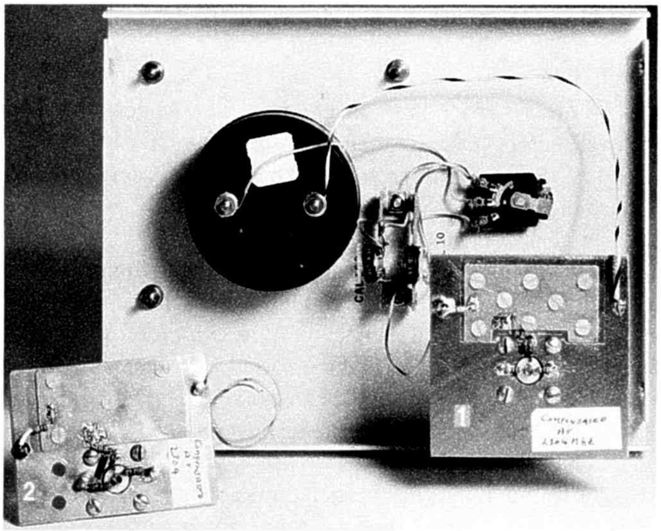


Fig 3—The K6UQH Milliwattmeter can be used with two different detector heads, labeled 1 and 2 in the photo. See text.

anything but simpleminded. In fact, they're quite subtle in terms of behavioral characteristics. The "physics" of the junction type and the diode's physical construction must be considered for RF applications. The junction capacitance and the lead inductance can create unwanted resonances at microwave frequencies. This is one reason why the wattmeter's upper usable limit is set at 1296 MHz (using the hot-carrier diode). Although the input circuit represents a good impedance match at 1296 MHz—it's not perfect, and the input SWR starts to act up in the 2- to 4-GHz region.

Fig 3 shows two different "detector heads." Detector head no. 1 is made with the most readily available parts. Note the circuit-board shim used to keep the Type-N fitting flush with the parts side of the single-sided, ¹⁄₁₆-inch circuit board. Also note the "carry-through" shim strips from the ground plane to the Type-N connector.

The basic Milliwattmeter can also be used to measure levels of stray RF in your shack (up to 1296 MHz). If you see levels greater than 5 mW, you'd better shut down and add improved shielding to your gear, or relocate your antenna to ensure your safety.

Before we get into the nuts and bolts, there are a few time-savers worth mentioning. The internal resistance and current rating of the meter will affect the value of the calibration resistor. Make sure the calibration pot has enough range to accommodate the meter.

Flat-plate capacitors have to be flat—otherwise, regardless of the dielectric material, the measured value of the capacitor will be less than its calculated value. The flat-plate capacitor was around long before the ceramic chip type became available, and still represents a good, low-inductance capacitor.

The 100-Ω input resistors should be mounted flat against the ground plane, with a minimum lead length. This provides a wide-band 50-Ω termination.

Be sure the input resistors are the carbon-film type—not the standard composition type.

Table 1

Detector Head Construction

Detector Head No. 1	*Detector Head No. 2*
Base plate—¹⁄₁₆-in. single-sided glass-epoxy circuit board. Size = 3 × 3-in.	Base plate—Brass plate, 0.05-in.thick. Size = 3 × 2¼-in.
Capacitor plate—Brass plate, 0.02- to 0.032-in. Size = 1 × 2-in.	Capacitor plate—Brass plate, 0.02- to 0.032-in.-thick. Size = 1.2 × 2.3-in.
Input fitting—Type-N.	Input fitting—Type N.

Construction

To build the Milliwattmeter you'll need standard UHF construction tools. Table 1 lists the materials used for the detector heads. Fig 4 shows a typical parts layout for the detector head, and Fig 5 shows the schematic diagram.

The detector heads are designed to provide a flat plate capacitance value of about 800 to 1100 pF, depending upon the type and thickness of your mica stock. In any case, a value of 600 pF or greater seems to work well.

The resistor in series with the diode is quite important. For the hot-carrier diode, its value will range from 39 to 82 Ω. Its effect is most noticeable at the low frequency calibration point.

This resistor will also interact with the calibration resistors, so if you change its value, start the calibration process over. When you find the right resistance value for your meter's diode, the frequency response will be acceptably flat from 28 to 1296 MHz.

Calibration

First, you'll need a power reference standard—usually

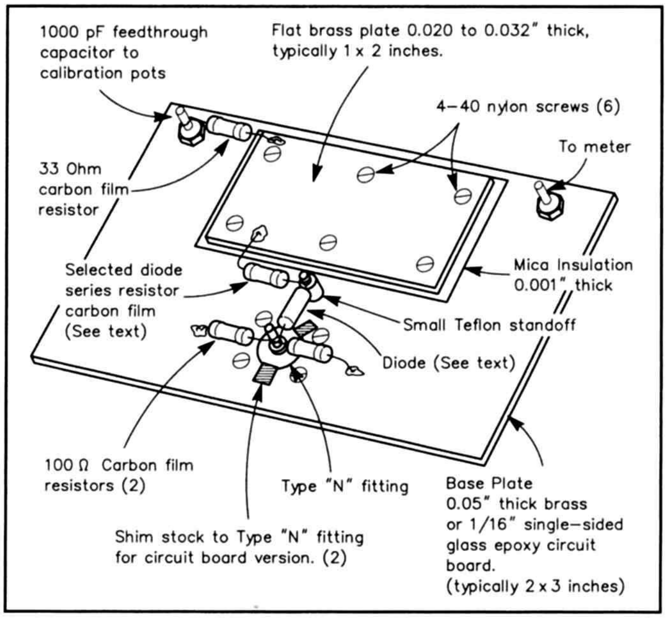


Fig 4—Construction details of a typical detector head. See text and Fig 3.

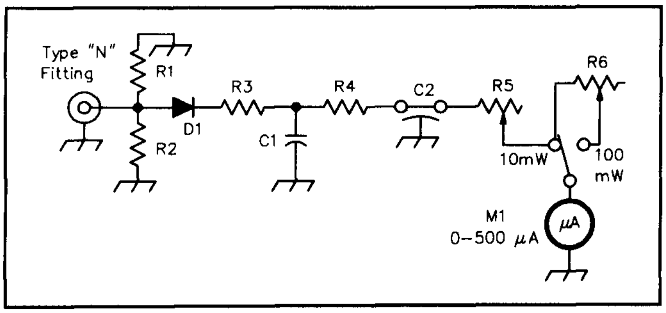


Fig 5—Schematic diagram of the Milliwattmeter. Fixed resistors are carbon film, not carbon composition.

R1, R2—100-Ω.
R3—Selected carbon-film resistor (see text).
R5—5- to 10-kΩ calibration potentiometer (for 10-mW range).
R6—25-kΩ calibration potentiometer for 100-mW range.
D1—HP 5082-2835.
C1—Flat-plate capacitor, approximately 800 pF. See text.
C2—100-pF feedthrough capacitor.

available at surplus stores, flea markets or from a friend. This meter was calibrated at 1296 MHz using an HP-430C power meter. For the dual-range meter, calibrate the 10-mW range first, since its calibration trimmer is in series with the trimmer used for the 100-mW range. All measurements of the detector response are made using a 5-mW variable frequency source, referenced to the meter scale reading.

The frequency sources can be your station's exciters, properly "padded," to provide a 5-mW signal to your power standard (and then switched to your meter). Be sure to use a clean signal for this purpose—use high-Q filters if possible.

