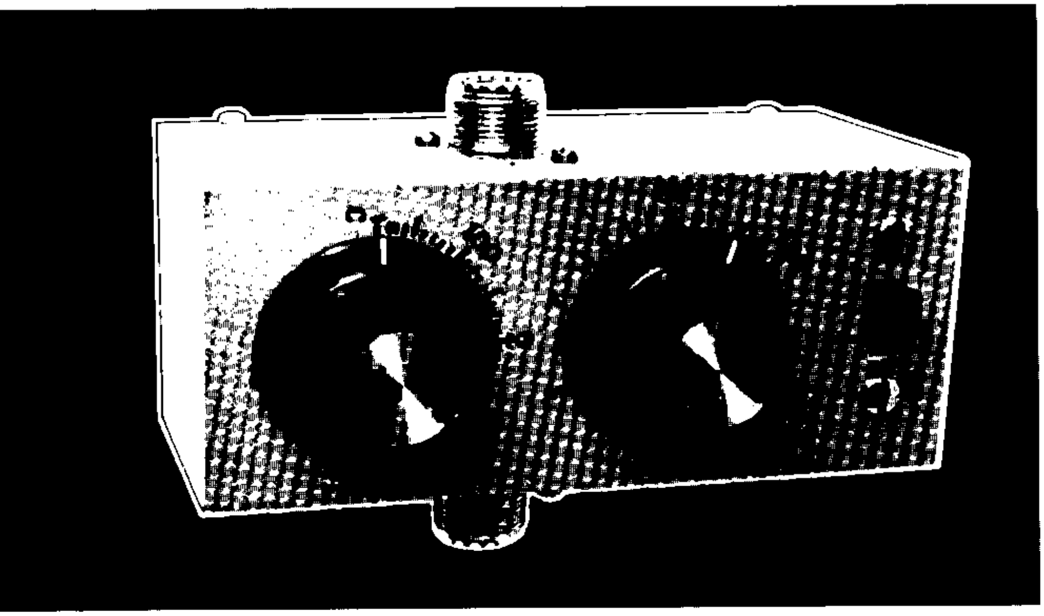


improvements to the RX noise bridge

Circuit and construction improvements for this simple device improve accuracy and measurement range for impedance measurements from 3.5 to 30 MHz

How often have you wanted to know the exact impedance — both the resistive and reactive components — of your high-frequency antenna? Although the venerable swr meter recognizes the situation which makes the transmitter happy, it very often is misleading with regard to actual antenna resonance and related matching adjustments. Amateur literature has approached the subject of measuring complex impedance in the high-frequency range a number of times. In spite of all this, the use of these techniques by amateurs is still relatively uncommon. This is indeed unfortunate, since many antenna adjustments become systematic if resistance and reactance are accurately known.

The rf noise bridge is one of several devices that can be used to make this type of measurement. YA1GJM's excellent article originally got us interested in this device.[1] Subsequently, we have pursued modifications to the instrument which both improve its accuracy and extend its range. Ultimately, an rf noise bridge unit resulted which has the following significant characteristics:

1. Measurement accuracy of 3 ohms rms* for both the resistive and reactive components of representative rf loads.

A parts kit for the noise bridge is being made available in conjunction with this article. For information and prices write to G. R. Whitehouse & Co., 10 Newburg Drive, Amherst, New Hampshire 03031.

By Robert A. Hubbs, W6BXI, and A. Frank Doting, W6NKU. Mr. Hubbs' address is 2927 Roberta Drive, Orange, California 92669; Mr. Doting can be reached at 13031 Ranchwood Road, Santa Ana, California 92705

2. Capability, from 3.5 to 30 MHz, to measure complex impedances equivalent to any point within a 5:1 swr circle on a Smith chart with a $Z_o = 50$ ohms.

3. A calibration concept which does not require laboratory standards and is valid over the entire frequency range up to 30 MHz.

4. Construction cost under $20.00.

With reasonable care, anyone can achieve these same results. All you have to do is follow the instructions on construction and calibration which are included in this article. In addition to these tips we will cover a wide range of other topics relative to the noise bridge. The first two sections describe the basic principles of the noise bridge and some history regarding its development. The next three sections describe our principal contributions — extending the range and improving the accuracy of the instrument. Details on how to build, checkout, calibrate, and use the improved instrument are concentrated in the last three sections.

noise bridge description

An rf noise bridge may be used to measure complex impedances — normally a very difficult thing to do without laboratory instruments. It works as shown in the block diagram of **fig. 1**. An unknown impedance to be measured is connected to the input terminal of the bridge; a receiver is connected to the output terminal where it is used as a sensitive, frequency-selective detector. Contained within the bridge is a wideband noise source which provides a signal over the frequency range of interest. Reference devices with known rf impedances are used to balance the bridge section and to measure the unknown impedance.

A schematic diagram of the bridge portion is shown in **fig. 2**. It works in the following manner: Wideband noise is injected into two legs of the bridge in equal quantities via a core transformer, T1. With the unknown impedance connected and the detector (your receiver) set to the desired frequency, R_p and C_p are adjusted for the deepest obtainable null. There is some interaction between the two adjustments, depending upon frequency, so that a readjustment between the two controls may be required to obtain the deepest null. When the null condition is achieved, the value of the unknown impedance is equal to the parallel combination of R_p and C_p. To permit measurement of both positive and negative values of parallel capacitance, the zero value of the C_p dial is set with the variable capacitor, C_p, approximately half-meshed. To balance the bridge with this offset, a fixed capacitor, C_f, is placed across the unknown terminal. This forces the bridge to balance with a purely resistive load when the half-meshed C_p is equal to C_f. However, by labeling the C_p dial zero at this point, the correct capacitance is registered by the instrument.

Several advantages of the noise bridge concept now become apparent:

1. The frequency at which the measurement is being made is determined by the detector (your communications receiver), and this should be very accurate.

2. Measurement of inductive reactance does not require an accurate variable inductance which, in practice, is difficult to build.

3. Very little power is required from the noise source generator since the detector is quite sensitive.

On the negative side, you might expect some difficulty with this circuit for the following reasons:

1. The capacitance dial does not read out directly in ohms since this parameter is a function of the measurement frequency.

2. The resulting R_p and C_p values are parallel-circuit values, and series parameters are required in many practical applications.

3. A large amount of parallel capacitance is required to achieve balance with some modestly reactive impedances, particularly at low frequencies.

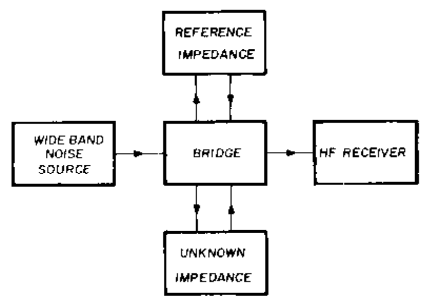


fig. 1. The noise bridge includes five major elements, three of which are internal to the instrument. The heart of the instrument is the bridge section itself. This is excited by a broad spectrum noise source. The unknown impedance and a reference impedance form separate legs of the bridge section. The reference impedance is varied until it equals the unknown impedance. When this occurs, the bridge is nulled and the output of the high-frequency receiver goes to a minimum.

