Summer 1954

Number 2

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

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A Wide-Range VHF Impedance Meter

JOHN H. MENNIE, *Senior Engineer*

The design of a completely self-contained instrument for measuring impedance at Very High Frequencies entails, as might be expected, a number of interesting engineering problems. Mr. Mennie describes some of those encountered and solved in the development of the BRC RX Meter Type 250-A.

Ever since the middle of the last century when Sir Charles Wheatstone first put to practical use a curious electrical balancing network which had been devised a decade before by a fellow Englishman named Christie, the bridge circuit has been accepted as a valuable and powerful tool in the field of electrical measurement.

Even though today bridges, like almost everything else, have become increasingly complex in order to meet the demands of highly specialized applications, they retain, for the most part, the fundamental advantages of convenience, sensitivity and accuracy which characterized their more straightforward antecedents.

One of the applications in which bridge circuits have been particularly useful in recent years is the measurement of impedances at radio frequencies. A number of specialized circuits have been evolved for this purpose. One, the Schering Bridge, possesses certain features which make it outstanding. These features may be summarized as follows:

1. A constant relationship between the bridge elements is maintained regardless of the frequency impressed on the network.
2. Both of the basic variable bridge elements can be air capacitors, which are infinitely superior to other types of variable impedances for high frequency measurement work.
3. The circuit residual impedance can be kept small enough to permit compensation over a wide frequency band.
4. When arranged to measure parallel components of impedance, shielding problems are drastically reduced.

The fundamental circuit was first worked out by Schering in 1920 and proposed as a means for measuring dielectric losses at high voltages. Many other applications of this circuit have been suggested and used since then. For instance, in 1933 Dr. E. L. Chaffee suggested a form of Schering Bridge, as a method for measuring the dynamic input capacitance and resistance of vacuum tubes.

BALANCE EQUATIONS

The simplicity and wide frequency range of this bridge network can be appreciated by an analysis of the impedance relationships of Fig. 2 for the balance condition (i.e., zero voltage across the null detector).

$Z_{AB}\,Z_{CD} = Z_{AD}\,Z_{BC}$ at balance, or

$$\left(R_2 + \frac{1}{j\omega C_2}\right)\left(\frac{1}{\frac{1}{R_4} + j\omega C_4}\right) = \frac{R_3}{j\omega C_1}$$

$$R_2 + \frac{1}{j\omega C_2} = \frac{R_3}{j\omega C_1}\left(\frac{1}{R_4} + j\omega C_4\right)$$

$$= \frac{R_3}{j\omega C_1 R_4} + \frac{C_4 R_3}{C_1}$$

Equating reals...

$$R_2 = \frac{C_4 R_3}{C_1}, \text{ and } \frac{R_2}{C_4} = \frac{R_3}{C_1}$$

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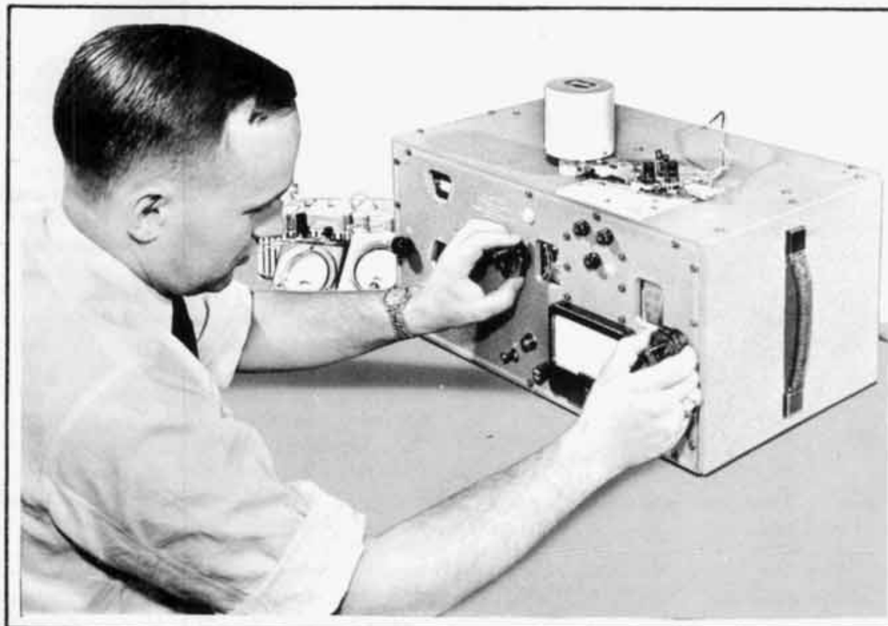

Figure 1. The RX Meter provides a simple, accurate means of measuring, independently, the RF resistance and reactance of a wide variety of materials, components and circuits. Dynamic measurements are possible, such as this one being made by C.G. Gorss, BRC Development Engineer, on a junction transistor.

YOU WILL FIND...

