

# An HF In-Line Return Loss And Power Meter

Adjust your antenna tuner with one mW of RF and then measure up to 300 watts, or measure the passband of that filter you just built -- all with the same instrument!

Simplicity in some designs sacrifices performance. In this case, simplicity is only external and no performance is lost. Through the use of a high performance and internally sophisticated integrated circuit, an accurate HF in-line return loss and forward power meter is possible over an 80 dB input power range with four integrated circuits.

## Measurement Overview

A block diagram for this instrument is shown in Figure 1. The in-line directional couplers provide RF power samples,  $P_{fs}$  and  $P_{rs}$ , that are 40 dB down from the corresponding forward and reflected mainline powers,  $P_f$  and  $P_r$ . The RF samples are not rectified, as done in many SWR meters, but applied directly to the log amplifiers, A1 and A2. The log amp is *demodulating*, in that it converts the RF input voltage to a DC output voltage that is linearly related to the logarithm of the RF input voltage.

The integrated circuit used here for the log amps is the Analog Devices AD8307<sup>1</sup>. Its typical conversion accuracy is +/-0.3 dB over an 80 dB input power range. This range extends from about -95 dBW to -15 dBW at the log amp input. Accounting for the 40 dB coupling factor in the directional couplers, the log amp input range corresponds to a mainline power range of -55 dBW to +25 dBW. Converting from dBW to watts results in a measured mainline power range of 3.16 microwatts to 316 watts, accurate to about +/-7%!

The high impedance log amp outputs are buffered by unity gain amplifiers, A3 and A4, to eliminate errors caused by loading of the subsequent stages. After calibration, the mainline forward and reflected powers in dBW are related to the unity gain amplifier outputs by:

$$P_{f,dBW} = K_1 \times V_{fl} + K_2 \quad (1)$$

and

$$P_{r,dBW} = K_1 \times V_{rl} + K_2 \quad (2)$$

Calibration is required to obtain near equality of like constants  $K_1$  and  $K_2$  in the forward and reflected paths. After calibration, we can measure DC voltages and relate them to mainline forward and reflected power, in dBW or watts, over the 80 dB range.

Instead of measuring SWR directly, we can simply measure return loss in dB, RL, by taking the difference of the log amp outputs<sup>2</sup>:

$$RL = P_{f,dBW} - P_{r,dBW} \\ = (V_{fl} - V_{rl}) \times K_1 \quad (3)$$

In this design,  $K_1$  is set to a nominal value of 50 dB/V. The above difference can be taken externally using a single floating input digital voltmeter for enhanced precision or internally using the A6 amplifier, which drives the return loss meter, M2.

The DC forward voltage and an offset voltage are processed through the A5 amplifier which provides the needed gain and offset to drive the forward power meter, M1. The offset voltage allows the user to select the desired power range to be displayed on the forward power meter.

All circuits are powered by a single low dropout positive voltage regulator to facilitate battery operation. The combination of powering, the op amp input polarities and gains, and the ground referenced analog meters provides meter protection.

Meter disabling, used in some designs to eliminate spurious or erratic readings when RF is not present, is also not needed here since the two log amps typically have nearly identical outputs without RF.

Analog meters are used for the primary display to facilitate adjustment of antenna tuners. Watching a needle swing to a peak or a null rather than reading digits on a DVM is usually easier for most users. Both meters are calibrated using linear dB scales for both return loss and forward power. Each swings upward with increased return loss and forward power.

While the analog meters are sufficient for many uses, the auxiliary metering port and a DVM can be used to more

accurately measure the entire -55 dBW to +25 dBW (3  $\mu$ W to 320 W) range for both forward and reflected power as well as return loss. Since numerical values for  $K_1$  and  $K_2$  can be determined after calibration, it is a simple matter to obtain accurate values for forward power, reflected power, and return loss using the measured DVM voltages and a calculator. Figure 1 shows the DVM connections required to directly measure forward power, reflected power, and return loss.

## Return Loss and SWR

The key features of Figure 1 are simplicity, accuracy, wide power range, and independence of measured return loss from forward power. This last feature is particularly striking when transmitting SSB where the forward power meter continually fluctuates and the return loss meter barely moves.

Most radio amateurs use meters that measure the voltage standing wave ratio, VSWR, which is generally simply called SWR. This instrument breaks away from that trend by measuring and displaying return loss in place of SWR. Figure 2 provides a conversion between return loss and SWR, using this instrument's 0 to 30 dB return loss scale.

Most amateurs will want return loss higher than 10 dB (1.9:1 SWR) in normal operation and higher than 20 dB (1.2:1 SWR) in some cases. As you can see, you adjust your antenna tuner for a return loss peak rather than a SWR null. But that will take maybe 5 minutes to get used to.

The rationale for limiting maximum displayed return loss to 30 dB (1.07:1 SWR) is twofold. The first is to minimize the forward power required to adjust an antenna tuner and still be able to measure maximum displayed return loss, in this case 30 dB. Using -55 dBW for the minimum measurable reflected power and 30 dB return loss results in a forward power of -25 dBW, or equivalently, 3 mW. In actual practice, 1 mW of forward power has been routinely used since the log conformance does not deviate too abruptly at the low end.

The second reason for the 30 dB limit is that the error in measured return loss increases with return loss, as will be seen

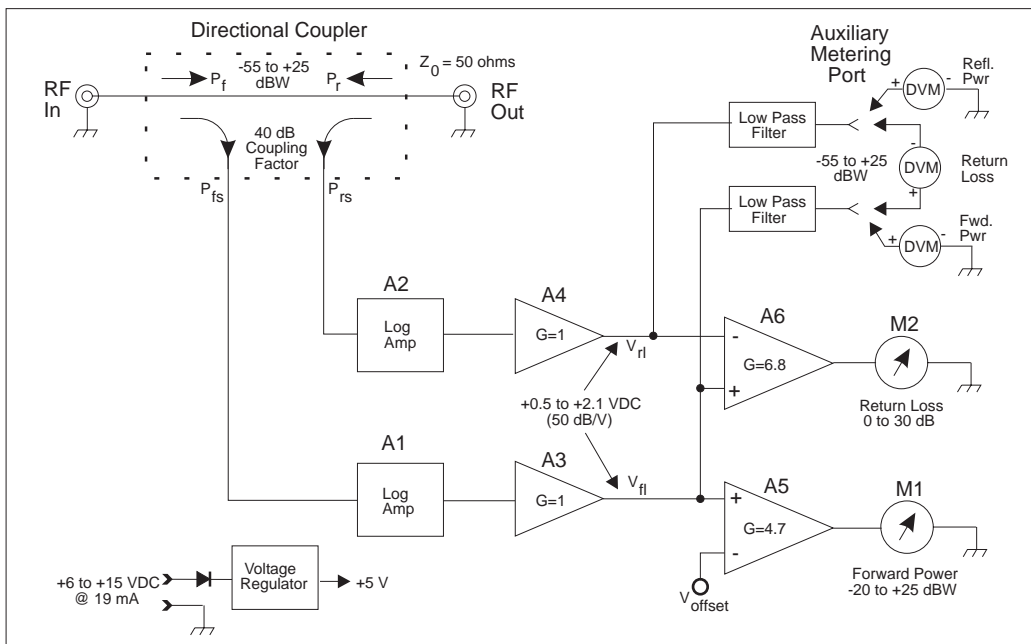


Figure 1 - Block Diagram of HF In-Line Return Loss and Power Meter (3  $\mu$ W to 300 Watts)