Don't become overly discouraged over the negative aspects — we would like to add that none of these are of serious concern and, in fact, the first two reduce to simple calculations with a hand-held calculator. The third may be handled with the aid of a range extension assembly which we discuss in a later section. The required mathematical relationships will be covered in a section devoted to practical applications.

initial units

Our exposure and efforts on the practical noise bridge circuit had an innocent beginning when we built the noise bridge described by YA1GJM with the modification suggested by K2BT.[1,2] Initial experiences were

*Rms stands for root-mean-square which is a measure of dispersion. Mathematically it is equal to the square-root of the average or mean value of the squares for a series or data points. For random errors, usually about two-thirds of the errors will be smaller than the rms error.

delightful. Extremely deep and repeatable nulls could be obtained while measuring complex impedances. Actual use of the device on antenna measurements and adjustment of a transmatch were not, however, nearly as fruitful. Adjustment of a transmatch to obtain a measured 50-ohm resistive impedance with a 21-MHz antenna system did not coincide with the adjustments to

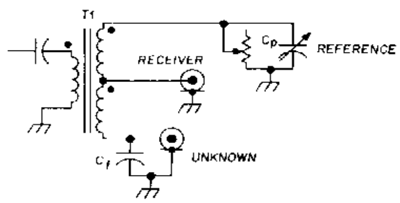


fig. 2. The bridge section of previously recommended noise bridge designs included a trifilar wound transformer. This injects wide spectrum noise approximately equally into the two halves of the secondary. When the parallel combination of the unknown impedance, shunted by a capacitor, C_f, is equal to the parallel combination of R_p and C_p, the bridge is balanced and the receiver detects a null condition.

achieve minimum swr meter readings. Similarly, we found that the familiar Heath dummy load did not appear to be close to the 50-ohm design value at the higher frequencies. Somewhat perplexed at this point, we fell back to consider the next step.

K2BT mentioned evidence of calibration shifts when using the device at higher frequencies and this seemed to be related to our experience. We also began to suspect that our knowledge of the impedance of carbon resistors which we had been using to evaluate performance of the bridge was not satisfactory.

We decided to explore the intrinsic performance of the noise bridge in a more rigorous manner. Fortunately, we have access to a Boonton Radio RX-Meter, model 250A. This is a highly respected, laboratory quality instrument which also measures the R-C parallel equivalent representation of an rf impedance. Test circuits were constructed of various resistances, capacitances, and/or inductance combined in parallel or series. These were measured first on the Boonton and next on a second noise bridge unit which we built with more care to shorten leads in the bridge circuit. The results are summarized in **table 1**.

These data were mostly concentrated at the higher frequencies (21 and 28 MHz) where problems, if encountered, were expected to be most pronounced. **Table 1** indicates some interesting consistencies as well as some interesting anomalies. The noise bridge nearly always indicated a more inductive (smaller) C_p than the RX Meter. The only exception was test circuit 1. Here the error was not only capacitive but also extremely large. There was also a tendency for the noise bridge to indicate a higher than correct value for R_p, particularly for higher values of load resistance.

It is interesting to compare these results with the words of K2BT.[2] He thought his unit exhibited an "inductive rotation" of the C_p calibration at the higher frequencies. That is a succinct way of summarizing our C_p data in **table 1**. Conversely, however, K2BT did not notice R_p errors at the higher frequencies. YA1GJM claimed that 2-watt composition resistors in the 10- to 150-ohm range exhibit "about 14 pF inductance in parallel with their indicated resistance values."[1] Test circuits 1 and 4 are composition resistors, number 1 being a 15-ohm value, and number 4 a 150-ohm value.

We find it very interesting that our noise bridge also indicated roughly this value of shunt inductance for these resistors. Note, however, that the readings are *incorrect*! In general, our findings indicate that composition resistors above 50 ohms exhibit relatively small values of shunt reactance. In fact, these are handy devices for use in calibrating the noise bridge as we shall explain later.

We were basically dissatisfied with the performance of this latest noise bridge. We think it is fair to say that our unit was probably as good as those made by either YA1GJM or K2BT. Indeed, as pointed out above, there appear to be significant correlations in the way our unit performed with the descriptions provided by these authors. Therefore, it seemed probable to us that some systematic errors were inherent to all these units. This prompted us to vigorously attack the problem of accuracy improvement.

improving accuracy

We could easily fill this entire issue of *ham radio* if we tried to relate, in any detail, all the avenues pursued in trying to improve noise bridge accuracy. A total of six different units were built with variations in each which we thought would be helpful. Many of our initial recipes for improving accuracy turned out to be blind alleys; others showed promise but did not provide the desired improvement. In the end we found four distinct design changes which definitely improve the performance of the instrument.

All of our improvements relate to the fundamental problem of achieving balance in the bridge section. Consider **fig. 2** which shows the previously recommended trifilar-wound transformer, T1. Clearly the desire is to balance both sides of the secondary circuit. Indeed, if the parallel combination of the unknown impedance and the fixed capacitor, C_f, is equal to the parallel combination of R_p and C_p, the secondary should be balanced. However, this will be true only so long as there are no additional coupling paths which affect secondary circuit balance. Unfortunately, the toroidal transformer couples energy between the primary and the secondary circuits *both* inductively and capacitively. Thus, if the primary circuit is unbalanced as it is in the trifilar configuration, then you can expect this capacitive coupling to slightly unbalance the secondary circuit as well.

To understand this point, imagine small fixed capacitors from the primary circuit to the secondary circuit.

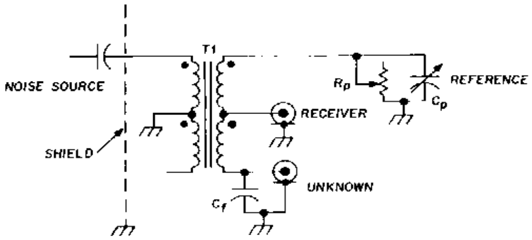


fig. 3. The recommended improved bridge section contains a quadrifilar wound transformer. This winding technique significantly reduces unbalanced capacitive coupling between the primary and secondary circuits. An electrostatic shield between the bridge and the amplifier electronics is also recommended.

Since the *dot* ends of each winding are physically near each other, then you expect the *reference* side of the secondary circuit to be most tightly coupled to the *ungrounded* side of the primary, whereas the *unknown* side would be more tightly coupled to the *grounded* part of the primary, circuit. It is not difficult to imagine rather significant unbalancing effects from this unbalanced capacitive coupling – particularly so at the higher frequencies.