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Impedance Meter (continued)

Equating imaginaries...

$$\frac{1}{j\omega C_2} = \frac{R_3}{j\omega C_1 R_4}$$

$$\frac{R_3}{C_1} = \frac{R_4}{C_2}$$

$$\therefore \frac{R_2}{C_4} = \frac{R_3}{C_1} = \frac{R_4}{C_2} \qquad (1)$$

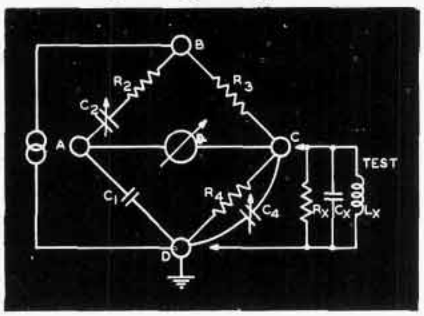

Figure 2. The Schering Bridge, arranged to provide measurement of parallel impedance components.

The test sample is connected across corners C and D of the bridge, and its parallel components of resistance and reactance effectively change the values of C4 and R4 in the circuit. In order to restore phase and amplitude balance conditions, the variable bridge capacitor C4 must be decreased by an amount equal to the equivalent parallel capacitance of the test sample. If the test sample is inductive, the capacitance of C4 is increased by an amount equal to the resonating capacitance of the parallel inductance.

The parallel resistance of the test is shunted across R4, reducing its value by a certain percentage which changes the R4/C2 ratio and unbalances the bridge. To restore phase and amplitude balance, variable capacitor C2 is reduced in value by the same percentage that R4 was reduced when shunted by the test resistance. The variable capacitor C_2 can thus be calibrated directly in terms of the parallel resistance (in ohms) of the component being measured.

THE RX METER

A refined version of the basic circuit described above forms the heart of the BRC RX Meter Type 250-A, which is designed to measure parallel resistive and reactive impedance components at frequencies from 0.5 mc to 250 mc. Unlike most RF impedance measuring devices using the null technique the RX Meter is a self-contained instrument including with the bridge circuit its associated oscillators, detector, amplifier and null indicator.

Figure 3 shows a block diagram of the instrument, the operation of which may be briefly described as follows:

The output of the test oscillator, (F1), which covers a frequency range of 0.5 to 250 megacycles, is fed into the bridge circuit. When the impedance to be measured is connected across one arm of the bridge, its parallel resistance and reactance components cause bridge unbalance, and the resulting voltage is applied to the mixer stage. The output of the local oscillator, (F2), which tracks at a frequency 100 kilocycles above F1, is also applied to the mixer, where it heterodynes with the bridge unbalance output frequency, F1, and produces a 100 kc difference frequency, having a magnitude proportional to the bridge unbalance voltage. This voltage is then amplified by a selective 100 kc amplifier to provide the desired bridge balance sensitivity. When the bridge controls are adjusted for balance (minimum indication), their respective dials serve as an accurate indication of the parallel impedance components of the test sample.

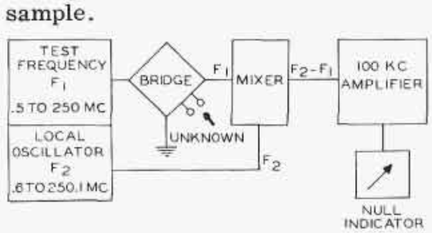

Figure 3. Block diagram of the RX Meter Type 250-A.

Particular care has been taken, in the mechanical design of the instrument, to provide adequate shielding of the oscillator and mixer in order to prevent spurious coupling between these stages. The detector is provided with automatic gain control which prevents the meter from reading off scale, thus greatly facilitating the balancing operation when approximate values of the impedance being measured are not known.

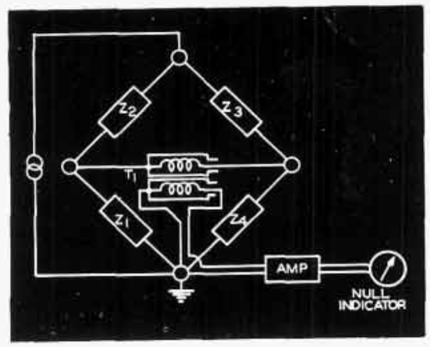

Figure 4. Conventional transformer coupling from bridge corners to amplifier.

Use of rugged castings, precision bearings and anti-backlash gears has permitted expansion of the calibrated parallel resistance scale to a useful length of twenty-eight inches, covering from 15 ohms to 100,000 ohms. The "R_p" scale of each instrument is individually calibrated and engraved in spiral form on the surface of a drum dial, the readability being one percent up to 5,000 ohms with an accuracy of approximately two percent over this range.

BRIDGE COUPLING CONSIDERATIONS

The absence of frequency terms in the balance equation (1) above suggests the applicability of this bridge circuit to impedance measurements over an almost unlimited frequency range. Several practical limitations do exist, however, and their effects must be considered in establishing the upper frequency limit for such a device. One of these involves the means used for connecting the signal source and null detector amplifier to the bridge. If the connection introduces excessive capacitance across the bridge arms or excessive loss in the oscillator or detecting system it can seriously affect the operation of the instrument at higher frequencies. This problem can be more readily appreciated by a study of Figure 4, which shows a conventional connection between the high bridge corners and the null detector, employing a special type of double-shielded transformer, T1.

A transformer of this type not only is difficult to design and expensive to manufacture but also has an upper frequency limitation imposed by leakage inductance as well as the usual low frequency limitation resulting from low impedance of the windings. The inter-shield capacitance is another handicap, as it must be incorporated in the bridge network, thus placing a severe limitation on both frequency and impedance ranges of the bridge. These shortcomings have been overcome by a new approach to the problem, as shown in Figure 5.