A PRD Type S712A Signal Source was used to measure the frequency response from 1900 to 4000 MHz.

Comments

Schottky-barrier, hot-carrier diodes were selected for the Milliwattmeter because of their consistent characteristics from HF to 1296 MHz. Similar results can be obtained using germanium point-contact diodes such as the 1N82AG-C.

Between 1900 and 4000 MHz, the diode junction capacitance, the inductance of the "catwhisker," the load presented to the diode, and the input circuit all have considerable effect on stray resonances that can occur at these high frequencies. If you have the equipment and the patience, you can take advantage of these characteristics to obtain a useful detector above 1296 MHz. A sweep frequency generator will make this job much easier.

The response curve of the detectors above 1296 MHz will vary considerably, depending upon which frequency the meter is optimized for. With no adjustment, the detector response typically falls off rapidly above 1296 MHz, and then rises between 3.5 to 4 GHz (the limit of my measurements). If you choose to make the adjustments, the hot-carrier diode needs more stray capacitance, and the point-contact diode needs less stray capacitance to compensate the detector for the frequency you select.

This is accomplished by moving a capacitive "tab" near the diode series resistor.

By using these adjustment tricks, it's possible to obtain a properly-calibrated response at 2304 or 3456 MHz.

Surplus stores and flea markets are a source of many hard-to-get parts. I use a Micromatch 40-dB throughline directional coupler (rated at 400 watts at 1296 MHz) in conjunction with a low-power, 10-dB attenuator (both flea-market purchases) with the Milliwattmeter to monitor the output power of my 1296-MHz amplifier (typically 400 watts).

1296 Power and SWR Indicator

By Bob Atkins, KA1GT

(From *QST*, Nov 1980)

Many power meters used on the lower bands are unsuitable for use on 1296 MHz. A constant (50-Ω) impedance must be maintained in the transmission line to make meaningful SWR and power measurements. The stripline forward and reverse power indicator shown in Figs 1 and 2 will enable you to make such measurements at low cost. Since all the lines on the PC board are 50-Ω microstrip lines with a width of 2.6 mm, they can be laid out using standard 1⁄10-in. wide (2.54 mm) PC drafting tape as the etch-resistant material, with very little resultant error. This is somewhat easier for most constructors than the use of photoresist etching techniques. You can calibrate this instrument by comparison with a power meter of known accuracy at 1296 MHz, if one is available. The unit may be used as an indicator for tuning up a low-power transmitter (maximum forward power) or for tuning an antenna (minimum reflected power).

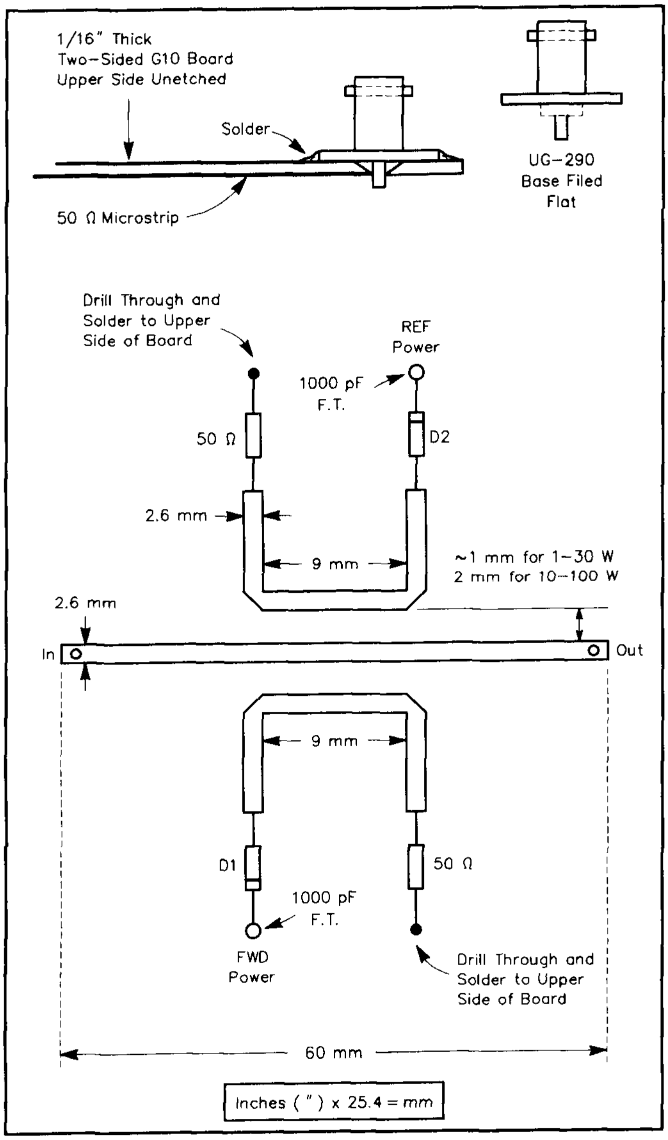


Fig 1— Configuration and design data for a 1296-MHz power and SWR indicator.

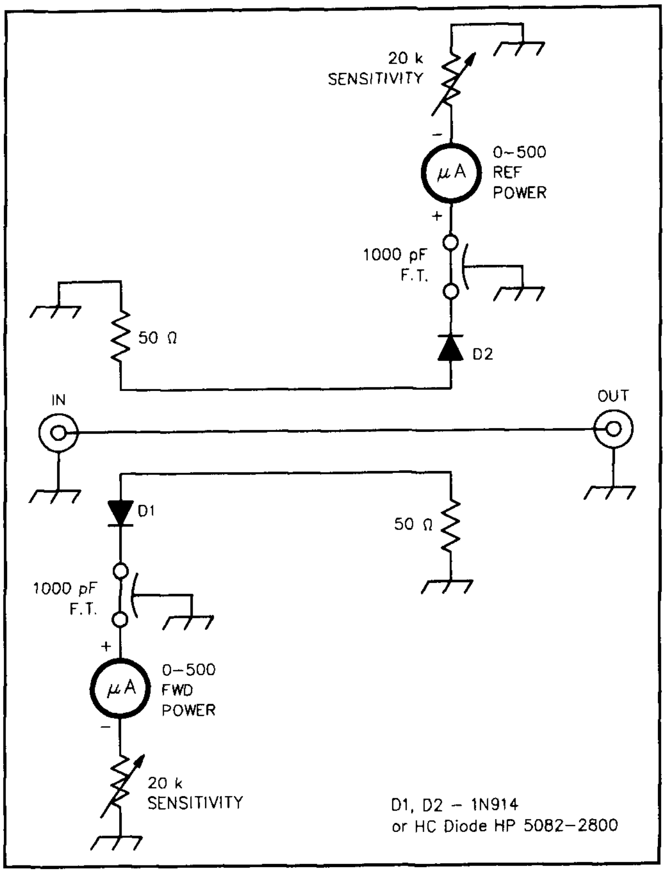


Fig 2— Schematic diagram of a power and SWR meter suitable for use on 1296 MHz.

5.7-GHz Waveguide Probes

By James D. Green, K5JG

A wealth of mathematical and practical data is available on coax-to-waveguide transitions. These usually take the form of a shorted section of waveguide with a coaxial connector and probe. Very little information is available on the insertion of a probe in the main transmission guide, even though they are quite often needed for sampling without disturbing the transmission system. A simple means of sampling the output of your transmitter or microwave test gear to measure frequency or relative power without interfering with the signal being transmitted, would certainly be valuable. The probe arrangements described here will accomplish this handily.