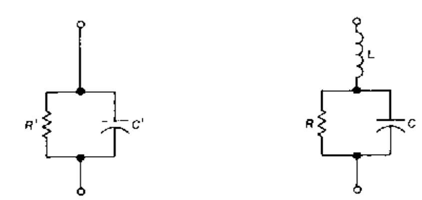
Fig. 3 schematically shows the recommended transformer configuration. The toroid is wound in a quadrifilar configuration. The additional winding serves to balance the primary to secondary circuit capacitive coupling. Note that one end of this primary circuit is left floating. To achieve the most perfect balance, it should be tied back to ground through an impedance equal to the driving impedance of the amplifier stage. However, from a practical standpoint, our results indicate no measurable improvement in trying to simulate this amplifier impedance with various terminations. Hence, we recommend leaving the other end floating as shown in **fig. 3**.

Note also that a grounded electrostatic shield is installed between the amplifier and the toroid. This serves the same purpose as above, reducing stray capacitive coupling to the secondary circuit. These are two of the four improvements we made, and in retrospect it is clear that the quadrifilar winding is the more important improvement.

Ground loops can easily exist in rf circuits. Unfortunately, they are difficult to diagnose and their effects on circuit performance are sometimes difficult to predict. We had observed strange behavior in several of our early noise bridge models which we now ascribe to this phenomenon. The mechanism is relatively simple. The chassis is necessarily part of the secondary circuit. It is also possible to have currents from the primary circuit return through the chassis; however, when this occurs, the primary and secondary circuits are again coupled in a way which may lead to unbalanced behavior. Rather than pre-empt the section which carefully details how to build our improved noise bridge, we will conclude discussion of this third recommended improvement by simply saying it is necessary to carefully ground the amplifier and bridge circuit components to achieve the best accuracy.

After making all three of the above noise bridge improvements, we were still noting some residual high-frequency unbalance. Our results, measuring carbon resistors in the 150-200 ohm range with the Boonton 250A, indicate that these resistors have very small equivalent parallel capacitance – usually 1 or 2 negative picofarads. Hence, as we shall see in the section on checkout and calibration, they are convenient standards for estimating noise bridge accuracy. At 3.5 MHz, measuring a 150-ohm resistor, the modified noise bridge would correctly indicate very small parallel capacitance, but at 28 MHz it could indicate nearly −10 pF and, furthermore, the indicated resistance value was higher than it should be. We soon discovered that by reversing the secondary winding of the quadrifilar transformer, the effect could be made to reverse. Hence, the error at 28 MHz would now be in the capacitive direction and the indicated resistance would be too low. This clue was sufficient to point us in the right direction.

Consider **fig. 4**. Suppose we have a parallel R-C circuit to which we add a very small series inductance. We wish to find the parallel $R'C'$ circuit which, at some particular frequency, has the same impedance as the circuit with the added series inductance. The algebra you have to grind through to get expressions for R' and C' in terms of R, L, C, and frequency is a bit tedious. However, if you are willing to make the approximation that the



$$Z_o = \frac{R'}{1 + j2\pi f R'C'} \qquad Z_i = j2\pi f L + \frac{R}{1 + j2\pi f RC}$$

$$Z_o \cong Z_i \text{ IF } R' = \frac{R}{1 + 8\pi^2 f^2 LC} \text{ AND } C' = C(1 + 4\pi^2 f^2 LC) - L/R^2$$

fig. 4. A small amount of series inductance added to a circuit results in a shift in the R-C parallel representation of the impedance of that circuit. The shift in impedance can be represented by another parallel R' - C' circuit. Numerical values for the shifted resistance and capacitance are found from the equations in the text. These are approximate in that they assume the reactance of the series inductor to be smaller in magnitude than the reactance of either R or C.

reactance of the small inductor is much less than that of any of the other components, then a relatively simple result emerges:

$$R' \approx \frac{R}{1+8\pi^2 f^2 LC}$$

$$C' \approx C(1 + 4\pi^2 f^2 LC) - L/R^2$$

where f is frequency.

Now to the point of this whole discussion. Suppose you wished to measure the impedance of a 100-ohm resistor with your noise bridge. On the unknown side you would have the 100-ohm resistor in parallel with the fixed capacitor. For the moment, assume the fixed capacitor to be 180 pF. Ideally, the bridge section would balance if you turned the pot and the variable capacitor on the reference side to these same values. Suppose, however, that one of these sides, either the reference or the unknown, had a small 10 nanohenry inductance in series with it. At 3.5 MHz, using the equations shown above, $R' = 99.8$ ohms and $C' = 179.2$ pF.

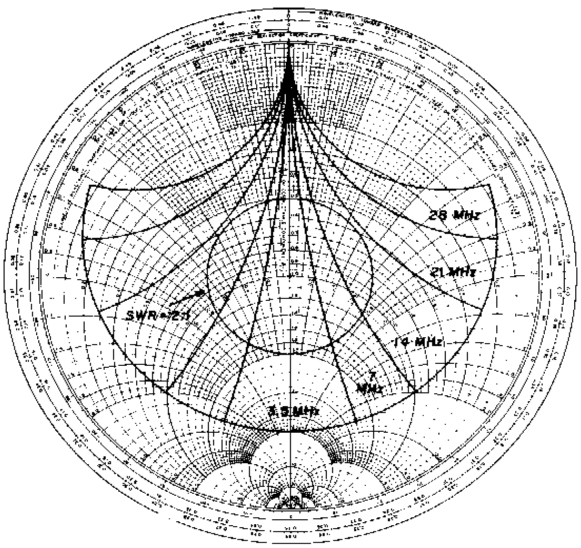


fig. 5. The range of impedance which can be measured using the basic noise bridge configured with a 365 pF variable capacitor and a 250-ohm potentiometer is shown in this diagram. Since capacitor reactance varies with frequency, the range of the bridge increases with increasing frequency as shown. The center of the Smith chart is 50 ohms and other impedances are scaled relative to this value. A 2:1 swr circle is shown on the chart for comparison.

Hence, at this frequency, the very small series inductance causes a small but essentially immeasurable shift in bridge balance. However, at 28 MHz, using the same values as before, $R' = 90.0$ ohms and $C' = 189.0$ pF. Thus, the very small series inductance becomes a significant bridge unbalance at this frequency. If it exists on the known side, it causes the bridge to overestimate the resistance and to miss the correct capacitance on the inductive side. If it exists on the unknown side, the errors are reversed.