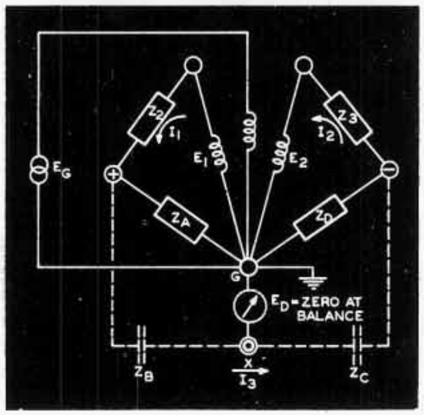

Figure 5. Specially devised system for coupling oscillator to detector.

The essential feature of this bridge network is that it is divided in two halves, one half being driven by voltage E_1 and the other half by E_2. Let us assume that voltages E_1 and E_2 are exactly equal in magnitude but opposite in phase, thus producing instantaneous currents I_1 and I_2 in the direction indicated. This arrangement enables us to detect the bridge balance conditions by coupling to the null detector through two very small but exactly equal capacitors, Z_B and Z_C. In the balanced condition, voltage E_D becomes zero, as indicated by our null detector, and the voltage across Z_B is exactly equal to that across Z_C. This is evident, since the same current, I_3, flows through both Z_B and Z_C when zero current is drawn by the detector branch. As X and G are at the same potential, it follows that Z_B may be considered effectively in parallel with Z_A, and Z_C is likewise in parallel with Z_D. Thus the voltage across Z_A is equal to that across Z_D. Now let Z_A and Z_B in parallel = Z_1 and Z_C and Z_D in parallel = Z_4. Figure 6 represents a simplified form of the bridge network for analysis purposes.

In Figure 6

$$E_1 = I_1Z_2 + I_1Z_1$$

$$E_2 = I_2Z_3 + I_2Z_4$$

$$E_1 = E_2 \text{ (by design)}$$

and $I_1Z_1 = I_2Z_4$ at balance

then $I_1Z_2 = I_2Z_3$

$$\therefore \frac{I_1Z_1}{I_1Z_2} = \frac{I_2Z_4}{I_2Z_3} \text{ and } Z_1Z_3 = Z_2Z_4 \qquad (2)$$

This impedance arm relationship is the same as that of a conventional bridge network.

COUPLING TRANSFORMER DESIGN

The above theory is based on the assumption that the transformer secondary voltages be extremely well balanced, and it is essential that this be the case over the entire frequency range of 0.5 mc to 250 mc. A very simple transformer was developed (Figure 7) consisting of three Formex-insulated wires twisted tightly together and wrapped once around a ferrite core ring. By using one turn the shunt capacitance and leakage inductance are held to a minimum for best high frequency performance. Leakage inductance is only 0.014 microhenry. The high permeability ferrite core serves to hold up the impedance of the transformer at the low frequency end. Interwinding capacitance C_{12} is ineffective because 1 and 2 are equipotential points. Although the transformer impedance is quite low, it effectively matches the very low output coupling loop impedance of the oscillator and drives the bridge with approximately one volt over the entire frequency range from 0.5 to 250 mc.

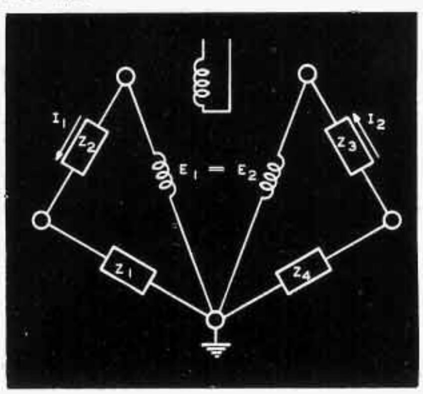

Figure 6. Simplified bridge network.

COUPLING CAPACITOR DESIGN

The capacitive network used to couple the unbalance output of the bridge to the detector must necessarily be very accurately balanced. To achieve such balance the following requirements were found to be essential:

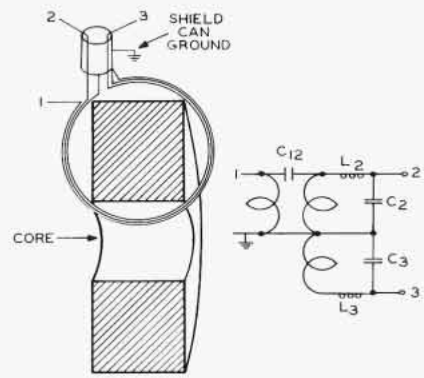

Figure 7. Balanced coupling transformer design.

1. Extremely fine adjustment to $\pm 0.001\ \mu\mu f$.
2. Loss factor uniformly small for both capacitors.
3. Temperature coefficient uniformly small for both capacitors.
4. Mechanical stability to stand vibration and aging without change of balance.
5. Complete shielding to avoid pick-up to grid of detector from any other source than the two bridge corners.
6. Equal lead inductance to maintain effective capacitance balance to 250 mc.

The dual capacitor unit which was specially designed to fulfill these requirements is illustrated in Fig. 8.

RESIDUAL BRIDGE IMPEDANCES

As previously discussed, the bridge will balance at all frequencies, providing proper relationship between the effective capacitance and resistance is maintained up to the highest frequency to be used. Serious deviations will occur, however, if residual impedances are not either compensated for or made negligible by design.