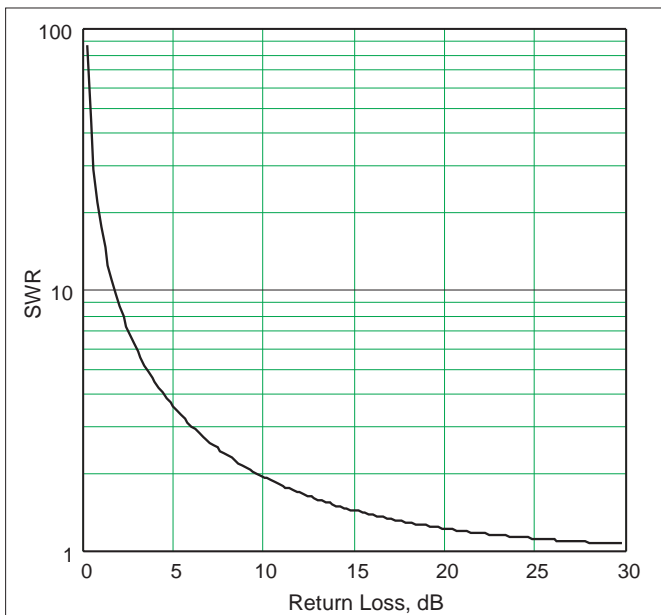


Figure 2 - Convert Return Loss to SWR

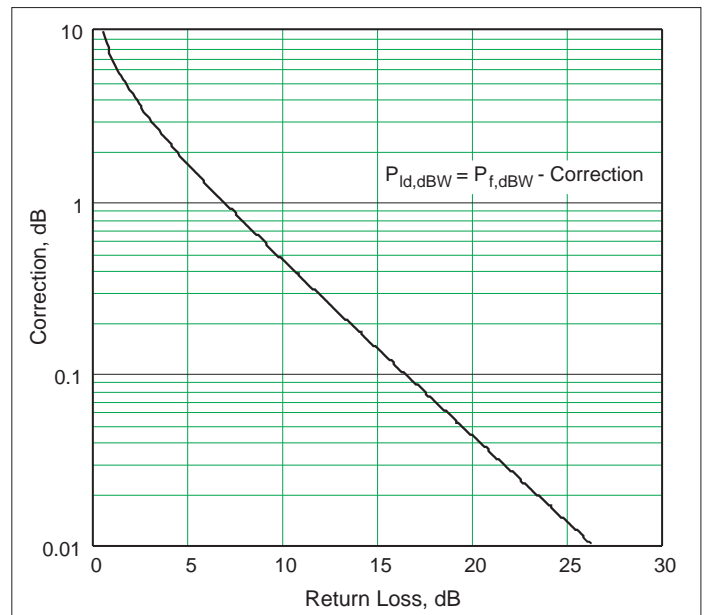


Figure 3 - Load Power Correction Using Measured Forward Power and Return Loss

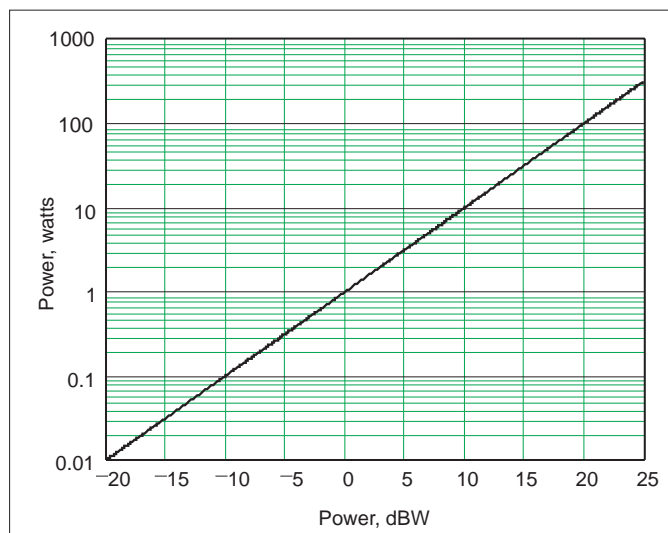


Figure 4 - Convert Power in dBW to watts

later. This limit represents a trade-off between increasing measurement error and diminishing returns in efficiency.

A forward power range of 1 mW to over 300 watts, while maintaining good return loss accuracy, makes this instrument compatible with a wide variety of available RF sources.

#### Forward Power vs. Load Power

This instrument measures forward power instead of load power. For high values of return loss, forward power can be used to approximate load power. Using the measured return loss and forward power, Figure 3 shows a correction factor in dB that can be subtracted from the measured forward power to get load power.

The error in using forward power for load power, without any correction, is less than 0.5 dB for return loss higher than 10 dB (1.9:1 SWR) and less than 0.05 dB for return loss higher than 20 dB (1.2:1 SWR). The added circuit complexity required to display load power instead of forward power is not warranted during normal operation. If needed, accuracy can be improved using either Figure 3 or the formulas in the sidebar.

This instrument displays forward power in dBW instead of watts. Figure 4 provides a conversion between the two. Using power in dBW instead of watts is useful in that it more accurately conveys the impact of making power changes. For example, going from the typical transceiver output of +20 dBW (100 watts) to a QRP level of +7 dBW (5 watts) is a 13 dB decrease, or a little more than two S-units<sup>3</sup>. The difference seems smaller expressed that way and makes you want to give QRP a try, doesn't it?

#### Directional Couplers

Several directional coupler and bridge circuits were studied for possible use in this instrument. The choices were narrowed to two circuits: the coupler used in the Tandem Match<sup>4</sup> and variations of the popular Bruene bridge<sup>5</sup>. The Tandem Match coupler is shown in Figure 5(A) and one variation of the Bruene bridge is shown in Figure 5(B).

The Tandem Match coupler has some nice features: simplicity, excellent directivity, scalable to other power levels, and 50-Ω load impedances on all ports. The Tandem Match coupler was built and tested.

While directivity<sup>6</sup> exceeds 40 dB on all HF amateur bands and coupling factor is flat to within +/-0.35 dB, it does have one disappointing characteristic for general HF use. Figure 6 shows the

return loss measured at the *input* of the Tandem Match coupler with a 50 dB return loss termination at its output. The input return loss for the Tandem Match coupler is about 30 dB or better above 7 MHz, but steadily degrades to only 18 dB at 1.8 MHz, which is equivalent to an input SWR of 1.3:1.

So, while the Tandem Match coupler is capable of accurately sensing low SWR on the transmission line connected to its output, it does not do as well in presenting a low SWR load on the lower amateur bands to the transmitter. Surprisingly, no explicit mention has been previously made of this characteristic; perhaps, no one else has thought about measuring the SWR of the SWR meter!

While improving this characteristic through design changes was not exhaustively explored, it did appear that improving low frequency input return loss would likely result in reduced high frequency directivity<sup>7</sup>.

The Bruene bridge, in various forms, has been used over the years in many homebrew and commercial in-line HF SWR meters ranging from QRP to QRO capability. Input return loss is usually not a problem as long as the coupling factor in dB is high.