To put this in perspective, no. 28 (0.3mm) wire has an inductance of about 9 nanohenries/centimeter. In the process of winding a toroidal transformer and in wiring the bridge section it is not hard to imagine getting an extra centimeter of copper in one side of the bridge. Hence, our fourth and last suggested improvement is to achieve final bridge balance using a short piece of bare hook-up wire on the unknown side of the bridge to balance this effect. We will discuss this procedure in more detail in the section on checkout and calibration.

range extension

In terms of parallel equivalent circuits, the range of the noise bridge using a 250-ohm pot for R_p, a 365-pF variable for C_p, and 180 pF for C_f is roughly:

$$0 \leqslant R_p < 250 \text{ ohms}$$

$$-180\,pF \leqslant C_p < 180\,pF$$

This representation, however, is not very informative. Instead, let's look at this range as a contour plot on a Smith chart. W1DTY discusses the use of these charts in the November, 1970, issue of *ham radio*, and this discussion assumes that you are familiar with using them.[3]

Since the reactance of a fixed capacitor is frequency dependent, we can expect the range of the bridge to also be a function of frequency. This is shown in **fig. 5**. In this presentation the center of the chart is 50 ohms. Five contours are shown; each represents the range of the noise bridge at the indicated frequency.

Suppose you wished to measure the impedance of an antenna at 3.5 MHz. Suppose further that the antenna swr at this frequency is 2:1. Therefore, its impedance is somewhere on the 2:1 swr circle on the Smith chart. Note, however, from **fig. 5** that only a very small portion of the possible impedances on a 2:1 swr circle are within the bridge's range at 3.5 MHz. K2BT noted this same problem, and recommended using a 365 pF variable instead of the 140 pF variable originally suggested by YA1GJM to increase the bridge's range. K2BT also suggested building the bridge so that fixed capacitance

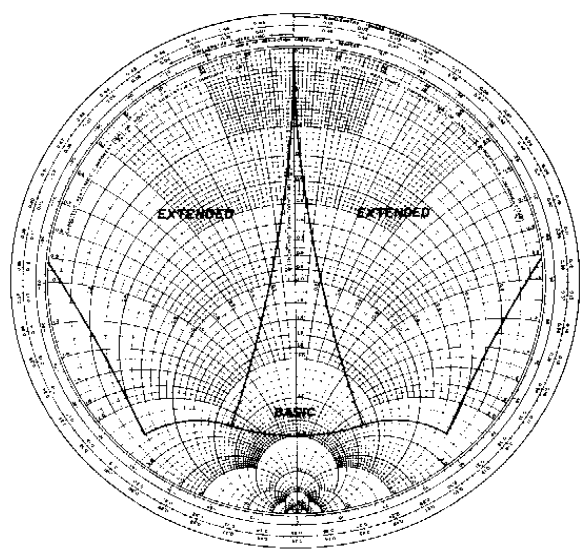


fig. 6. The range of the basic noise bridge can be greatly extended by adding a 100-ohm resistor in series with the unknown impedance. The extended range of impedances which can be measured using this technique is shown on this Smith chart. The center of the chart is 50 ohms, and the extended range includes all points on and within a 5:1 swr circle at 3.5 MHz.

could be added in parallel to either the reference or unknown sides of the bridge. We feel this is an excellent technique. Its only limitation is the required size of the fixed capacitance and possible construction difficulties. For instance, to measure any point on a 2:1 swr circle at 3.5 MHz requires the addition of up to 510 pF to the bridge. Extending the requirement to a 5:1 swr circle requires up to 2010 pF fixed capacitance.

An alternative which we feel is often more attractive is the addition of 100-ohm resistor in series with the unknown. In effect, this lowers the Q of the circuit to be measured and brings the resultant within the range of the bridge. The possible range extension at 3.5 MHz is shown graphically in **fig. 6**. This extended range includes all points on and within a 5:1 swr circle at 3.5 MHz and becomes even larger with increasing frequency.

A convenient range extender assembly will be described in the construction section; the mathematical formulas required to process the noise bridge readings with the 100-ohm range extender in place are included in the applications section.

results

So far we have stated that accuracy improvements to the noise bridge can be made with certain design changes. We have also alluded to a technique for extending its range. The last three sections of this article will describe in detail just how to accomplish these results. However, before we proceed with these detailed descriptions, it's logical (and motivational) to demonstrate our claims.

Table 1 summarized data taken on a noise bridge in accordance with earlier prescriptions. By way of comparison, **table 2** summarizes measurements made on these same terminations, but taken with two units built as described in this article. These are labeled, respectively, Unit 2 and Unit 3. Each of us built one of these, and other than the specific requirements imposed in the construction section, they are dissimilar. For purposes of comparison, the data from **table 1** (and Unit 1) are repeated.

Units 2 and 3 show a marked improvement in their correspondence to the Boonton measurements. The "inductive rotation" error evident in Unit 1 all but disappears in both these units. Further, the rather sizable error in measuring the C_p of the first load has disappeared. The only significant errors which remained in the Unit 2 and 3 data are on the last test circuit where errors of 12 and 17 pF, respectively, were made. But these are one-half the corresponding error in Unit 1. Also, any substantial indication of a systematic error in resistance measurements has disappeared. The rms error pretty well summarizes the data. Units 2 and 3 outperformed Unit 1 by a factor of more than three to one!

Although the results shown in **table 2** pretty clearly indicate an improvement in measurement accuracy for both Units 2 and 3, we felt that additional data was desirable to more fully explore the accuracy potential of the improved units. Using a number of different building blocks, including discrete resistors, capacitors, and inductors and, in several cases, short coaxial lines with resistive termination, we built a number of additional test loads. Our intention was that these loads should include both inductive and capacitive reactance and both small and large values of resistance. They were synthesized to be roughly arranged around the 3:1-swr circle on a Smith chart with $Z_o = 50$ ohms. Some of

table 1. A noise bridge unit built according to previously described articles as compared to a Boonton 250A RX meter.

MEASURED IMPEDANCE PARAMETERS

test circuit number	freq. MHz	Boonton R_p	Boonton C_p	noise bridge R_p	noise bridge C_p	error relative to Boonton ΔR_p	error relative to Boonton ΔC_p
1	7.2	15.5	−156	17	−18	1.5	138
2	28.2	138	−62	160	−65	22	3
3	21.2	162	−43	162	−52	0	−9
4	28.2	149	−1	160	−15	11	14
5	14.2	141	59	145	45	4	−14
6	21.2	37.5	182	35	150	−2.5	−32

table 2. Improved noise bridge units 2 and 3 as compared with unmodified unit 1 and the Boonton RX meter.