With the trend toward miniaturization, the SMA-type connector was chosen for coaxial connection to the waveguide probe. This offers several advantages; particularly, it presents only a very small obstacle in the main guide. The specific connector used here is an E. F. Johnson type JCM-142-0294-001. The JCM connector is a semi-precision miniature (3-mm) connector fully compatible with all SMA types. When used with semi-rigid coax its performance is satisfactory well beyond 6 GHz. Even though it is gold plated, the price is reasonable. As with most connectors of this type it has a characteristic impedance of 50 Ω.

The object of this project was to accumulate sufficient data that the average amateur microwave experimenter could sample energy from any WR-137 waveguide without the usual expensive "cut and try" approach. The curves shown in Fig 1 were derived by first assembling three short-flanged sections of WR-137 waveguide. Holes to accommodate the coax connector were drilled in each as per spacing indicated. The connectors were measured and averaged before mounting. These measurements are shown in Fig 2. The probe length referred to here is measured from the flat nut portion of the connector to the tip that coax would normally be soldered (see Fig 3).

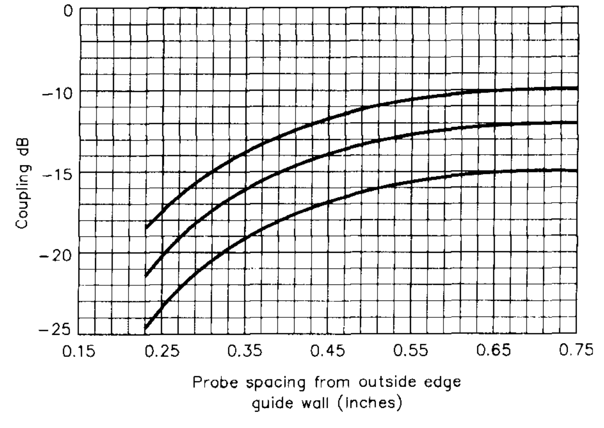


Fig 1—Graph of probe coupling for various spacings and probe lengths.

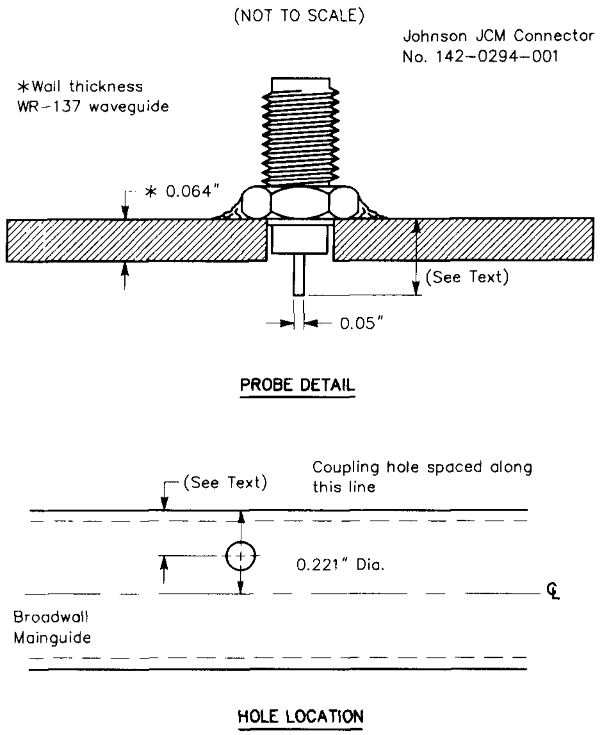


Fig 2—Probe construction and mounting.

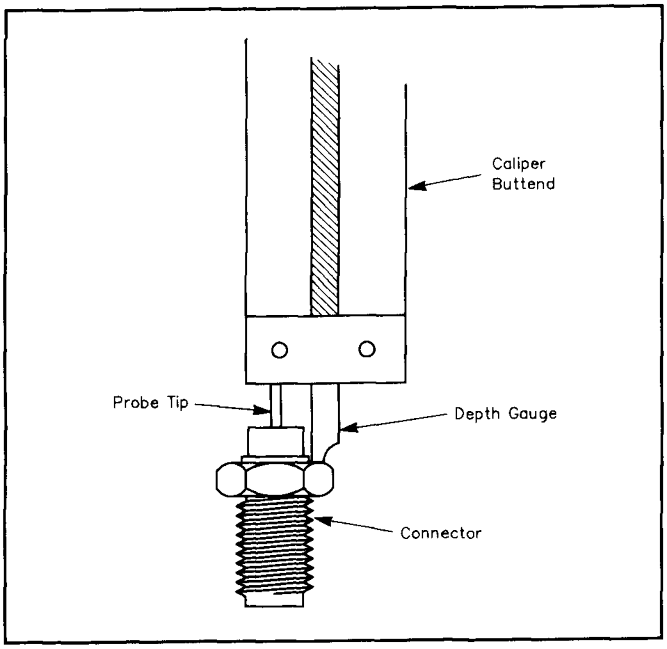


Fig 3—Probe length measurement.

This was done because of ease in measuring this distance with the depth measuring facility of a standard caliper. Each connector was lightly soldered into position with the probe (tip) length as they came from the factory.

Measurements were made on each section for coupling, frequency response of each probe and return loss and frequency response through the main guide. A standard sweep-reflectometer setup was used to conduct these measurements. Connectors were removed and probe lengths shortened successively and measurements conducted for each length and spacing. The results of these measurements are summarized in the curves shown in Fig 1. From these you can decide the spacing and probe length for the coupling desired. For example, if –20 dB coupling is desired, the curves indicate that a spacing of 0.32 in. from the main guide side wall and a probe length of 0.120 in. would be satisfactory. It can also be seen that a probe spacing of 0.25 in. with a probe length of 0.150 in. would also supply –20 dB of coupling.

The probe protrusion into the main guide has much less effect on transmission in the main guide than might be expected. This is mainly due to its small size. A Type-N connector, for example, would have a much greater effect for the same degree of coupling. The worst case is where the longest probe length is used and mounted in the center of the guide (0.75 in. spacing). This effects only a 24-dB return loss across the band. This is equivalent to approximately 1.14:1 SWR. As the probe is shortened and moved toward the side wall, the return loss gets even better. Frequency response across the band is rather good, exhibiting no more than 1.0-dB variation regardless of the combination used.

Construction

Location of the probe in line with the long dimension of the main guide is unimportant and may be placed at the most convenient point. After this location is determined measure and scribe the location of the probe along the perpendicular axis by measuring with a caliper from the outside wall toward the guide center. This hole should be marked with a center drill and then drilled with a no. 2 (0.221-in. diam.) drill. Clear the hole of any burrs.

If some dimension other than the normal length of the connector is desired, file the tip carefully. Hold the connector with the threaded portion down and probe tip up. Seat the butt end of the caliper on the probe tip. Run the sliding arm of the caliper depth gauge down until it barely touches the flat side of the nut-shaped portion of the connector (Fig 3). This will give you the probe length. File only a small portion of the probe at one time and remeasure. The tip is small, delicate and easy to over file. When the probe length is correct the connector may be soldered into the main guide. Use only soft solder and do not apply heat directly to the connector body. Heat applied to the waveguide will be conducted sufficiently to the connector for good solder flow. Clean up any flux residue and the probe is ready for use.