Figure 9 shows two of the most troublesome residual inductances, L2 and L4. L2 represents the total inductance of series condenser C2

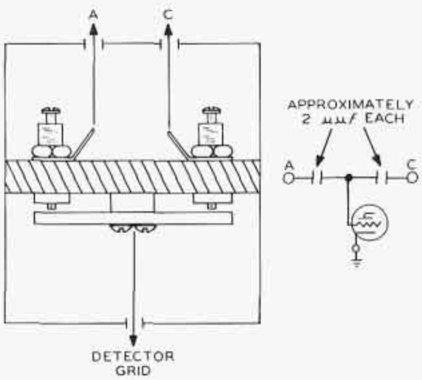

Figure 8. Balanced coupling capacitor design.

and resistor R2 and has a value of approximately 0.03 microhenry. This is sufficient reactance to completely neutralize the effective capacitance of the 20 μμf condenser C2 at 200 mc. Fortunately this residual inductance, L2, can be effectively removed from the circuit by means of a small shunt capacitance, Cp, across resistor R2. This shunt capacitance, as shown below, is equivalent to a series capacitance (C_s) whose reactance is very nearly equal and opposite to that of L2 up to the maximum frequency of 250 mc.

$$Z \text{ of parallel } R_2 C_p = \frac{R_2 - j\omega C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2}$$

The series capacitance reactance

$$\frac{1}{j\omega C_s} = \frac{-j\omega C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2}$$

For neutralization, let reactance

$$j\omega L_2 = \frac{-1}{j\omega C_s} = \frac{j\omega C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2}$$

$$\text{then } L_2 = \frac{C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2} \cong C_p R_2^2$$

(within 2.4% at 250 mc.)

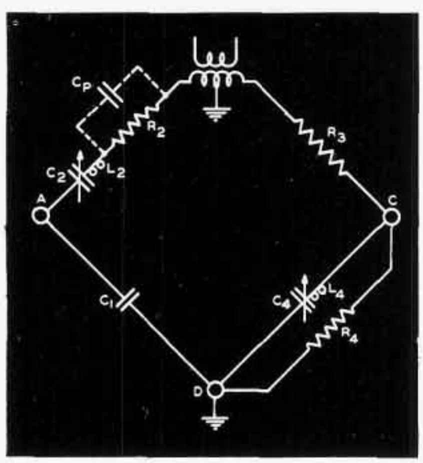

Figure 9. Residual impedance in bridge arms.

The second troublesome residual is the inductance, L4, associated with the variable condenser standard, C4. This inductance not only causes an effective capacitance increase in C4 at high frequencies but also produces an error in the resistance readings of high Q reactive components. The latter effect is the result of coupling between L4 and the inductive component of resistor R4. The best solution to this problem was found to be a series of fourteen specially designed edge-wiping rotor contact springs which provide fourteen individual parallel inductance paths through the condenser, each path having extremely low inductance. By this means, the inductance L4 is reduced to a value of only 0.0005 microhenry.

APPLICATION

The RX Meter has been found valuable in a number of diversified industrial and experimental applications. For example, it is being used to measure the RF characteristics of balanced and unbalanced transmission lines, of electrical components such as resistors, and of high and medium loss insulating materials such as phenolic vacuum tube bases.

The direct measurement of equivalent parallel resistance (R_p) of a circuit is particularly useful, since it represents the impedance seen by a vacuum tube or transistor when working into the circuit. Measurement of R_p also facilitates the determination of power dissipation in a tuned circuit since, when the voltage E across the tank is measured, power dissipation = E^2/R_p.

The RX Meter is particularly applicable to the measurement of transistor impedance since DC bias currents up to 50 milliamperes can be passed directly through the sample binding posts without harming the bridge elements. In such applications it is possible to lower the RF test voltage to as low as 20 mv with useable sensitivity.

CONCLUSION

The bridge methods conventionally used for RF impedance measurements generally require a composite test circuit consisting of a number of separate interconnected instruments. Setting up such a circuit for specific measurements is usually tedious and time consuming; leakage and undesirable coupling are frequently a problem at higher frequencies, and the resulting circuit is often limited in range and application. By combining all the necessary test components in a single, integrated instrument, the RX Meter provides ease and rapidity of measurement as well as unusual flexibility of application.

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3. "A Radio Frequency Bridge for Impedance Measurements from 400 kc to 60 mc", Proceedings of the IRE, Nov., 1940.
4. U. S. Patent No. 2,607,827.

THE AUTHOR

John Mennie joined BRC as a Senior Engineer in 1951. His activities include the development of new instruments and equipment from idea to working model stage, and he is personally responsible for the basic design of the RX Meter Type 250-A.

Prior to 1951 he was employed as a design and development engineer at Western Electric Co., where he worked in close conjunction with Bell Laboratories on the development of specialized manufacturing techniques and a variety of measuring and test equipment.

Mr. Mennie was graduated from Stevens Institute of Technology in 1929 with a degree in Mechanical Engineering. He is a member of the Industrial Electronics Committee of I.R.E.

NOTE FOR OWNERS OF Q METERS TYPE 260-A

Final Instruction Manuals for this instrument are now off the press. If you have not received your copy please let us know. It will be forwarded immediately.