MEASURED IMPEDANCE PARAMETERS

test circuit number	freq. MHz	Boonton R_p	Boonton C_p	Unit 1 R_p	Unit 1 C_p	Unit 2 R_p	Unit 2 C_p	Unit 3 R_p	Unit 3 C_p
1	7.2	15.5	−156	17	−18	14	−160	16	−160
2	28.2	138	−62	160	−65	142	−55	140	56
3	21.2	162	−43	162	52	161	−41	160	−40
4	28.2	149	−1	160	−15	150	−4	147	1
5	14.2	141	59	145	45	142	60	136	58
6	21.2	37.5	182	35	150	35	165	34	170
rms measurement error relative to the Boonton				10.2	58.5	2.1	7.8	2.9	5.9

table 3. Improved noise bridge units 2 and 3 as compared with the Boonton RX Meter for a number of representative test circuits. All recorded data was converted to series impedance and therefore the entries have units of ohms.

test circuit number	freq. MHz	100-ohm resistor used	measured impedance Boonton	measured impedance Unit 2	measured impedance Unit 3
1	7.2	no	15 + j2	14 + j1	16 + j2
2	14.2	yes	19 + j48	22 + j49	19 + j49
3	28.2	no	42 + j63	49 + j67	48 + j66
4	21.2	no	86 + j81	91 + j80	93 + j79
5	28.2	no	149 + j6	148 + j16	147 + j4
6	14.2	no	91 − j67	90 − j67	91 − j63
7	3.7	yes	51 − j45	52 − j44	51 − j44
8	21.2	no	21 − j19	22 − j17	21 − j17
9	3.5	yes	17 + j16	20 + j17	18 + j15
10	7.0	yes	23 + j34	28 + j35	27 + j32
11	14.0	no	120 + j61	122 + j58	123 + j54
12	21.0	no	55 − j59	54 − j61	54 − j58
13	28.0	no	21 j16	21 − j17	19 − j13
14	3.5	yes	85 − j67	81 j66	80 − j66
15	7.0	yes	39 j51	38 − j53	38 j49
16	14.0	yes	19 − j14	17 j13	19 − j14
17	21.0	yes	20 + j14	22 + j12	20 + j16
18	28.0	no	43 + j47	48 + j49	46 + j52
rms measurement error relative to the Boonton				ΔR = 3.2 ohms, ΔX = j2.9 ohms	ΔR = 3.0 ohms, ΔX = j2.8 ohms

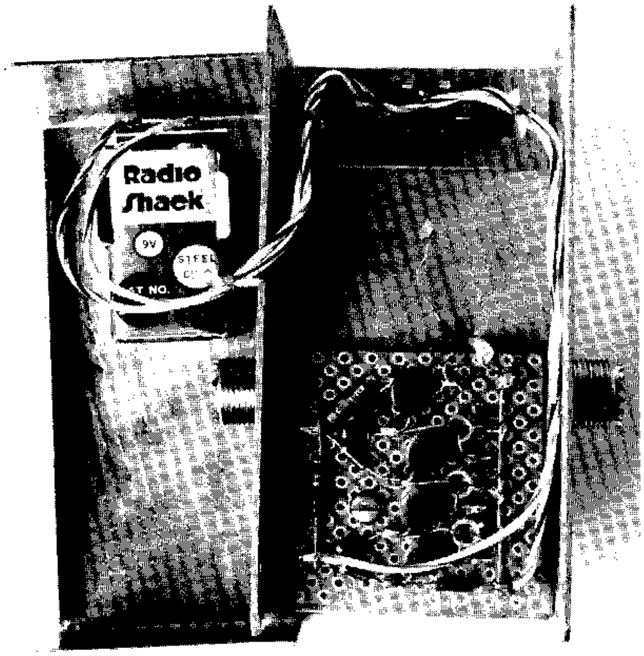


fig. 7. Interior layout of the improved noise bridge is shown in this photograph. A copper electrostatic shield encloses the bridge section. The three leads from the bridge section can be seen at the upper right-hand corner of the perf board. One is a ground lead and the other two are leads from the primary of the toroid. One primary lead is connected to the amplifier output; the other is left unattached as explained in the text.

these impedances are clearly outside the range of the noise bridge in its usual configuration. In these cases, the 100-ohm resistor technique briefly described above is required to obtain a measurement.

Table 3 shows the results of measuring these new test loads on Units 2 and 3. In all cases, the measurements were converted into the series circuit equivalent of the measured impedance. As before, the data are compared with those taken on the Boonton 250A RX-Meter. For convenience, these data were also converted to series equivalent.

There are several ways of describing the accuracy of these measurements. The rms error for each unit was about 3 ohms in both the resistive and reactive components. Considering the distribution of errors, one is 10 ohms, several are in the 5 to 7 ohm range, but *nearly half are 1 ohm or less*! While we shall purposely avoid making a global statement about the accuracy of these units, the data is encouraging well beyond our original hopes. Many high-quality instruments have accuracy expressions relating to "percent of full-scale reading;" our noise bridges appear to be in the "few" per cent category.

constructon

Figs. 7 and **8** shows the details of construction we recommend. The 2¼ by 2¼ by 5-inch (5.7x5.7x12.7cm) box allows easy construction as well as a component arrangement which minimizes lead length. A conventional 365-pF air-dielectric variable can be made to fit in this same box, although it is recommended that the more compact Archer 272-1341 capacitor be used to improve access to the actual bridge circuit wiring during calibration. The amplifier circuit, **fig. 9**, was built on a small piece of perf board in the model pictured although two other units we built used a printed-circuit board. Both construction techniques resulted in identical performance and the circuit itself is immune to layout variations. The potentiometer should be a linear taper, carbon variety such as an Allen-Bradley, type J. Wirewound pots are unacceptable because of their high inductance.

Shielding is arranged so that the battery, electronics, and on/off switch are external to the critical bridge circuit components. Copper sheeting was formed so that one end is secured under the variable resistor and the sides are clamped between the flange of the chassis-mount coax connectors and the box. The electronics board was then mounted with a single screw and spacer on the major surface of the copper shield.

Short leads are important in the bridge circuit area with the exception of the leads going through a single hole in the shield near the output end of the electronics board. Of particular importance is the detail involved with grounding. The ground lug of the variable capacitor is used as the focus of a single-point grounding system; an insulated ground wire is routed from this point to the electronics along with both ungrounded primary leads from the toroidal transformer. Both ends of the primary

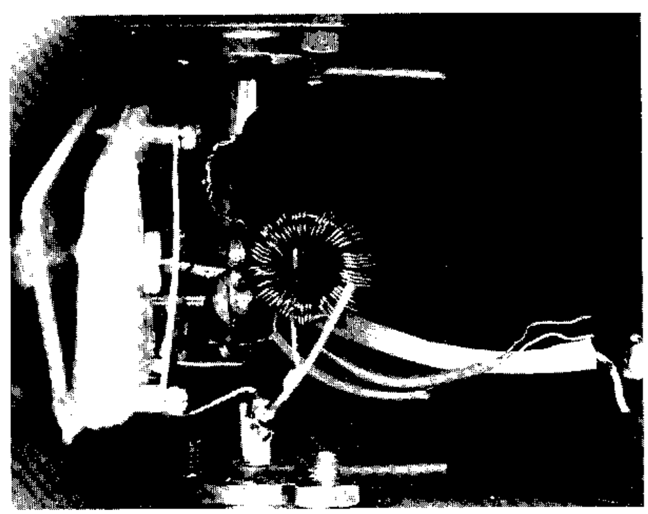


fig. 8. This closeup photograph of the bridge section shows the location of major components. The small 365 pF variable is on the left while the 250-ohm pot is in the background behind the toroid. The SO-239 coaxial connectors are on either side of the box between the capacitor and resistor, providing a compact layout. The leads to the amplifier ground and from the toroid primary, as well as the hole through the copper shield to accomodate them, are in the lower right. The series inductance balancing device can be seen emanating from SO-239 connector on the bottom.

windings will be required during the calibration sequence. Overall construction of the noise bridge is shown in **figs. 7** and **8**, and retention of this component layout is encouraged.