While there are obstacles to achieving good directivity, it can be done. Referring to Figure 5(B), the main obstacles to good directivity are a) parasitic lead inductance associated with C2, b) high values for C2, c) excessive secondary wire length on T1, and d) impedance control in the bifilar secondary winding. The lead inductance and C2 result in a series resonance that progressively deteriorates bridge balance as the frequency is raised. Figure 7 shows the deterioration in 28 MHz directivity with increased parasitic lead inductance in an otherwise ideal Bruene bridge.

The Bruene bridge used in this meter, including the log amps, was built and tested. Figure 8 shows the input return loss, directivity, and coupling factor. Some key characteristics are:

- input return loss exceeds 50 dB<sup>8</sup> for 1.8 - 30 MHz (41 dB at 54 MHz),
- directivity exceeds 40 dB for 1.8 - 30 MHz (33 dB at 54 MHz), and
- the coupling factor is flat to within +/- 0.4 dB for 1.8 - 54 MHz.

While the Bruene bridge is more complex, requires adjustment, and loses the 50-Ω termination feature when compared to the Tandem Match coupler, it was decided that the improved input return loss overcame those disadvantages.

#### Circuit Description

The complete circuit is shown in Figure 9. Much of the circuit has been described in Measurement Overview and is supplemented here.

C1 provides the coarse bridge balance while C6 provides the fine adjustment. Four paralleled capacitors are used at C2 through C5 to provide low parasitic inductance. C2, C3, and C4 are chip capacitors that are mounted between the plates of C5. Refer to the next section for construction details.

The parallel combination of R1 and the log amp common mode input impedances extends the low frequency directivity of the Bruene bridge. It was necessary to keep T1 secondary inductance low and its total load resistance high to accommodate the relatively low common mode impedance of the log amps while maintaining high frequency directivity.

To accommodate the above characteristics and to maintain bridge directivity as high as possible, a modification to the Bruene bridge is made. The single resistor normally across the transformer secondary is split into three resistors, R2, R3, and R4. This provides reduced amplitude signals to the log amps and better impedance matching to the bifilar wound transformer secondary.

Each of the voltage and current senses has a nominal 46 dB attenuation. When the appropriate phases are summed for the forward signal, this amounts to a nominal 40 dB coupling factor. The difference provides the reflected signal. Ideally, the difference would be zero (infinite directivity in dB), but various imperfections limit the actual directivity. The differential inputs provided by the log amps are used to form the required sums and differences involving the capacitive voltage divider and the appropriate phases of the inductive current sense.

Pots R10 and R12 at the log amp outputs provide the adjustment of each gain to the nominal 50 dB/V. Pot R6 provides the offset adjustment to set the indicated reflected power equal to the indicated forward power when the actual return loss is 0 dB, an open or a short for example.

The U3A and U3B unity gain buffers are required to be nearly negative rail compatible since the log amp outputs range from about 0.17 V to 2.1 V. The auxiliary external metering port at J3 is filtered to protect the unity gain buffers against inadvertent shorts and reduce the effects of external noise. The log amp data sheet recommends that care be

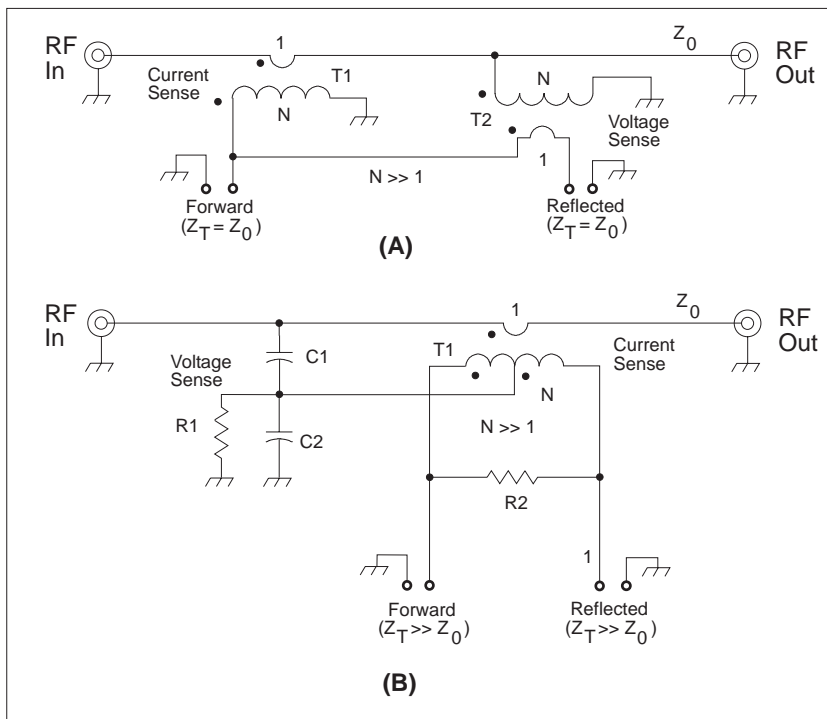


Figure 5 - Directional Couplers (A) Tandem Match Coupler (B) Bruene Bridge (Phase Compensated)