Conclusions

The probe provides one of the simplest methods of allowing one to check the frequency or power of a transmitter while it is in operation without disturbing the transmitted signal. Determine what level RF is required. By using the data supplied here a simple probe can be installed that will do the job without the usual "cut and try" approach. If it is desired to use more than one probe in the main guide this should not present any problem.

Waveguide Loads

By Kent Britain, WA5VJB

(From *Proceedings of Microwave Update '89*)

Simple, quick, and common stuff: that was the idea. I went through about two dozen designs and these are the two that worked best. The absorber is that black foam stuff you store ICs in. The foam conductivity varies quite a bit. Put your ohmmeter probes on the foam, about an inch apart. *H*ighly conductive (approx 10 kΩ), *M*edium (approx 50 kΩ and *L*ow/Static Dissipative (>200 kΩ). This carbon-loaded foam usually comes in thin sheets, so you may have to build the load up with several layers.

My commercial loads showed from 32- to 45-dB return loss so these foam versions worked pretty well. For the purist, Eccosorb worked best, but isn't commonly available. Finally, the lowest SWR occurs if you leave the back end of the waveguide open, but there will be some leakage.

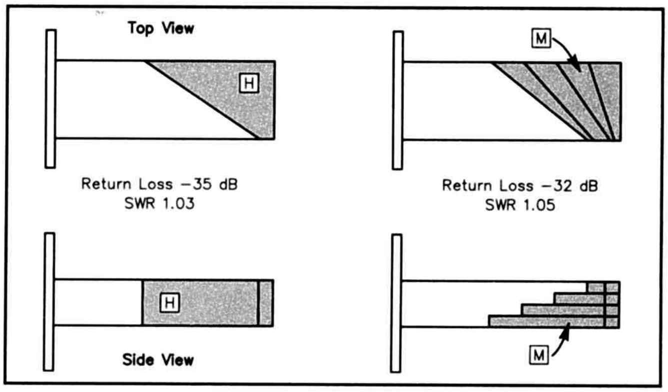


Fig 1—Various methods of making waveguide loads from conductive foam. H = Highly conductive, M = Medium conductive, L = Low conductive (see text).

A Single-Knob, Single-Crystal VHF/Microwave Calibrator

By Bill Troetschel, K6UQH

This unit, shown in Fig 1, provides a highly accurate set of RF signals from VLF to the microwave spectrum—directly referenced to WWV. It can be used to calibrate and align HF or VHF receivers and provide accurate reference signals in the VHF-microwave spectrum—simultaneously.

Error sources that normally affect the accuracy of frequency measurements are typically:

- Basic receiver calibration and tracking
- Crystal oscillators in converters
- General confusion in terms of the myriad ways receivers and transmitters internally use "offset" oscillators for various modulation modes.
- The inability of the human ear to hear a true subaudible zero-beat signal.

The first three items generally represent "fixed" conditions, while the last illustrates the flaw in the general technique used to identify the others.

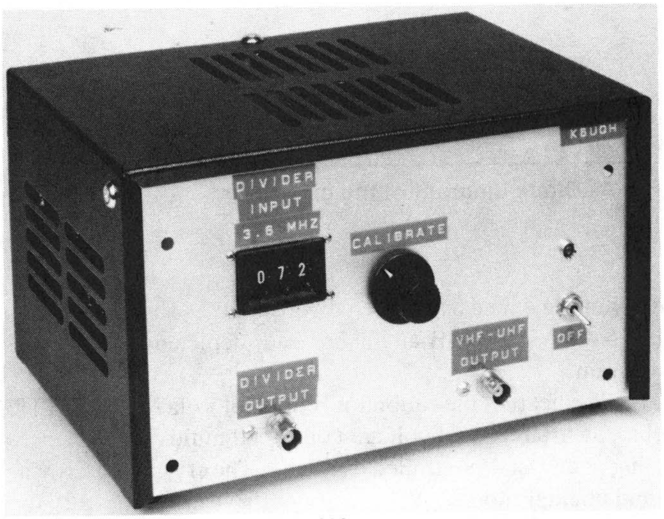
Nearly all amateur receivers have very little audio response below 100 Hz, and few people have good audio perception below 20 Hz. Therein lies the problem with obtaining an accurate zero beat.

Most hams use a transfer oscillator to zero-beat WWV, and then use its harmonic to zero beat at the desired receive frequency. A few hams have "error-free" frequency counters calibrated to WWV. The problem with either approach is that an audible WWV zero beat is generally a "dead space" on the dial, typically 30 Hz wide. Assuming you use a 10-MHz WWV signal to calibrate a 2304-MHz signal, a 20-Hz error at the fundamental means an error greater than 4.6 kHz at the harmonic. The visual method described in this article is much more accurate.

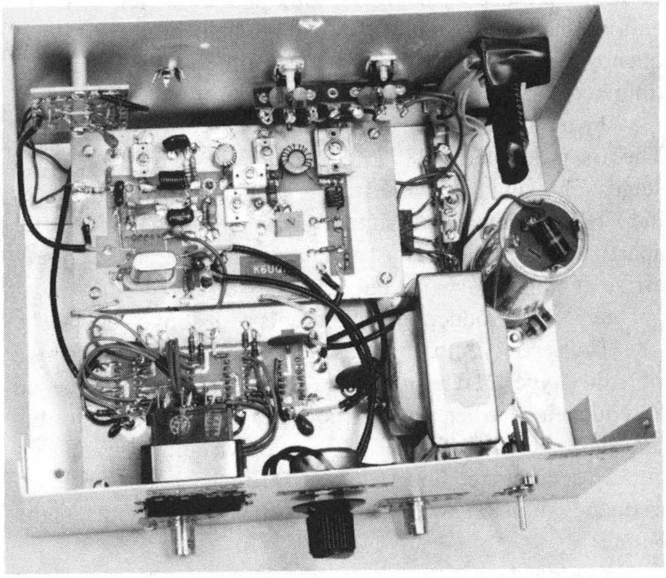
The Single-Knob, Single-Crystal Calibrator provides an easy-to-adjust signal with visual subaudible zero-beat indicator. (You watch the receiver S meter as the WWV signal and the calibrator signal combine in the receiver.)

Circuit Description

A block diagram of the calibrator, and one possible application are shown in Figs 2 and 3. The unit is shown schemati-



(A)



(B)

Fig 1—At A, a front-panel view of the Single-Knob, Single-Crystal Calibrator. At B, an interior view shows the placement of the three circuit boards and the power supply.