Construction of the 100-ohm range extender is quite simple with the use of a PL259-to-Motorola type pin plug (Archer 278-208). This unit fits conveniently into the back of a PL-259 plug (Archer 278-205). The adapter plug must be modified to allow for series connection of the 100-ohm, ½-watt carbon resistor. All but ¼ inch (6.5mm) of the Motorola plug is cut off, including the center pin. The 100-ohm resistor is soldered to the shortened inner pin and the assembly is then inserted into the PL-259 plug.

Several spots can be soldered to secure the two units together and the resistor lead projecting out of the PL-259 plug may then be soldered. The layout of the assembly is shown in **fig. 10**; the completed unit is pictured in **fig. 11**. It might be tempting to consider installing the 100-ohm range extender inside the noise bridge with a switch to more easily select its use. This was tried, with mixed results, so we recommend you use the external device if comparable accuracy is desired.

checkout and calibration

Initial checkout of the noise bridge can be accomplished by connecting the unit to your receiver and leaving the unknown coax connector unterminated. There should be generous amounts of noise anywhere between 3.5 and 30 MHz! Although the output falls off somewhat at the higher frequencies, an S-9 signal should be expected on 10 meters. No null can be expected under these conditions. By varying the R and C dials, however, some variation in the S-meter reading is normal. Failure to observe these results requires checks of wiring and measurement of bias levels on the zener and transistor terminals. Access to an oscilloscope is helpful for signal tracing, but it was our experience that voltmeter measurements are quite satisfactory.

The next step in the checkout is to see if nulls can be achieved and that the bridge circuit is basically working. Connect the range extender assembly to the *unknown* coax connector and short the output of the 100-ohm resistor to ground through a physically short connection. By placing the receiver on 80 meters with a short agc time constant, if available, it should now be possible to find an extremely sharp null on the S-meter by adjusting both the R_p and C_p dials. Since there will be some interaction between the adjustments of the dials, the sequence must be repeated until the minimum S-meter reading is obtained. This null should be all but absolute, that is, near zero on the S-meter.

Since the unknown being tested is near 100 ohms with little reactance, the resistance dial, which is linear, should read approximately 40 per cent of scale from the zero position, and the capacitance dial would be expected to read near the center of its range. Inability to obtain these results indicates a problem in the bridge circuit and transformer connections should be checked.

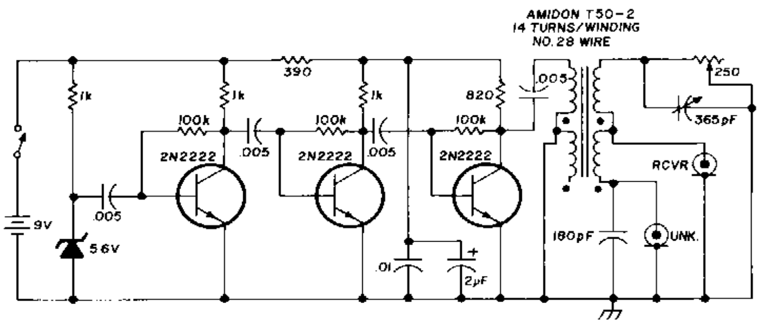


fig. 9. Schematic diagram for the improved noise bridge unit. Construction details are discussed in the text.

Offset of the capacitance dial from approximate center would indicate difficulty with the 180 pF fixed capacitor.

Assuming a successful checkout, you can now proceed with calibration. Preparation includes lightly steel-wooling the paint under the R and C dials to remove the gloss and permit the use of India ink or a fountain pen for scale markings. Later, a clear alcohol-based artist-type aerosol spray can be applied to protect the markings and make them more durable. Some clear sprays interact with ink, so a preliminary try is advised to avoid an upsetting experience after you have finished the calibration sequence.

With both the receiver and unknown disconnected, the resistance dial can be calibrated by attaching an accurate ohmmeter to either coax connector and ground. Surprisingly, dc resistance readings and accompanying marks on the resistance scale are very accurate from 80 through 10 meters. Tick marks every 10 ohms are a good balance between readability and scale appearance. Lacking an accurate ohmmeter, it might be advisable to locate several 1 per cent resistors in the 10 to 250 ohm range for reference.

Calibration of the capacitance dial is most easily accomplished with a capacitance meter. The variable capacitor must be temporarily disconnected from the rest of the bridge. Establishing the 180 pF dial setting as zero, the scale can then be marked in both directions to an "end-of-scale" reading of approximately 180 pF in either direction. Tick marks every 10 pF are desirable; the most clockwise 180 pF value should be marked with an L to indicate that inductive reactance is in that direction.

Lacking access to a capacitance meter, calibration of the C dial requires the following sequence: Connect the noise bridge to your receiver and select the 80-meter band. A 100-ohm, ½-watt carbon resistor and a selection of accurate capacitors in the 10 to 180 pF range should be at hand. Start with the 100-ohm resistor only and attach it to the unknown coax connector with short leads. Establish a null by observing the receiver S-meter and adjusting both the R and C dials. The resulting position of the C dial can then be marked as zero. By adding small values of capacitance in parallel with the

100-ohm resistor and rebalancing the bridge each time, the dial can be marked progressively until the 180 pF value is reached.

The next step is to temporarily remove the 180 pF fixed capacitor inside the noise bridge. Returning to the 100-ohm resistor measurement at 80 meters, capacitance should be added in parallel until the C dial reads zero as marked in the earlier step (approximately 180 pF will be required). Small values of capacitance should then be removed from the 100-ohm resistor until the scale has been marked as desired. Since the C scale is quite linear, some liberties may be taken in the form of estimating intermediate scale markings as long as accurate capacitors are used during the calibration process.