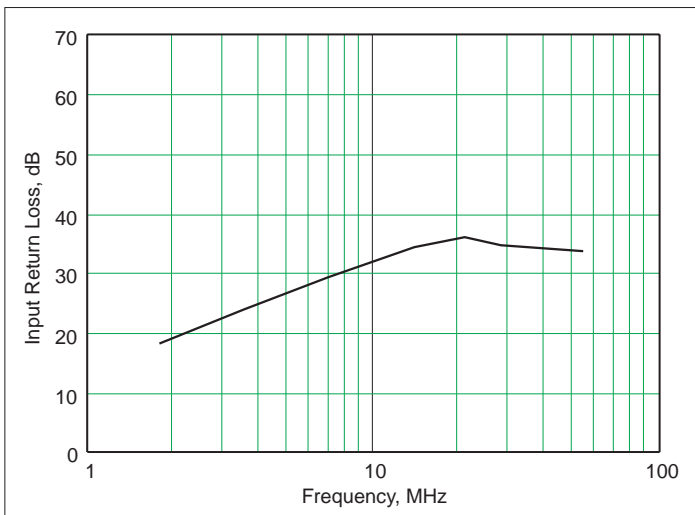


Figure 6 - Measured Tandem Match Coupler Input Return Loss

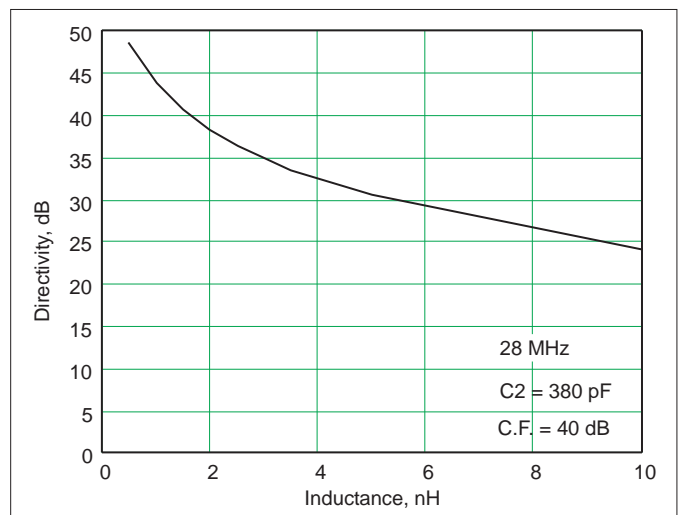


Figure 7 - Bruene Bridge Directivity Sensitivity to Lead Inductance

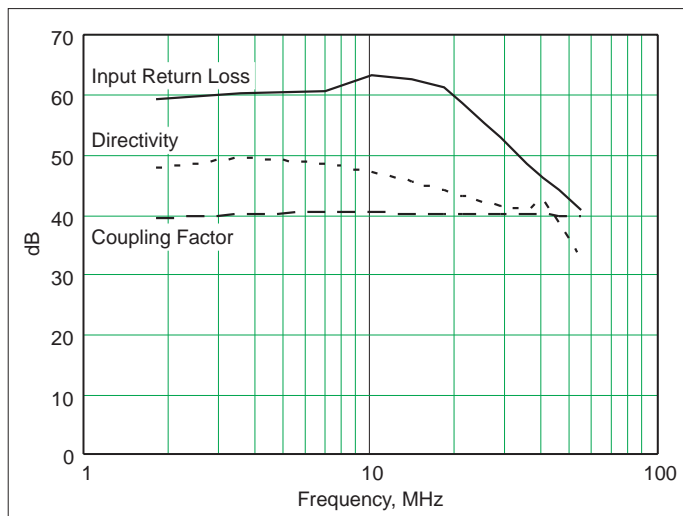


Figure 8 - Measured Bruene Bridge Characteristics (Including log amps)

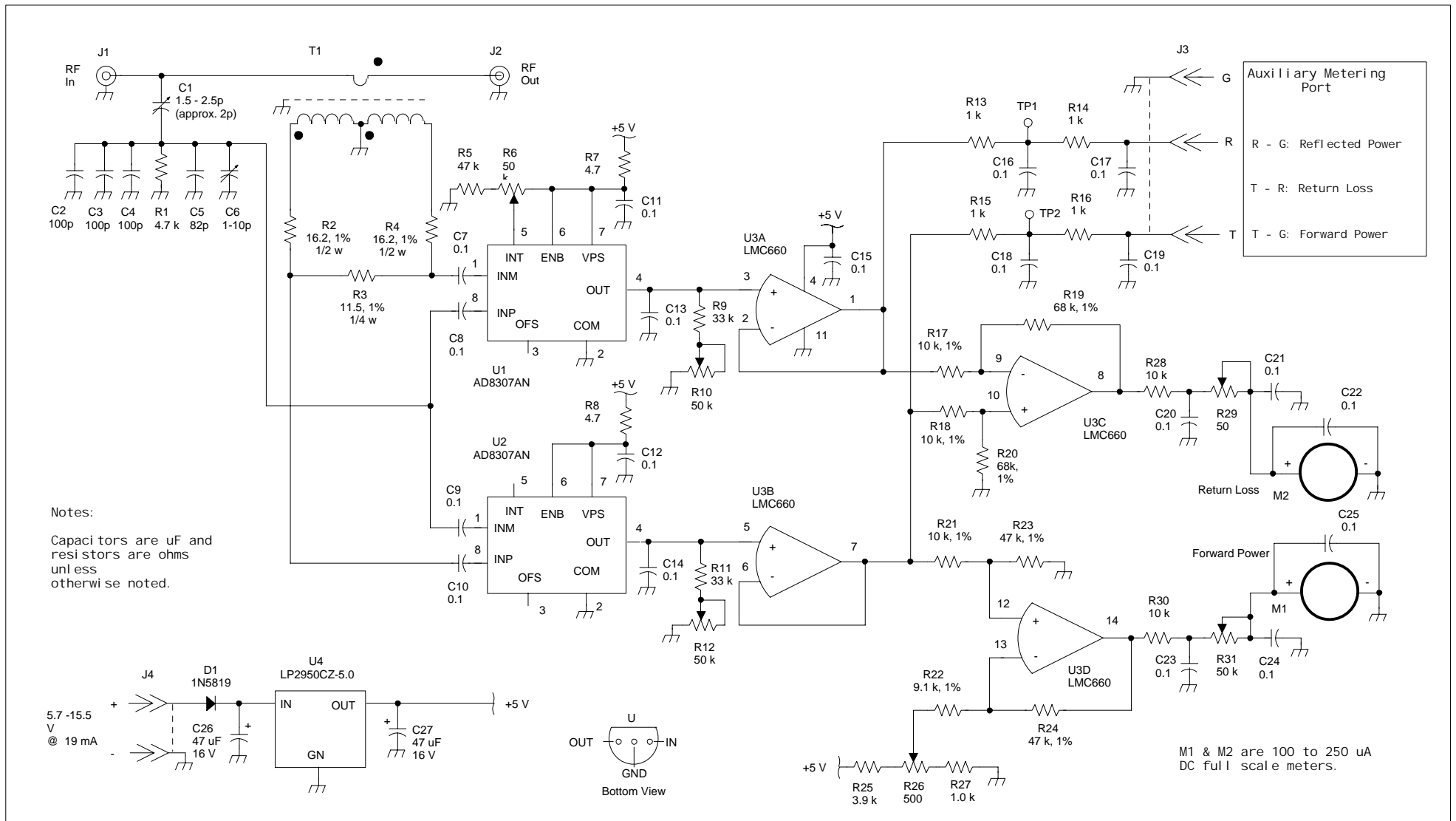


Figure 9 - Schematic of HF In-Line Return Loss and Power Meter

taken to keep unwanted signal sources away from these highly sensitive and wideband amplifiers.

The log forward and reflected voltages are subtracted in U3C which drives the reflected power meter M2. Since there is a large common mode voltage at the U3C input when the forward power is high, the R17 - R20 gain setting resistors are 1% tolerance to reduce the error caused by common mode response.

By providing the forward power signal to the positive input of U3C and the reflected to the negative input, the output voltage at U3C is 0 V for 0 dB return loss and +4 V for 30 dB return loss. Pot R29 provides the full scale adjustment for indicated return loss on M2.

Return loss meter protection is provided in two ways. Reverse meter swing is not possible since the U3C op amp output can not go below ground and the minus lead of the meter is connected to ground. Forward swing meter protection is provided by the U3C op amp maximum output limit of +5V. That level is equivalent to a 25% over-current through the meter and has not been a problem.

Disabling the return loss meter in the absence of RF is not required since both log amps produce approximately 0.17 V output. This is displayed as 0 dB, within about 0.5 dB, or essentially no meter deflection.

The log forward voltage and an offset voltage are subtracted in U3D. Pot R26 provides control of the offset voltage which sets the minimum power point on the forward meter scale. Forward power meter protection is provided by essentially the same means as used on the return loss meter. Pot R31 provides the full scale adjustment for indicated forward power on meter M1.

The voltage regulator used at U4 is a low drop-out type (about 250 mV) to maximize utilization of battery capacity. When combined with the reverse input protection provided by the D1 schottky diode, the minimum required DC input is +5.7 V and the current is 19 mA which is suitable for 9 V battery usage. At the high end, +15.5V should not be exceeded.

## Construction

The entire circuit, except for C17, C19, J3, J4, C22, C25, M1, and M2, is constructed on double-sided copper clad board. The board dimensions were selected to permit mounting inside an MFJ-948 antenna tuner. While this board was tailored to the MFJ-948 environment, it should be usable as is, or with minor modification, in many environments that have a pair of SO-239 connectors.