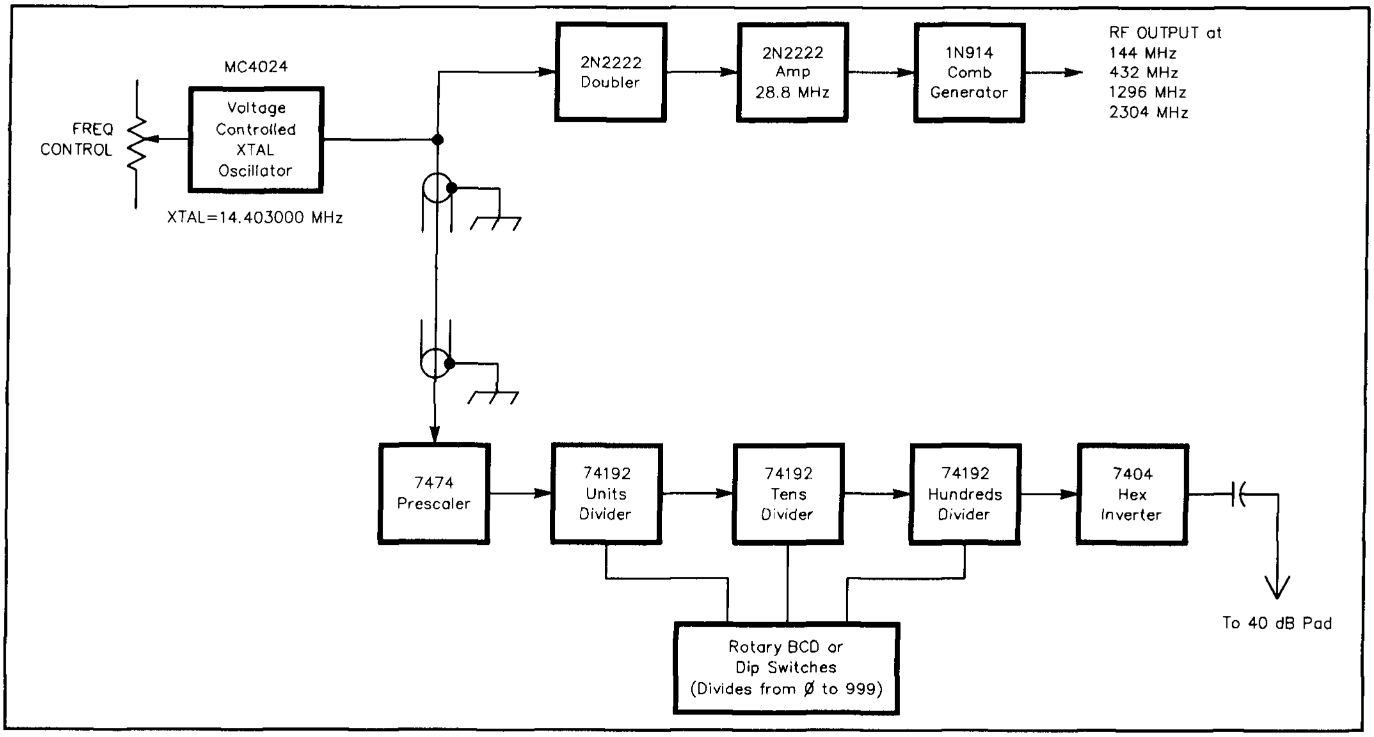


Fig 2—Block diagram of the calibrator.

cally in Fig 4, and a suitable power supply is shown in Fig 5. Fig 6 shows a 40-dB attenuator pad, depicted in the block diagram.

The heart of the calibrator is the Motorola MC4024 TTL chip, an astable, dual voltage-controlled multivibrator, operating as a crystal-controlled oscillator. The crystal is a Croven fundamental-mode, AT-cut, high-drive crystal cut for 14.403000 MHz. It provides a rock-solid output at 14.400000 MHz when tuned by the calibration potentiometer. This circuit is designed to oscillate far enough below the crystal frequency to provide mid-range tuning for the calibration potentiometer.

Two outputs are taken from the oscillator at 14.4 MHz. The first is doubled to 28.8 MHz and amplified by the 2N2222 stages. This signal is applied to a comb generator that provides the VHF/microwave output (at frequencies spaced 28.8 MHz apart). A 1N914 diode is used in the comb generator. A more-efficient step-recovery diode is discussed in the references, but the 1N914 produces useful outputs well above 2304 MHz.

The second output is prescaled by a factor of four by a 7474 dual type-D edge-triggered flip-flop. This stage is needed to reduce the divider's input frequency to compensate for the time delays incurred in the 74192 up/down counters, and provide an accurate reading on the BCD (binary-coded decimal) frequency division switches. A 7404 hex inverter buffers the output.

By selecting the proper division ratio on the BCD switches (described later), many VLF signals and their harmonics can be selected for calibrating your HF/VHF receiver, and for referencing the WWV signal available in your area. For example: 14.4 / 4 = 3.6 MHz. This 3.6-MHz signal, divided by 144 provides 25-kHz reference markers, plus the usual WWV markers at 2.5, 5, 10, 15 and 20 MHz. Other useful divider ratios are 36 (100 kHz), 72 (50 kHz) and 360 (10 kHz). To determine the harmonic interval of the marker generator, use the formula

3.6 MHz / 3-digit BCD display × 1000 = marker frequency (kHz)

The output of the 7404 hex inverter is quite strong, so a 40-dB attenuator attenuates the signal level to make its harmonics approximately equal in strength to the WWV signal.

Construction

Point-to-point wiring or a PC board are equally suitable for this circuit. I built mine on a PC board. PC board templates and part-placement diagrams are given in Figs 7 through 12. Perhaps the most tedious part of the job is wiring the BCD switches. Remember that the first divider is for units, the second for tens and the third for hundreds. When mounting the BCD switches on the front panel, be sure they read hundreds-tens-units. The BCD switch is wired as a 5-V common system, and *not* as a BCD complement (where the common is ground).

Construct the comb generator circuit carefully in terms of layout and lead dress. Remember, the RF output frequency can be as high as 10,368 MHz if you are careful.

Alignment

There is no question but that the best way to tune up a comb generator is with a spectrum analyzer. By using your receivers as signal-strength indicators, however, you can do an acceptable job. I assume you can find your frequency to within 14 MHz. Since the basic oscillator frequency is 14.4 MHz, there are several unwanted frequency spikes spaced 14.4 MHz apart in the comb, particularly at the lower frequen-