Following the scale marking sequence, all circuits can be returned to their original state in readiness for compensation adjustments which will assure accurate performance over the 3.5 to 30-MHz frequency. A 150-ohm, ½-watt carbon resistor should be placed on the unknown coax connector with short leads. The noise bridge may then be nulled on 3.5 MHz; both the resistance and capacitance dial should read fairly accurately at this point (150 ohms, 0 pF). Possibly the C dial might read slightly off zero at this time and the knob can be moved on the capacitor shaft to correct this situation. Moving your receiver to 30 MHz, again measure the 150-ohm resistor. If you get the same readings as you did at 3.5 MHz, the effort is over and you have a good instrument!

More than likely you will see a small shift at 30 MHz. If the R_p dial reads higher, and the C_p dial reads on the inductive side of zero, then your bridge needs additional series inductance on the unknown side. We recommend that you solder a short piece (about 1 inch or 25mm) of bare hookup wire to the SO-239 coaxial connector

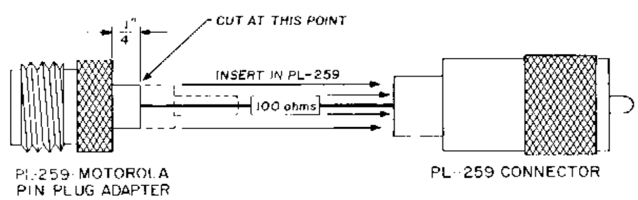


fig. 10. Construction details for the 100-ohm range extender unit. A PL-259 to Motorola pin plus adapter is first prepared by cutting off all but ¼" (6.5mm) of the pin and shank. A ½ watt, 100-ohm resistor is then inserted in the pin of this adapter and soldered. The shortened shank and resistor are then inserted in the back of the PL-259 plug. The other resistor wire is soldered to the center pin of the PL-259 connector. Soldering the exterior surfaces together completes the assembly.

marked *unknown*. The fixed C_f, 180 pF capacitor should remain soldered at this connector with a short lead to ground. The wire from the transformer secondary should be removed from the SO-239 connector and fastened instead to the 1 inch (25mm) piece of wire. Connect it to the far end of this wire and rerun you balance test.

You should now find that your error at 30 MHz has reversed (instead of R_p reading too high, it will now be too low). In addition, the C_p reading should be on the capacitive side of zero. If this is the case, move the wire from the transformer a little closer to the SO-239

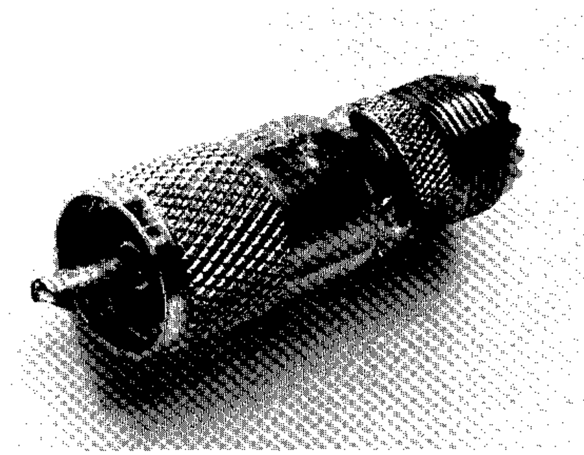


fig. 11. The completed range extension assembly.

connector along the 1 inch (25mm) wire and rerun your test. You should find a point where the R_p and the C_p dials read very nearly the same at 30 MHz as they did at 3.5 MHz. When they do, your work is complete.

Suppose, instead of R_p too high and C_p inductive at 30 MHz, your first measurement (without the 1 inch or 25mm wire) indicates the opposite. This means that your bridge has too much inductance on the unknown side. In this case, instead of trying to add wire to the known side it is easier to reverse the transformer primary windings to shift the error in the other direction. Then you can proceed to finish the calibration as discussed above.

Suppose you find a high-frequency shift that is different from those discussed above. Perhaps R_p increases and C_p also increases (goes capacitive). If this is the case, it indicates your unbalance is more than just a problem of unbalanced series inductance and you should check your layout, particularly the winding and placement of your transformer. Try and make this layout as clean as possible. Make sure your shielding is effective and that a chassis ground loop does not exist. If all this is done properly, the residual unbalance should be such that it can be compensated with the series inductance technique.

applications

There are a number of applications in the shack which are ideally suited for the noise bridge. The first and most obvious is to use it to measure antenna impedance. Since the measurements will be made with the noise bridge located near the receiver, you will want to transform the values measured to those that apply to the antenna's feedpoint. Again, the Smith chart is recommended and W1DTY's article or the latest issue of the ARRL *Antenna Handbook* discusses the technique in detail.[3,4] An example of the results that might be obtained are shown in **table 4** and **fig. 12.** These data are from an actual 80-meter inverted-vee antenna with isolation traps resonant on 40 meters. It is fed with 60 feet (18.3m) of RG-8/U coaxial cable.

The actual mathematical steps involved in noise bridge impedance measurements are quite easy and relatively fast once the pattern is established. First, record the bridge readings; then compute the reactance of the parallel capacitor. The parallel circuit elements are then converted to series equivalent elements. These transformations are performed using the following equations:

$$X_p = \frac{-159{,}000}{f\,C_p}$$

where f = frequency in MHz

C_p = capacitance in pF

For the 4.0-MHz data point,

$$X_p = \frac{-159{,}000}{4.0(-84)} = 473 \text{ ohms}$$

Converting to series equivalent values required use of the following equations:

$$R_s = R_p \frac{X_p^2}{R_p^2 + X_p^2} \qquad X_s = X_p \frac{R_p^2}{R_p^2 + X_p^2}$$

Substituting 4.0 MHz values of R_p and X_p,

$$R_s = 117 \frac{(473)^2}{(117)^2 + (473)^2} = 110 \text{ ohms}$$

$$X_s = 473 \frac{(117)^2}{(117)^2 + (473)^2} = 27 \text{ ohms}$$

If the range extender was used, then actual values of the 100-ohm resistor must be subtracted from R_s before it is entered in the corrected series impedance column.*

At this point, the equivalent series representation may be plotted on a Smith chart, making allowance for the rotation required due to the electrical length of the transmission line used. The electrical length is easily established by using the velocity factor (v) for the coax in use (66 per cent for RG-8/U) and the following relationship

table 4. The noise bridge measured impedance of an actual 80-meter inverted vee antenna. Starting with the recorded values of R_p and C_p, the data are converted to equivalent series impedance, and finally "rotated" through the 60-foot (18.3m) length of RG-8/U used to feed the antenna. The final column tabulates the measured impedance of the antenna at its feedpoint.