Figure 10 shows three pictorial views to assist in locating key components. The board is fastened to the shield side of the Transmitter SO-239 connector on the MFJ-948 using two shortened ring terminals soldered to one of the board's copper planes. The original pop rivets securing the SO-239 connectors were removed and replaced with 4-40 screws, nuts, and lockwashers.

The bridge components are split between both sides of the board for shielding. T1 and part of C1 are mounted on the side of the board that accesses the center conductor of the SO-239 connector. The metal plane on this side of the board is dedicated exclusively to ground. The primary winding of T1 uses the center conductor of a short section of RG-8X. The shield of this coax is grounded at one end only using a small piece of tin sheet stock that also supports the coax and T1. The coax jacket is left on where it passes through the core. The fit is quite tight and will require a careful flattening of the secondary turns against the core and possibly either sanding down the jacket or softening the jacket with a heat gun just prior to sliding the core over the coax jacket. The jacket could also be replaced with teflon or fiberglass tape.

The plates of C5 use the above solid ground plane on one side and a section of the metal plane on the other side. The remainder of the split plane, except for some small pot mounting pads, is dedicated to ground and is connected to the plane on the other side using six jumpers equally spaced along the split plane dividing line.

The relatively thin 0.031" board was used to maximize the capacitance available for C5 in a given board area. Standard 0.062" board can also be used with possibly some loss of high frequency directivity.

C1 is formed with a short section of RG-8X coax with the outer jacket removed. The center conductor is attached to T1 in a fashion that permits the coarse bridge balance by sliding the center conductor in and out of the coax. The coax shield is soldered to C5 where it passes through C5. The ground plane around C1 on the other side is cleared to avoid shorting. Three holes are drilled through C5 for C2, C3, and C4, which are chip capacitors that just fit between the planes of C5. The chip capacitors are carefully centered between the planes to avoid shorting C5.

Two holes are drilled through C5 for two of the three T1 secondary leads. The center-tap for T1 is soldered to the ground plane. R2 - R4 and the remaining board components are located on the opposite side of the board.

Just a few general comments about the style of dead bug (ugly) construction used here are in order.

First, the DIP module ground lead(s) are carefully bent and trimmed in length to be even with the top surface of the module and as close as possible to the body of the module without snapping off the leads. The remaining leads are cut at the point where the lead steps down in width. The modules are positioned with the top of the module placed against the board surface and soldered to the board ground plane.

Decoupling capacitors are used on all component power leads and soldered directly from the module or component lead to the ground plane as close as possible. All other decoupling capacitors or components, grounded at one lead, are mounted in the same way. The remaining components are soldered point to point and directly to module leads wherever possible.

Generally, only one layer of components above the modules is used, but two layers are used when components can cross without danger of shorting. Insulated wire is attached to a component which has been secured directly to a module at only one end. All insulated wiring is dressed against the ground plane.

Trimming the module leads and keeping the components and wiring around module lead height helps to minimize stray coupling, unit to unit variability, and inadvertent shorting. In some cases, isolated pads to secure larger or adjustable components are carved out of the board plane using an Exacto knife. Alternatively, hot melt glue can also be used to secure these components to the board without pads.

Board planes are never carved up near high frequency components or components that require shielding. Op amps, while not particularly high frequency components, are likely to rectify RF and generate offset errors. So, the board plane is left intact around the op amp module as well as the log amps.

Dead bug construction, as used here, allows the two board planes to shield the sensitive amplifiers from the high RF environment near the primary of T1 on the other side. However, there's no doubt that dead bug construction is ugly to many people, has virtually zero manufacturability, and needs greater care in handling.

The usual pin in hole and printed wire construction with components on the same side of the board as T1 may be used but may require additional measures for shielding.

Printed wiring should be restricted to the non-module side of the board to

## Parts List for Figure 9:

Unless otherwise specified, resistors are 1/8 W, 5% tolerance carbon composition or film units. All 0.1  $\mu\text{F}$  capacitors are 50 Volt axial ceramic (Digi-Key 1210PHCT-ND or Mouser 147-72-104). Equivalent parts can be substituted.

- C1 - 25 mm RG-8X
- C2, C3, C4 - 100 pF, NPO, 50 WVDC, Size 0603, ceramic chip capacitors (Digi-Key PCC101ACVCT-ND)
- C5 - 40 mm x 33 mm isolated area of copper plane on one side of 0.031 inch thick double sided copper clad board. The opposite side is ground.
- C6 - approx. 1 - 10 pF trimmer. (Mouser 24AA071 or Digi-Key SG-1034ND))
- C26, C27 - 47  $\mu\text{F}$ , 16 V, electrolytic (Digi-Key P969-ND or Mouser 140-HTRL16V47)
- D1 - 1N5819 Schottky diode (Mouser 583-1N5819 or Digi-Key 1N5819GICT-ND)
- J1, J2 - SO-239 chassis mount connector
- J3 - 1/8" (3.5 mm) panel mount, 3 cond., phone jack (RadioShack 274-249)
- J4 - 3/32" (2.5 mm) panel mount, 2 cond., phone jack (RadioShack 274-292 or 274-247)
- M1, M2 - 100 to 250  $\mu\text{A}$  full scale meter(s)
- R2, R4 - 1/2 W, 1% metal film (Mouser 273-16.2)
- R3 - 1/4 W, 1% metal film (Mouser 271-11.5)
- R6, R10, R11, R29, R31 - 50 kohm, cermet trimmer pot (Mouser 72-T70XW-50K)
- R17 to R24 - 1/8 W, 1% metal film (Mouser 278-Value, for example, 278-10K)
- R26 - 500 ohm cermet trimmer pot (Mouser 72-T70XW-500)
- T1 - Secondary is 11 bifilar wound turns of #22 AWG enameled wire, tightly and uniformly wound around a T50-3 core. Primary is 33 mm of RG-8X with shield grounded at J1 end only.
- U1, U2 - Analog Devices AD8307AN (Allied Electronics 630-8006, Newark 83F3404, or Future Active AD8307AN)
- U3 - National LMC660CN (Digi-Key LMC660CN-ND)
- U4 - National LP2950CZ-5.0 (Digi-Key LP2950CZ-5.0-ND)