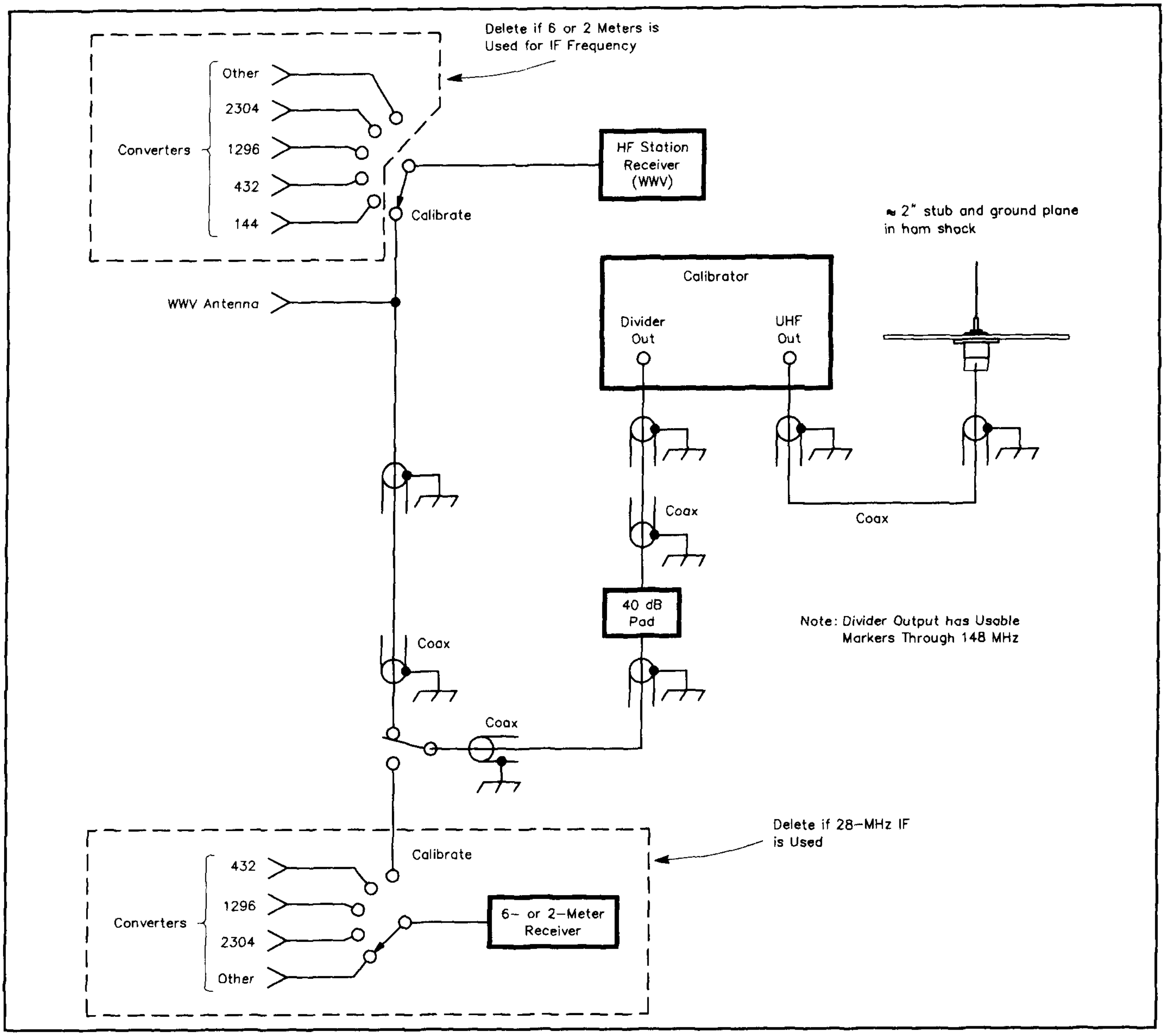


Fig 3—Here's one application of the calibrator. See text for more information.

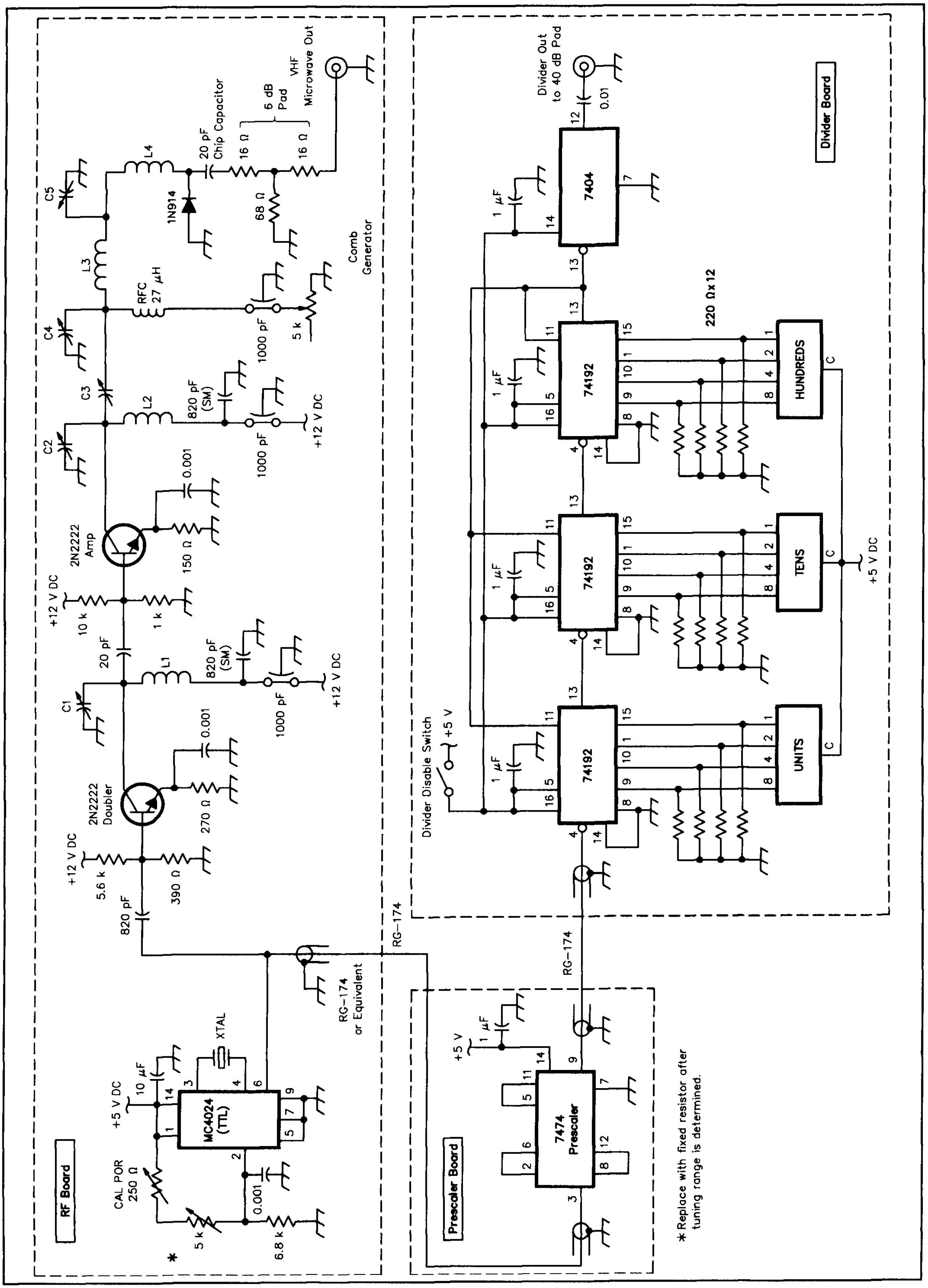
cies. You can minimize their effects considerably by tuning with the aid of a spectrum analyzer, but the spurs shouldn't be a problem if you use high-quality microwave receivers as tune-up indicators.

The first step in alignment without a spectrum analyzer is to set the 1N914 bias pot to its maximum value. Monitor the voltage across the pot with a sensitive voltmeter, and adjust the doubler and amplifier stages for maximum voltage—approximately 2 to 3 V. Turn the bias pot to approximately zero resistance. At this point, connect a 2-in. stub antenna to the VHF/microwave output. You should be able to hear the calibration signal on your receivers. Now, use your station receivers, preferably the highest frequency receiver available, and adjust the other controls for maximum signal strength—and you're done.

If your wattmeter has a 10-mW range, an alternative tune-up procedure is useful. Set the diode bias pot as above, and tune the entire unit for a maximum power indication. Output should be between 3 and 5 mW, which represents the summation of all harmonics from that port. Set the bias pot to zero resistance. The power output should drop to about 2 mW, which indicates proper operation.

Using the Calibrator

To calibrate an HF receiver, first tune in WWV at any frequency on which reception is good. Then couple the output of the decade dividers to the receiver antenna input, through the 40-dB attenuator pad. I use a coaxial tee connector at the receiver antenna jack. Set the BCD switches to provide the desired output markers. Now, carefully tune the calibration potentiometer while watching the receiver S meter. When the needle starts to flop around, you know you're getting close to



* Replace with fixed resistor after tuning range is determined.