Q Meter Comparison

The recent redesign of the Q Meter Type 160-A resulted in a more accurate and flexible instrument - the Type 260-A. Here is a detailed discussion of the changes which were made and how they affect comparative Q readings on the two instruments.

When the Q Meter Type 260-A was introduced in 1953, its predecessors, the Types 100-A and 160-A, had been giving satisfactory laboratory service for a period of over 18 years, and had long since become a standard for Q measurements.

In spite of the fact that, during recent years, the Type 160-A had proved itself a useful and dependable tool, it had become apparent, in the light of new developments and improvements in the art, that its circuit had certain limitations. For this reason, the Q Meter Type 260-A was developed, retaining all the valuable characteristics of the 160A, but incorporating features which eliminated or minimized the limitations of the older instrument.

Because these design improvements will often cause small variations in Q readings between the two instruments, it seems desirable to investigate in detail the reasons for such apparent discrepancies and to formulate an expression for predicting them.

DESIGN DIFFERENCES

Both the 160-A and 260-A use the same basic measuring circuit, and the block diagram shown in Figure 1 applies to either instrument. This design, which makes use of the low impedance system of injecting voltage into the measuring circuit, is based on the resonant voltage rise principle described in a previous article.[1]

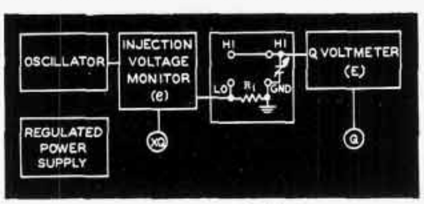

Figure 1. Q Meter block diagram.

Thus (for $Q \geq 10$) $Q = E/e$, where Q is the quality factor of the measuring circuit, e is the injected voltage and E is the voltage across the resonating capacitor.

The principle requirements for a successful Q measuring system are,

1. A low-distortion oscillator.
2. An accurate system for monitoring the injection voltage.
3. A low injection impedance across which e is developed.
4. A low-loss internal resonating capacitor.
5. An accurate high impedance Q voltmeter for measuring E.

Each of these must be carefully designed to minimize the residual parameters which tend to make the indicated Q (of the Q measuring circuit) differ from the effective Q of the unknown. Primarily, the discrepancies in Q readings which may be found between the two Q Meters at certain frequencies are caused by those improvements in the design of the 260A which result in lowered residual parameters.

SOURCES OF ERROR

1. Oscillator Harmonics

The calibration of the Q voltmeter is dependent on accurate indication of the ratio E/e. Errors in the Q readings can be caused by harmonics present in the oscillator output, due to the fact that the thermocouple used to monitor the current in the injection resistor produces a meter deflection proportional to the total heating effect of the distorted current, while, because of the selectivity of the measuring circuit, the indication of the Q voltmeter is proportional only to the fundamental component of the oscillator output. Although this effect is considerably reduced because of the square law response of the thermocouple,[2] it may still be noticeable when distortion is present. Early models of the 160-A (serial numbers below 4581) had some harmonic distortion in the oscillator output on the low frequency band (around 150 kc), which did introduce some error due to this effect. The effect is negligible in the 260-A because of the extremely low distortion present.

2. Injection Resistor

The injection resistor (R_i in figure 1) is in series with the resonant circuit, and when components are measured having low equivalent series resistance (high Q at high frequencies), the value R_i, which adds to the resistance of the unknown, becomes a large part of the total series resistance. This causes the Q Meter to indicate a Q value lower than that of the unknown component. Evidently it is desirable to keep the injection resistance as low as possible. For this reason, the 0.04 ohm injection resistor of the 160-A was reduced to 0.02 ohm in the 260-A.

It is possible for the reactive component of the injection resistance to cause a rise in injection voltage at high frequencies. Since the injection current is monitored, a constant e depends on a constant injection impedance, and it is important that the reactive component be reduced to a minimum. The dashed curve in figure 2 indicates the percent of rise in injection voltage caused by the residual 83 $\mu\mu$hy inductance present in the injection resistor of the 160-A. A new annular type of resistor has been used in the new Q-Meter, resulting in a residual inductance of the order of 0.035 $\mu\mu$hy, and reducing the error due to this inductance to a negligible amount, as indicated by the solid line curve.

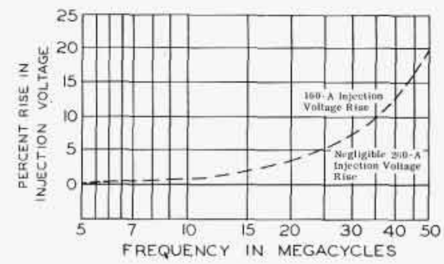

Figure 2. Calculated effect of injection resistor inductance on 160-A injection voltage at higher frequencies. 260-A resistor has negligible reactance, eliminating this source of error.

3. Resonating Capacitor

The construction of the resonating capacitor used in the Q Meter Type 160-A represents a satisfactory compromise between minimized residuals and the largest practicable capacitance range. As a result residual parameters are small enough to be ignored in the course of most measurements. This optimum mechanical design was retained in the new Q Meter and any residuals which exist may be regarded as the same for both instruments.