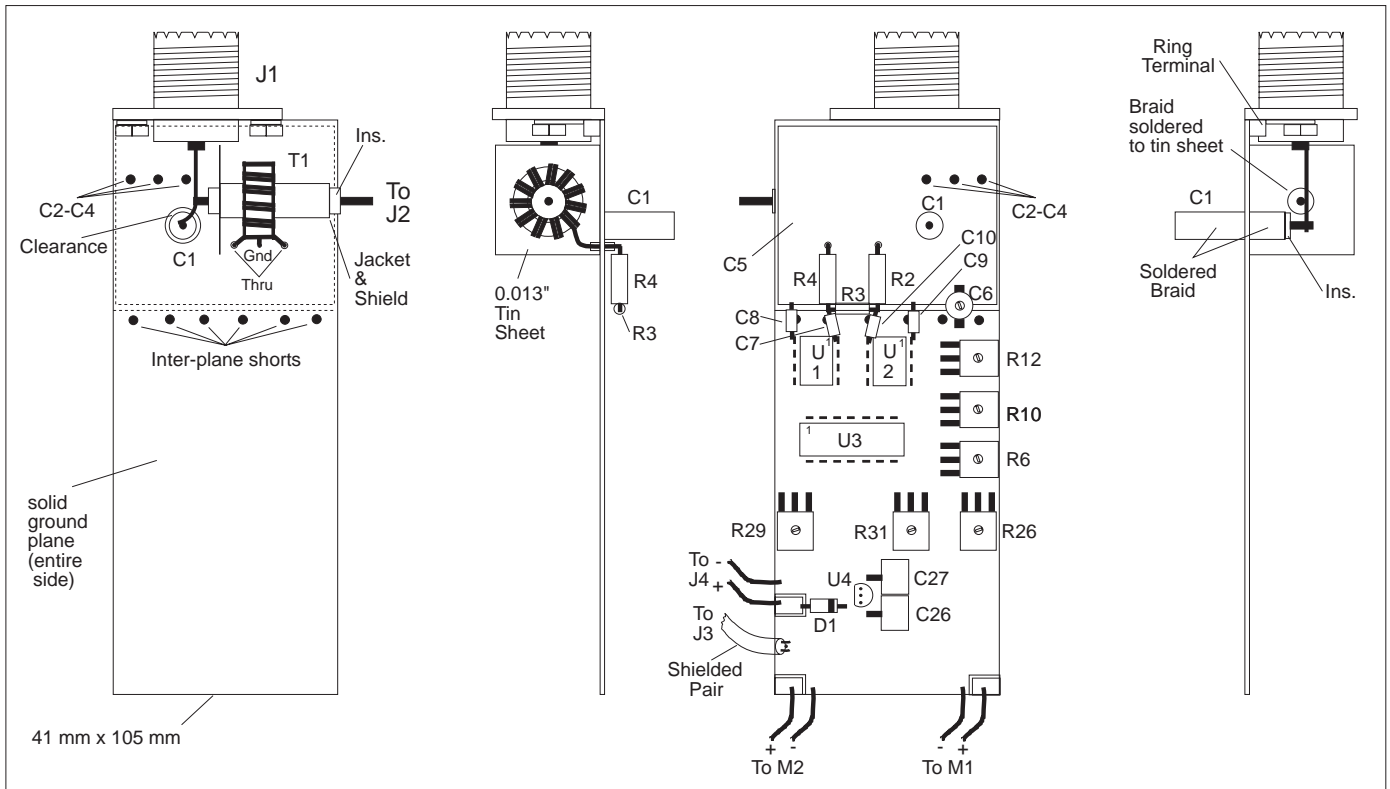


Figure 10 - Pictorial views of circuit board showing major components

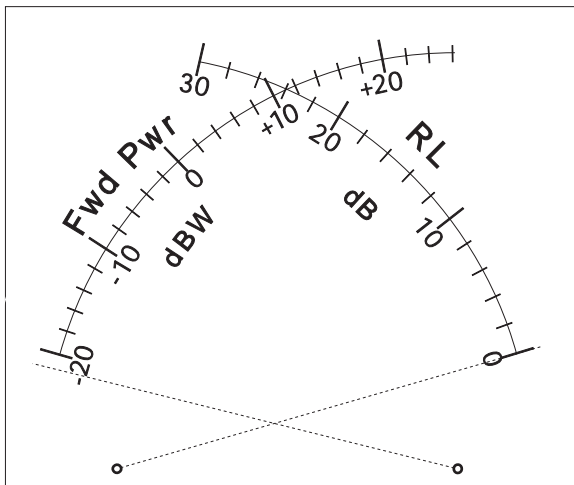


Figure 11 - Dual Movement Meter Scales, shows power off needle resting positions

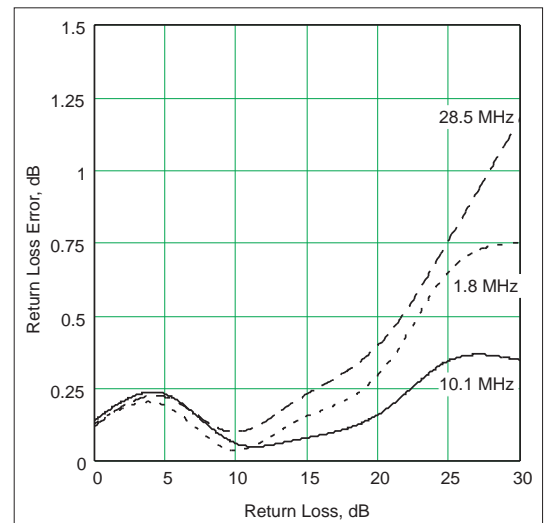


Figure 12 - Measured Return Loss Error

maintain the module side copper as a solid ground plane, except for the smallest possible clearances around component lead pads. Placing a grounded shield over the modules and mounting the circuit board on a metal box that surrounds the printed wiring may also be needed to provide RF shielding.

#### Meter Scales

The MFJ-948 antenna tuner uses a single meter enclosure with dual movements. A compatible overlay for the face of this meter is shown in Figure 11 and can be used as a basis for other meters.

The forward power scale ranges from -20 dBW to +26 dBW. Note that -20 dBW point is slightly above the forward power needle resting point. This scale, which is a fraction of the measurable power range, was chosen to improve accuracy over the anticipated range of transmitter output power levels. This range can be changed, but may require different values for R25 and R27.

The return loss scale ranges from 0 to 30 dB. The 0 dB point corresponds to its needle resting point.

#### Calibration

Referring to Figure 9 for connections, the calibration procedure is described as follows:

1. With power off, adjust the meter needle resting points as shown in Figure 11.
2. Connect a DC power source to J4. Connect a low power 7 MHz RF source (about 10 mW) to J1. Connect a high return loss 50- $\Omega$  termination<sup>9</sup> to J2. Set C6 for the center of its range. Monitor return loss at J3 with a voltmeter and adjust C1 for maximum return loss as close as possible. Then adjust C6 for maximum return loss. If desired, change to 28 Mhz and adjust the spacing between the wires in the bifilar pair on T1 to improve high frequency return loss.
3. Use the set-up from step #2 except monitor forward power at J3 using a DVM for best accuracy. Adjust R12 for 50 dBW/V using two known power levels. To determine the actual dB/V calibration factor, measure the forward voltage at each of the two power levels and take the difference; also calculate the difference in power levels in dBW. Then divide the power difference by the voltage difference and compare to 50 dB/V.
4. Use the set-up from step #3 but reverse the connections to J1 and J2. While monitoring reflected power at J3, adjust R10 for 50 dB/V using the two known powers and the calculation procedure from step #3. Precisely hitting

50.00 dB/V for steps #3 and #4 is not required as long as they match each other reasonably well.

5. Reconnect the signal source to J1 and leave J2 open. Make sure your RF source is capable of withstanding high SWR. While monitoring return loss at J3, adjust R6 for zero volts. For improved accuracy, alternate between a short and an open at J2 and balance the measured values around zero volts.

6. Cycle through steps #4 and #5 if needed.

7. Reconnect the 50- $\Omega$  termination to J2. Connect a known power RF source to J1 around -20 dBW (10 mW). Adjust R26 to set the meter to the appropriate mark on the scale.

8. Connect a known power source near +20 dBW (100 watts) to J1 and a suitably rated 50- $\Omega$  dummy load to J2. Adjust R31 to set the meter to the appropriate mark on the scale.

9. Cycle through steps #7 and #8 if needed.

10. Connect a low power RF source around 10 mW that can withstand high SWR to J1. Leave J2 open. Verify that the return loss meter reads zero within about 0.5 dB.