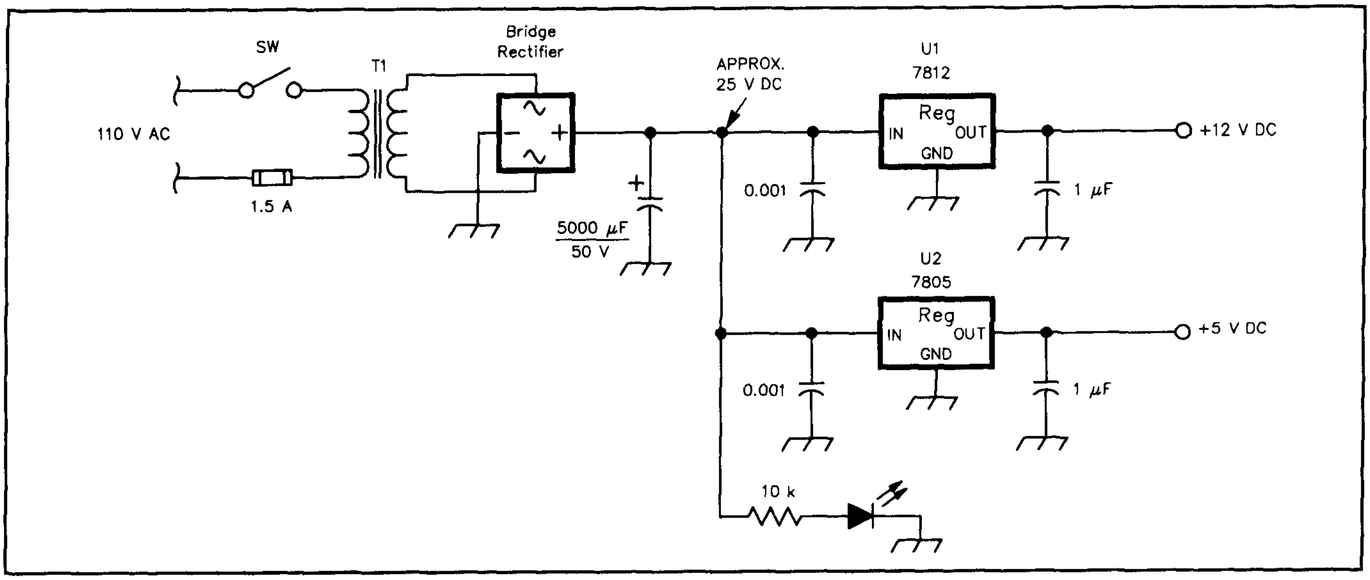


Fig 5—Schematic diagram of a power supply suitable for use with the calibrator.

D1—Bridge rectifier 100 PIV, 4 A.	**U1—Regulator IC, +12 V, 1 A, 7812 or equiv.**
D2—Red LED.	**U2—Regulator IC, +5 V, 1 A, 7805 or equiv.**
T1—Power transformer, 18-VAC, 2-A.	

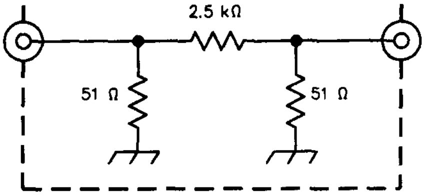


Fig 6—Schematic diagram of the 40-dB attenuator, used to make the calibrator signal strength equal to the received strength of WWV.

←

Fig 4—Schematic diagram of the calibrator.

C1, C2, C4—Trimmer capacitor, 4 to 40 pF.
C3—Trimmer capacitor, 1.5 to 20 pF.
C5—Trimmer capacitor, 30 to 450 pF.
L1—12 t no. 20, ¼-in. ID, air-wound, close-spaced.
L2—15 t no. 26 on T37-6 core.
L3—13 t no. 26 on T50-6 core.
L4—5 t no. 20, ¼-in. ID,

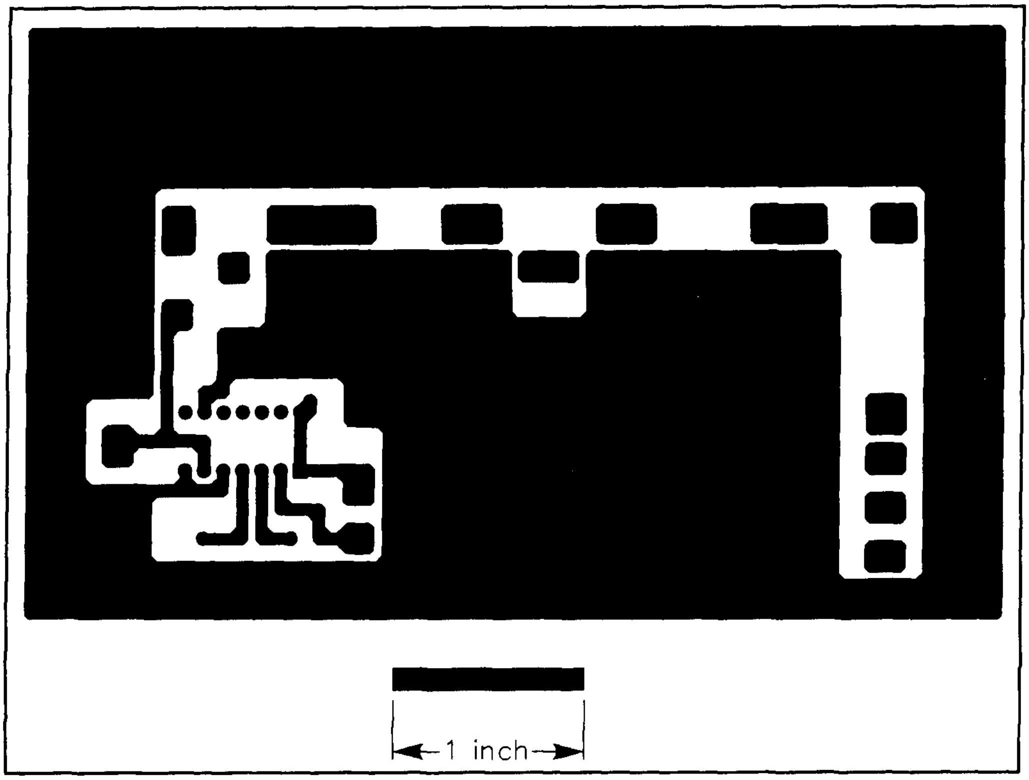


Fig 7—RF board template.