4. Voltmeter Circuit

The problem of providing an accurate voltmeter which will measure the RF voltage across the resonating capacitor to an accuracy of ±1% is a major one in the design of a Q Meter. In addition to maintaining the linearity (or scale calibration) characteristics, the voltmeter tube must have a very low grid current (with a grid leak of 100 megohms), the input capacity must remain nearly the same for different tubes, and the input conductance of the voltmeter circuit must be as high as possible.

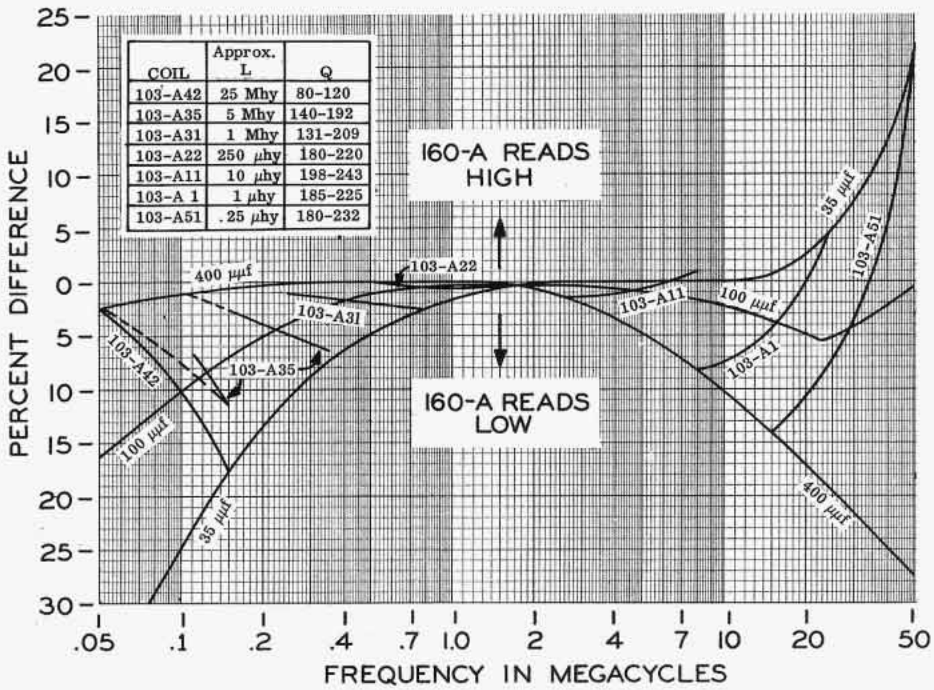

Figure 3. Difference in Q readings obtained by measuring a set of 103-A inductors on both 160-A and 260-A, with the readings of the latter used as a base. Each line labelled with a coil number indicates the percent difference in readings obtained by measuring the same coil, throughout its resonant frequency range, on both instruments. The lines labelled 35 uuf, 100 uuf, and 400 uuf respectively, indicate the percent difference in Q readings resulting at these internal resonating capacitances.

In order to achieve and maintain these characteristics, the use of a special tube is found to be imperative. The BRC 105-A is manufactured and tested specifically for application in the Q voltmeter circuit. A review of available tube types during the development of the Type 260-A confirmed the fact that this was the only tube which could meet such requirements, and it was therefore included in the design of the new instrument.

Several other changes in the voltmeter circuit were found desirable. The addition of "Lo Q" and "ΔQ" scales in the new instrument necessitated a slightly different operating point for the voltmeter tube. In addition, the physical arrangement of the grid circuit was changed to provide a higher natural resonant frequency, and additional bypassing was added in the plate circuit of the voltmeter, causing the Type 260-A to indicate Q more accurately at low frequencies and at low capacitance settings of the resonating capacitor.

TYPICAL EXPERIMENTAL DATA

To illustrate the differences which might be expected in Q readings between the Q Meters Types 160-A and 260-A, a set of standard inductors Type 103-A was measured on two representative instruments which were selected to be as nearly average as possible after a careful survey of Q readings on several hundred production units. Figure 6 is a plot of the readings obtained, and figure 3 indicates the percent difference in these readings vs frequency. Note that these graphs apply only to inductors with Q values in the ranges indicated. In general, the percent difference at any frequency is proportional to Q.

FORMULA FOR PREDICTING Q READING DIFFERENCES

The empirical formula at the bottom of this page has been suggested by R. E. Lafferty to describe the differences in Q readings obtained on the two instruments.

$$\frac{Q160A}{Q260A} = \underbrace{\left[\frac{1}{1 + \dfrac{10^{-6}Q260A}{\omega CK}}\right]}_{\text{Term A: 50-500 kc}} \underbrace{\left[\frac{1}{1 + \dfrac{\omega CQ260}{50}}\right]}_{\text{Term B: 5-10 mc}} \left[\sqrt{1 + 4.32 \times 10^{-18}\omega^2}\right]_{\text{Term C}}$$

10-50 mc

This formula is the result of combining actual measurements taken with both average instruments in a form indicated by the known theoretical differences in measuring circuits described previously. It will be observed that the terms of the formula are grouped according to frequency domain. Term A represents the difference in input conductance between the two Q voltmeters, term B corrects for the difference of insertion resistance value, and term C compensates for the inductance of the insertion resistor in the Type 160-A Q Meter. The latter term was derived using the average value of 83 µµhy for this inductance. It will be noted that each term is equal to 1 outside the frequency range specified. Thus, from 50 to 500 kc, only term A is significant, while from 500