11. Use the same RF source at J1 from step #10 and connect a termination with known return loss around 20 dB<sup>10</sup> to J2. Adjust R29 to set the meter to the appropriate mark on the scale.

Once calibration is complete, four numerical values can be calculated for  $K_1$  and  $K_2$  in equations (1) and (2) if the auxiliary metering port is to be used. This is done using the measured data from step #3 and step #4 after step #6 is done.

The two values for  $K_1$  should be close to each other and can be averaged for most purposes. Due to log amp differences, the two values for  $K_2$  may be different and both should be used in that case. All four values can be used for best accuracy. For reference, with  $K_1$  equal to 50.3 dB/V,  $K_2$  is typically about -79 dBW.

To provide the two known forward powers required in the calibration procedure, I modified a Heath Cantenna to include a 40 dB attenuator with 50- $\Omega$  input and output impedances and a linearized diode detector, similar to that used in the Tandem Match. The detector monitors RF voltage on the low power attenuator output.

With this modified Cantenna, either the +20 dBW (100 watts) input power can be used directly or the attenuator output at -20 dBW (10 mW) can be used. The 100 watt input is verified using the diode detector.

The entire calibration procedure can be performed at these two power levels

using a 100 watt transmitter, a high power attenuator such as the modified Cantenna, 50 and 60- $\Omega$  low power loads, and a DVM.

In general, a 5 - 100 watt transmitter, a signal generator, a calibrated attenuator, another power or wattmeter, and a dummy load are possible tools that can be used to establish two known forward powers. The calibration procedure can be modified as needed to accommodate available equipment.

#### Accuracy

In addition to the measured characteristics shown in Figure 8, estimates of forward power and return loss accuracy were also made using the auxiliary metering port.

Forward power accuracy was evaluated using 10 - 300 watts to the modified Cantenna at 7 MHz. Over this power range, the power measured by this meter was within 0.4 dB of that measured by the linearized RF detector on the Cantenna.

Return loss accuracy was evaluated using a set of metal film resistor terminations, each coaxially mounted in a male BNC connector. The set provides nominal 0 to 30 dB return losses in 5 dB increments. Two resistor values, one greater than 50- $\Omega$  and the other less than 50- $\Omega$ , will have the same return loss; both values were used here. The actual return loss for each termination was measured using the 4 wire DC resistance technique. Figure 12 shows the maximum deviation in dB of the measured return loss vs. the actual return loss. Typical accuracy is better than 0.5 dB rising to 1.2 dB at 28.5 MHz and 30 dB return loss.

#### Operation

In normal operation, this meter would be inserted in a 50- $\Omega$  coaxial line between the transmitter and either an antenna or an antenna tuner. In either case, as little as 1 mW of RF (-30 dBW) can be used with this meter and still accurately measure up to 30 dB return loss. Few, if any, in-line wattmeters accurately measure 1  $\mu$ W of reflected power (-60 dBW). As a result, the level of on-the-air QRM can be reduced.

Adjustment of antenna tuners can be done at any power level exceeding 1 mW, but preferably as close to that level as the transmitter will allow. Simply adjust the controls to obtain either peak return loss or return loss greater than some desired value such as 10 dB (1.9:1 SWR).

After the adjustment is complete, normal intermittent service for forward

power up to 300 watts can be used if the return loss is 10 dB or higher. Limited testing at 500 watts with 20 dB return loss was performed without problems.

There should be little change in the displayed return loss, from 1 mW to 300 watts of forward power, as a result of instrumentation error. Log amp tracking, common mode rejection, and residual calibration errors will typically limit the variation to under 1 dB. Heating of antenna tuner components, particularly the inductor, or the resistor in a dummy load can cause perceptible changes for high return losses that may have previously gone un-noticed.

Lower than 10 dB return loss will require a reduction in maximum forward power. For example, 6 dB return loss (3.0:1 SWR) will require that the intermittent service forward power be limited to 230 watts (+23.6 dBW), which is equivalent to about 170 watts of load power.

Depending on the proximity of local transmitters, such as TV, radio, pagers, or other amateurs, there is another possible source of apparent return loss variation. Since the log amps are wideband and very sensitive, they will respond to strong local sources of RF at their inputs. At one time, a neighboring amateur would induce about a -45 dBW signal on my antenna, which is sufficient to affect measured return loss for forward powers less than -15 dBW (32 mW). The solution is to either raise power or tune up when the interference is gone.

The wide frequency and power range of this instrument make it suitable for other general test purposes. For example, gain/loss and return loss measurements of a wide variety of 50- $\Omega$  devices, such as filters, cables, amplifiers, and attenuators, can be performed. While the 1 mW minimum forward power required for 30 dB return loss may be too high for some amplifiers, it can be reduced by 25 dB if only gain measurements are performed. Return losses lower than 20 dB can be measured with 100  $\mu$ W.

50- $\Omega$  cable loss can be measured in two ways. The first is to use the fact that the cable loss in dB, at any frequency, is one-half of the return loss in dB measured when the far end of the cable is either open or shorted. Connect one end of the cable to J2 and an RF source to J1 and divide the measured return loss by two. The advantage of this method is access is required to only one end of the cable as long as the other end is open or shorted.

The other way of measuring cable loss, which is typically more accurate, is the insertion loss method. Connect an RF source to J1 and attach a 50- $\Omega$  load to

J2. Note the forward power in dBW on either the analog meter or as calculated using a DVM attached to the auxiliary metering port. Then insert the cable to be tested between the RF source and J1 and again note the forward power in dBW. The difference between the two measured powers is the cable loss in dB.

If a DVM is used for either return loss or differences in dBW forward power, the calculation can be simplified by first subtracting the two measured voltages and then multiplying by the nominal 50 dB/V log amp gain (or the actual value established during calibration) to obtain the cable loss in dB.

Measurements of most other devices can be done using one or both of the above techniques. The auxiliary metering port and a DVM will prove useful in extending measurement range and accuracy, particularly in cases, such as amplifiers, where input power must be limited to avoid gain compression.

## Modifications

The 500 MHz frequency range of the log amp makes VHF and UHF possible by simply changing the directional coupler. While a detailed design has not been done, some preliminary modelling indicates that a single coupled stripline design could be used for both 2 meters and 70 cm. The log amp frequency roll-off should roughly compensate for the increased stripline coupling factor as the frequency is raised. Additional frequency shaping is also possible.

The following changes can be made either singly or in combination:

- If two analog meters are not available, then one meter and a SPDT switch can be substituted.
- If the auxiliary metering port is not needed, then R14, R16, C17, C19, and J3 can be eliminated. Calibration is then performed using TP1 and TP2.
- If high accuracy is not needed, then R10, R12, R5, and R6 can be eliminated. Specified log amp tolerances may be adequate for many users. R9 and R11 would be changed to 51 k $\Omega$  and grounded at one end. Nearest 5% values for R17 - R24 can also be substituted.
- If forward power measurement is not needed, then U3D, R21 - R31, C23, C24, and M1 can be eliminated.
- Substituting the SOIC AD8310 for the AD8307 would eliminate the unity gain buffers since there are buffers in the AD8310. This would also allow use of the 8 pin LMC662 in place of the LMC660, saving more space. The log amp gain and offset

setting resistors would also be changed.