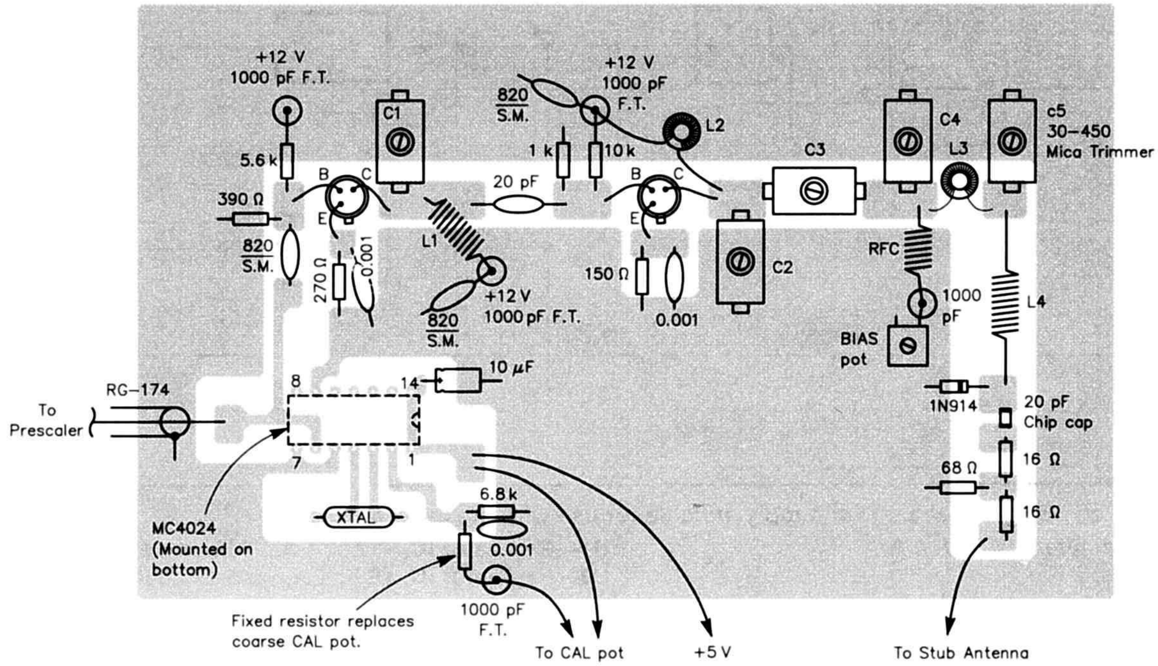


Fig 8—RF board part placement. All components *except* the IC are mounted on the foil side.

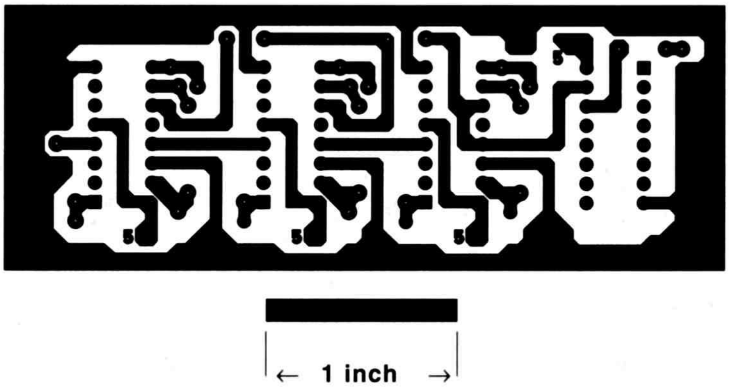


Fig 9—Decade-divider board template.

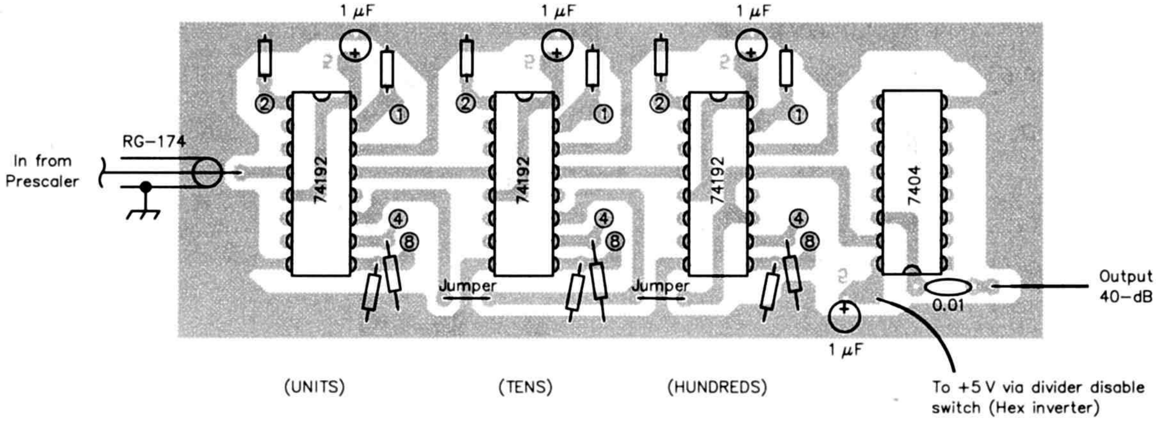


Fig 10—Decade-divider board part placement. All points labeled "5" are connected by jumpers.

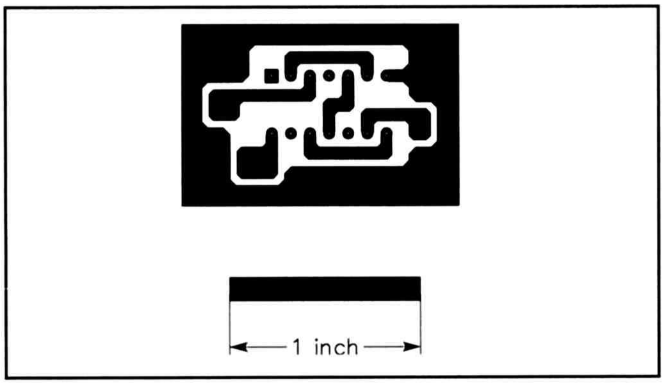


Fig 11—Prescaler board template.

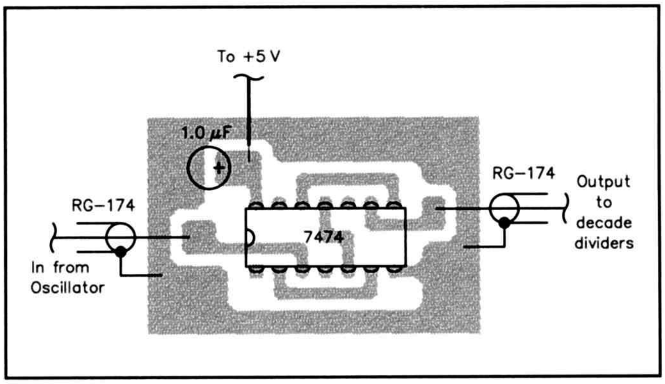


Fig 12—Prescaler board part placement.

zero beat. When the S meter fluctuates at 1 Hz or less, you've achieved true zero beat with WWV. You'll be able to hear the beating effects on your receiver as you adjust the calibration pot.

Once the oscillator has been calibrated against WWV, the output from its microwave port is also extremely accurate, and can be used as a frequency standard well into the microwave region. If you use the 20-MHz WWV signal and adjust the calibrator so the beat note is less than 1 Hz, the error at 1296 MHz is only about 65 Hz. You can get greater accuracy by adjusting the calibration pot until the S-meter pointer is nearly stationary for a long period, say 15 seconds. The divider on-off switch removes the 5-V supply from the dividers to eliminate the VLF output and its harmonics when only VHF/UHF output is desired.

If you use 6 or 2 meters as a tunable IF, first calibrate the unit against WWV, then use the divider output to check the calibration and tracking of your VHF receiver. Harmonics of the divider will show up as markers, up to at least 148 MHz.

Conclusion

Scheduled long-haul microwave tropo DX is made easier when you know you're on the right frequency. With GaAsFET amplifiers at the antenna, I receive the calibrator signal weakly on 1296 and 2304 MHz. For the higher frequencies you may need more direct coupling, unless you have terrific preamps. You'll find that the calibrator signal is remarkably stable—perhaps more so than some signals you are used to receiving!

Other crystal frequencies, prescaling integrals and more amplification prior to the comb generator will improve or extend the unit's capability to other bands, such as 50 or 222 MHz.