frequency MHz	noise bridge recorded data R_p	C_p	calculated capacitor reactance X_p	series impedance R_s	X_s	100-ohm resistor used	corrected series impedance R_s	X_s	transmission line length	antenna impedance R_s	X_s
3.50	149	−163	279	116	62	yes	16	62	0.323λ	30	−95
3.55	164	−163	275	121	72	yes	21	72	0.328λ	27	−85
3.60	202	−141	313	143	92	yes	43	92	0.333λ	25	−65
3.65	240	−98	445	186	100	yes	86	100	0.337λ	24	−50
3.70	129	−66	651	124	25	no	124	25	0.342λ	26	−31
3.75	78	152	−279	72	−20	no	72	−20	0.346λ	33	−9
3.80	144	20	−2092	143	−10	yes	43	−10	0.351λ	41	9
3.85	121	−9	4589	121	3	yes	21	3	0.356λ	48	45
3.90	118	−39	1045	117	13	yes	17	13	0.360λ	65	70
3.95	116	−65	619	112	21	yes	12	21	0.365λ	75	105
4.00	117	−84	473	110	27	yes	10	27	0.370λ	95	140

$$\lambda = \frac{\ell f}{984 v} \text{ wavelengths}$$

where ℓ = physical length of coax in feet

f = frequency in MHz

For 4.0 MHz

$$\lambda = \frac{(60)(4.0)}{984(.66)} = 0.370 \text{ wavelength}$$

The equivalent series impedance components must, of course, be normalized to the Z_o value (division by 50 in this case) prior to plotting on the Smith chart. Construction details for the 4.0 MHz data point are shown in **fig. 12.** Although the influence of feedline loss could have been included, the degradation is small on 80 meters and, therefore, was ignored. The final results as presented by **fig. 12** allow considerable insight into the workings of this antenna:

1. Resonant frequency is 3.78 MHz.
2. Feedpoint impedance at resonance is 37 ohms.
3. Bandwidth with swr ⩽ 2:1 is 100 kHz.
4. Bandwidth with swr ⩽ 3:1 is 200 kHz.

With an antenna such as this one, you should not expect to load your transmitter at the band edge with-

*You should measure the actual value of your 100-ohm resistor by shorting the output end of the range extender and nulling your noise bridge.

out causing some sparks in your final tank circuit! A transmatch must be used to make the transmitter happy. Then there is the problem of properly tuning the transmatch at the frequency of interest. A good way to do this is to place the noise bridge at the transmatch input terminal and set the dials to R_p = 50 ohms and C_p = 0. Then by turning the transmatch dials, look for a null.

If a null can be found, the resulting swr at the transmitter will be very close to 1:1. And the best part is that you've done this without radiating any power and causing interference. It's a good idea to log these results for future reference. Be cautious, however, with this test setup, because it's possible, by hitting the wrong switches, to apply power to the antenna through the noise bridge. If you do this, even for an instant, you'll be in the market for a new 250-ohm pot and perhaps a new transistor in the amplifier circuit.

The particular antenna discussed earlier did not have a balun installed at the time these measurements were taken. In trying to find a suitable core to wind a balun, we discovered another application for the noise bridge. A core of appropriate dimensions but uncertain ancestry was available. In practice the resulting balun transformer didn't work. To discover why, a 50-ohm carbon resistor was soldered across the secondary of the balum and the input impedance measured with the noise bridge. With a good 1:1 balun we would expect to see an R_p of approximately 50 ohms in parallel with a negative C_p representing the reactance of the primary of the transformer. If this parallel reactance is large with respect to 50 ohms, the transformer works as it should. In this case we found a very high impedance primary circuit, beyond the range of the bridge. This indicated that the core material was not designed for the frequency range desired so a different core had to be obtained.

Another pet application involves power meters. We have built several of these and have wondered about their accuracy. If you have both a 50-ohm dummy load and a transmatch, you can use the noise bridge to synthesize arbitrary transmitter loads of interest. Unless you spent a lot of money for your dummy load, it probably is not exactly 50 ohms at any frequency. If you used it to set the null on your power meter, that null will not be the best that can be obtained. By using the transmatch/dummy load combination, you can synthesize an accurate 50-ohm impedance.

In addition, do you have confidence that when your power meter indicates a 2:1 swr that this is in fact the case? With the above hardware, you can synthesize an rf impedance of either 100 ohms or 25 ohms. Both should yield an swr of 2:1 on a 50-ohm coaxial cable. You might try this to see if your power meter indicates the correct values.

The last application we shall discuss effectively demonstrates the extreme range capability of the instrument when using the 100-ohm resistor modification. It is often desired to know the inductance of an rf coil. With air-core coils having uniform dimensions, it is possible to accurately compute inductance using

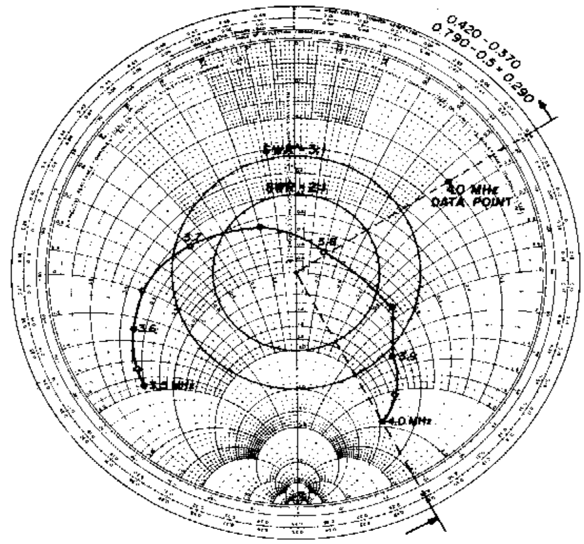


fig. 12. Smith chart plot of the noise bridge measured impedance of an actual 80-meter inverted vee antenna. The construction of the 4.0 MHz data point is shown for illustration. The measured impedance at 4.0 MHz must be rotated 0.370 wavelength "toward the load" to account for the 60′ (18.3m) RG-8/U transmission line. 2:1 and 3:1 swr circles illustrate the bandwidth of the antenna. The bandwidth is relatively narrow since the antenna is physically short (about 100′ or 30m long) and relies on the inductive loading effect of 40-meter traps to achieve resonance on 80 meters.

standard formulas. These formulas are much less precise and are more difficult to apply to inductors wound on a core such as a toroid. However, using the rf noise bridge, it is very easy to directly measure inductance in the "few" microhenry range.

By nulling the noise bridge with the 100-ohm resistor in the circuit and the unknown coil attached, observe the R_p and C_p readings. If a reading is not possible, try a lower frequency. The C_p reading should be negative, of course, indicating the inductive reactance of the load. The inductance of the unknown coil in microhenries is calculated by

$$L = \frac{-R_p C_p}{10{,}000} \ \mu H$$

We leave it as an exercise to mathematically inclined and interested readers to derive this equation. Most, I'm sure, would prefer to prove it by demonstration with a real rf coil. We earnestly invite you to build the rf noise bridge as described here so you can perform the demonstration yourselves!

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