- C2 - C5 can be combined or changed to pin in hole leaded components with loss of directivity. For leaded components, use 20 nH/inch, applied to the component lead length plus the lead spacing at the component body, and Figure 7 as a guide to estimate directivity. Paralleling equal capacitors also roughly divides the lead inductance for one capacitor by the number of capacitors to obtain the inductance used in Figure 7.
- Other full scale meters, both below 100  $\mu$ A and above 250  $\mu$ A up to about 1 mA, can be used by changing R28 and R30.

Changing the maximum power either higher or lower is possible. Reducing the maximum power to 10 watts, for example, would simply require changing R25 to 5.1 k $\Omega$  and a meter scale change. The forward power scale would be changed to range from -30 dBW (1 mW) to +10 dBW (10 watts). Increasing the maximum power would require changes to the bridge, R25, and the forward meter scale. These changes are left to the interested reader<sup>11</sup>.

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## Notes

1. The data sheet for the AD8307, Analog Devices, Norwood, MA, is available from their website at <http://www.analog.com>. It also contains a good overview of log amp theory as well as general application information.
2. The technique used here to measure return loss is not new. See Virgil G. Leenerts, W0INK, "automatic VSWR and power meter," *ham radio*, May 1980, pp 34 - 43.
3. "S-units" are neither universally defined nor adhered to. 6 dB per S-unit is used here.
4. John Grebenkemper, KA3BLO, "The Tandem Match - An Accurate Directional Wattmeter," *QST*, Jan 1987, pp 18-26. The Tandem Match also appears in various editions of *The ARRL Handbook* and *The ARRL Antenna Handbook*.
5. See *The ARRL Antenna Book*, 18th Edition, Figures 4(E) and 4(F), p 27-4. The difference between Figure 4(F) and the Bruene bridge, as shown in Figure 4(E), is the addition of R1 which improves the low frequency directivity of the Bruene bridge.

6. For a detailed treatment of directivity and other related concepts, see *Network Analyzer Basics*, available from Agilent (HP) at

[http://dragon11.cos.agilent.com/data/downloads/eng/tmo/techinfo/pdf/comptest\\_nabasics.pdf](http://dragon11.cos.agilent.com/data/downloads/eng/tmo/techinfo/pdf/comptest_nabasics.pdf)

7. The degraded low frequency return loss of the Tandem Match coupler is caused by the inductance of the voltage sense transformer which shunts the mainline. Since the design already uses a high permeability core suitable for HF use, increasing the shunt inductance will probably involve added wire length which will degrade high frequency directivity. The shunt inductance also increases an error, called source match error, that can be as high as +/-1.1 dB for 1.8 MHz return losses near 0 dB.

8. Coupler directivity accuracy depends on the quality of the 50-Ω load used for the measurement. Input return loss accuracy depends on the 50-Ω load as well plus the directivity of the test bridge used. While the accuracy is not traceable to NIST (preferable but expensive), the combination of the 50-Ω load described below and a carefully constructed type of Wheatstone bridge exceeds 50 dB measured directivity for 1.8 - 54 MHz. The DC resistance of the 50-Ω load was measured using the 4 wire technique (See p 26.7 of the 1998 ARRL Handbook.). This Wheatstone bridge is constructed with 49.9-Ω resistors for the three internal bridge arms, a single differential input AD8307 log amp for the detector, and a 52.3-Ω across the log amp input. All resistors are 1%, 1/4 watt, metal film resistors. The differential input of the log amp eliminates the transformer, usually present to accommodate unbalanced detectors. Eliminating the transformer greatly improves directivity. To achieve a good open / short ratio with this bridge, a good 50-Ω generator impedance must also be maintained.

9. A low power, high return loss, 50-Ω termination can be built using a 49.9-Ω, 1%, 1/2 watt, metal film resistor which is coaxially mounted in a modified PL-259 connector. The resistor leads are cut as short as possible. The main body of the connector is cut off or ground down to just back of the threads. A small disk made from sheet tin stock, with a hole in the center, is soldered along its perimeter to the back of the connector and to the protruding resistor lead. A higher test power will require a suitably higher power rated 50-Ω termination or dummy load.

10. A 60.4-Ω, 1%, 1/2 watt, metal film resistor mounted in a PL-259 connector in the same fashion as above will provide 20.5 dB return loss, accurate to +/-0.5 dB. Various other precision resistors can be mounted in similar fashion and used to verify accuracy at other points on the return loss scale if desired. Four wire DC resistance measurements can be used to reduce the error due to resistor tolerance and soldering.

11. Referring to Figure 5B, the phase compensated Bruene bridge design equations are provided as a guide:

Bridge balance:

$$\frac{C1}{C1 + C2} = \frac{R3'}{2 \times Z_0 \times N}$$

where R3' is the parallel combination of R3 and 2.2 kΩ (2 times the log amp differential input impedance), Z<sub>0</sub> is the line characteristic impedance, normally 50-Ω, and N is the total secondary to primary turns ratio, or 22 in the current design.

The coupling factor in dB is:

$$CF = 20 \times \log \left( \frac{Z_0 \times N}{R3'} \right)$$

The phase compensation resistor, R1, is calculated from R1', which is the parallel combination of R1 and 510-Ω (the combined log amp common mode input impedance), using:

$$R1' = \frac{Ls}{(R2 + R3' + R4) \times (C1 + C2)}$$

where Ls is the secondary inductance.

Power dissipation in R2, R3, and R4 is calculated using the resistor values and the maximum transformer secondary rms current, I<sub>s,max</sub>:

$$I_{s,max} = \frac{1}{N} \times \sqrt{\left( \frac{P_{ld,max} \times SWR_{max}}{Z_0} \right)}$$

where SWR<sub>max</sub> is the maximum SWR that can be present simultaneously with the maximum load power in watts, P<sub>ld,max</sub>.

#### Sidebar

#### Power, SWR, and Return Loss

The units *dBm* and *dBW* are measures of RF power in dB with respect to references of one milliwatt and one watt respectively. Conversions from watts to dBW and dBm are:

$$P_{dBW} = 10 \times \log(P_w)$$

and

$$P_{dBm} = P_{dBW} + 30$$

All logs used here are base 10.

Return loss in dB, RL, is related to the magnitude of the complex voltage reflection coefficient, |ρ|, by:

$$RL = -20 \times \log|\rho|$$

Voltage standing wave ratio, VSWR, is related to |ρ| by:

$$VSWR = \frac{1 + |\rho|}{1 - |\rho|}$$

and can be calculated from return loss using:

$$|\rho| = 10^{-RL/20}$$

To obtain return loss in terms of forward and reflected power, we start with:

$$|\rho| = \frac{|V_r|}{|V_f|}$$

and some logarithm relationships to obtain:

$$RL = 20 \times \log|V_f| - 20 \times \log|V_r|$$

And, since

$$P_w = \frac{V^2}{Z_0}$$

for each of the forward and the reflected voltages, some more algebra yields:

$$RL = P_{f,dBW} - P_{r,dBW}$$

Load power can be calculated using the measured forward power and return loss. Convert forward power in dBW to watts using:

$$P_{f,w} = 10^{P_{f,dBW} / 10}$$

Reflected power in dBW is calculated from:

$$P_{r,dBW} = P_{f,dBW} - RL$$

Reflected power is converted to watts using:

$$P_{r,w} = 10^{P_{r,dBW} / 10}$$

Load power is determined from forward and reflected power, all in watts, by:

$$P_{ld,w} = P_{f,w} - P_{r,w}$$

(End of sidebar)