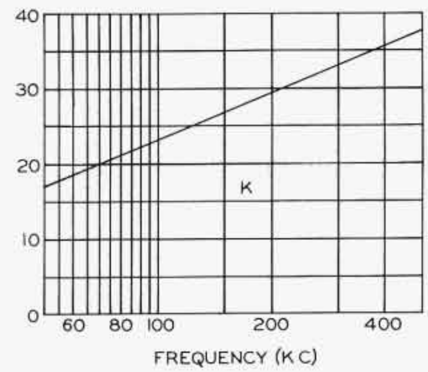

Figure 4. Values of the factor K, for use in term A of correlation equation.

kc to 5 mc, all terms are approximately 1 and the two instruments will have good agreement. From 5 to 10 mc, term B becomes significant, while terms A and C may be disregarded. Above 10 mc, B and C must be used. In general, the need for correction is proportional to the magnitude.

It should be noted that this correlation equation does not account for the effects of the harmonic content of the oscillator output of early Q Meters Type 160-A at the upper ends of the low frequency bands. Q

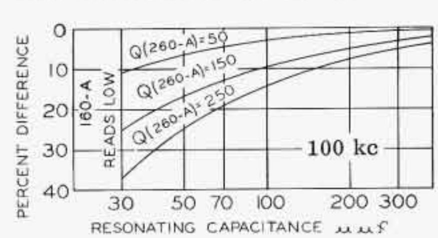

Figure 5. Calculated percent differences in Q readings between 260-A (used as base) and 160-A, at three values of Q as read on the 260-A.

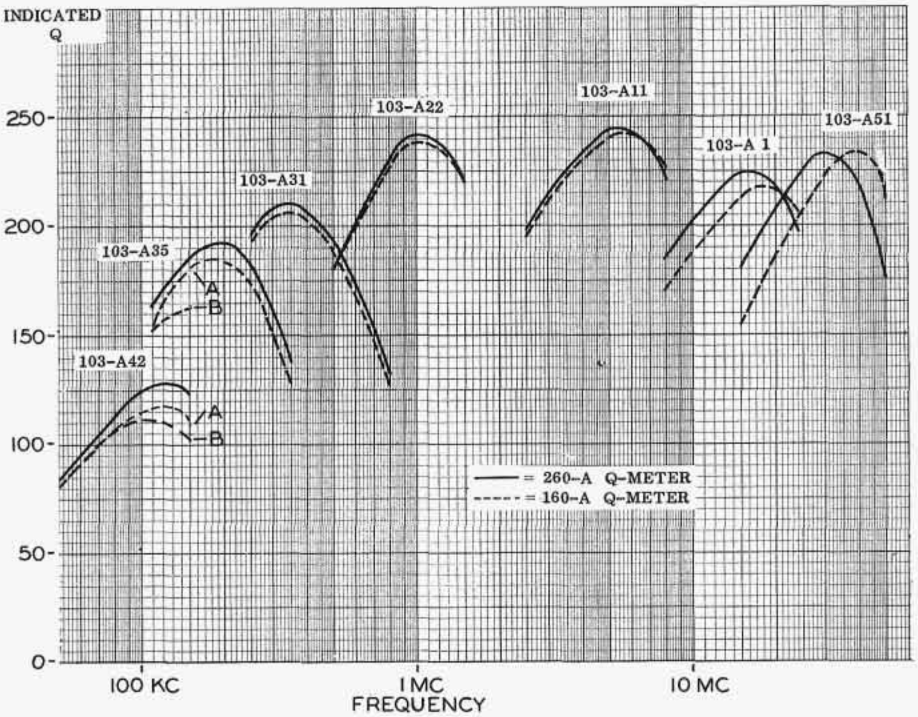

Figure 6. A comparison of Q readings obtained from duplicate measurements on both Q meters of a set of 103-A inductors. Dotted curves marked "B" indicate values obtained on older 160-A's having appreciable oscillator distortion in this range; curves marked "A" indicate 160-A's with reduced distortion.

Meters Type 160-A having serial numbers above 4581 have oscillators with reduced harmonic content.

Figure 5 illustrates the percent by which the older Q Meter will read low at 100 kc, for three values of Q. These curves were calculated by using term A of the correlation equation.

SUMMARY

The difference observed between the readings obtained on a Q Meter Type 160-A and a Q Meter Type 260-A for a given component is caused by the reduction of residual circuit parameters in the latter instrument, permitting more accurate measurement. The chart below is a summary of the causes and effects of these Q reading differences as they apply for each frequency range.

The correction equation and other data presented in this article are the result of information obtained from average Q Meter readings, and do not allow for variations within specified tolerances. Application in individual cases may require study of the parameters of the instruments concerned.

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1. "The Nature of Q", W. Cullen Moore, BRC Notebook, No. 1.
2. "Alternating Current Measuring Instruments as Discriminators Against Harmonics", Irving Wolff Proc. IRE, April, 1931.
3. "RF Microvoltages", Myron C. Selby, National Bureau of Standards.

Range	160-A reads	Cause	Term A	Term B	Term C
50-500 kc	Low	160-A voltmeter loads measuring circuit at low resonating capacitance and frequency. Early models read 3% lower at 100 kc and 8% lower at 150 kc because of oscillator harmonics. This effect is not included in Term A.	$\frac{1}{1 + \frac{10^{-6}Q}{\omega CK}}$	1	1
500kc - 5 mc	Substantially in agreement with 260-A		1	1	1
5 mc - 10 mc	Low	Effect of higher 160-A injection resistance becomes appreciable with increased frequency, Q, and resonating capacitance.	1	$\frac{1}{1 + \frac{\omega CQ}{50}}$	
10 mc - 50 mc	Low at high C High at low C	Injection resistance effect continues, causing low readings with high resonating capacitance. Reactance of injection resistor increases injected voltage with frequency, causing high readings at low resonating capacitance and high frequency.	1	$\frac{1}{1 + \frac{\omega CQ}{50}}$	$\sqrt{1 + 4.32 \times 10^{-18}\omega^2}$

Service Note

RX METER BRIDGE TRIMMER ADJUSTMENT

At frequencies above 100 MC the zero balance of the RX Meter bridge circuit is necessarily sensitive to extremely small variations in internal circuit capacitance. It is possible that minute shifts in the relative position of circuit components, caused by excessively rough handling in shipping, etc., may alter the effective capacity enough to make it impossible to obtain a null indication on the highest frequency range by adjusting ZERO BALANCE controls, "R" and "C".

In most cases, this situation can be corrected by the following screwdriver adjustment:

1. Allow the instrument to warm up, set the oscillator frequency at 200 mc, and adjust the detector tuning control as described in the Instruction Manual, with the C_p dial at 0 and R_p at ∞.

2. With a screwdriver, pry up the small metal cap located near the rear of the ground plate on top of the instrument. This provides access to a small trimmer capacitor having a vertical, slotted adjusting shaft.

3. Rotate the "R" knob and note whether the null indicator reading decreases with (a) clockwise, or (b) counter-clockwise, rotation.

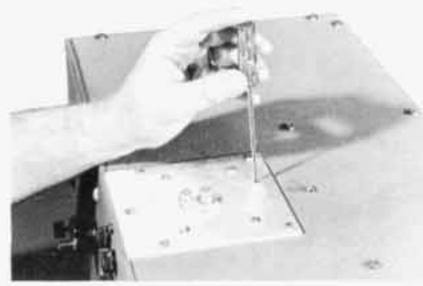

Figure 7. Adjusting RX Meter trimmer.

4. Using the screwdriver, rotate the trimmer shaft about 1/8 turn— clockwise in case (a) above, counterclockwise in case (b) above. Then remove the screwdriver * and try to obtain balance with the "R" and "C" knobs. If a null indication still cannot be obtained, rotate the trimmer another 1/8 turn in the same direction. Continue this procedure until balance can be obtained.

5. Check the balance at a frequency of 250 mc and repeat the above adjustment if necessary.

*Correct null indication can not be obtained while the screwdriver (or aligning tool) is near or in contact with the trimmer shaft.

BRC SALES ORGANIZATION

Frank G. Marble, *Sales Manager*

As is the case with most industrial firms, Boonton Radio Corporation makes most of its contacts with the outside world through its Sales Department. This, of course, is altogether reasonable, since the basic function of a Sales Department is actually that of communication-- the origination and transmission of data to customers, potential customers, and those seeking specific technical information.

BRC SALES OFFICES

Since many readers of THE NOTEBOOK will, at some time, fall in at least one of these catagories, we feel that it may be of interest to describe, very briefly, the set-up of our technical sales organization as it will affect them.

Our sales organization may be considered as having two major subdivisions; the internal Sales Department, and a nation-wide network of Sales Representatives. The former is an integral part of the company, and is responsible for a number of widely diversified functions. These include the routine work of processing orders, answering inquiries, expediting shipments, preparing quotations, etc., as well as supervising advertising and sales promotion activities and catalog and instruction material. The Sales Department also has the responsibility of developing and distributing new application information needed by customers, and of keeping abreast of the trends and requirements of the industry. In addition, company Sales Engineers handle, directly, instrument sales in the New York City, New Jersey area as far South as Washington, DC.

The second subdivision of our organization is comprised of Representatives who are located throughout the country in those areas where the electronics industry is most heavily concentrated. These organizations maintain staffs of experienced sales engineers who are throughly familiar with the operation and application of BRC instruments. Each Representative is the exclusive BRC agent in his territory, and is qualified to supply complete information on our full line of equipment.

A NOTE FROM THE EDITOR

We were extremely pleased and, to tell the truth, a little amazed at the response which greeted the first edition of THE NOTEBOOK. The number of returns received has forced us to revise completely our original estimates of the size of future printings. A few figures might be interesting. We mailed roughly 40,000 copies of the first edition to a selected group of engineers, scientists, and educators. As a direct result of the reply cards returned, we are mailing this second edition to a list of 19,000 readers, and it is still growing daily. In addition, we have received several hundred notes and letters expressing interest and encouragement.

Such a vote of confidence is deeply appreciated. We will do our best to merit it in the future by maintaining, as nearly as possible, the level established in our first issue.

CORRECTION: Somewhere along the way, in preparing NOTEBOOK No. 1, an exponent was lost. The first equation in column 1, page 6, should read:

$$L_e = \frac{L}{1 - \omega^2 L C_d'}$$

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Printed in U.S.A.