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TO

GENERAL RADIO

EXPERIMENTER

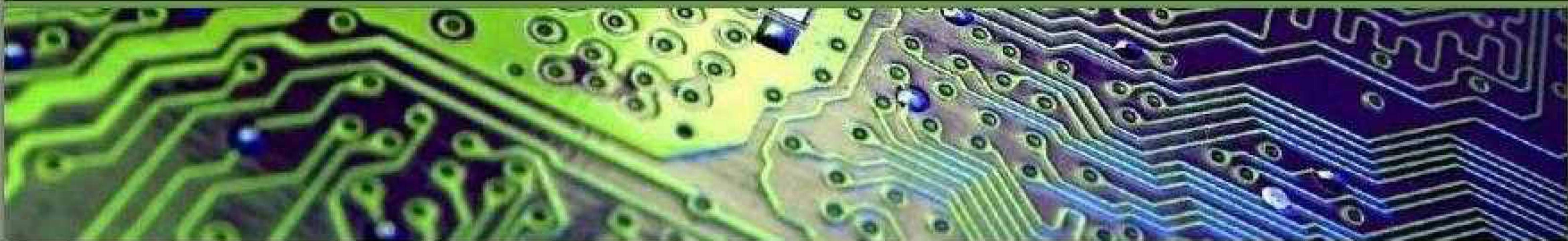
VOLUMES XVIII AND XIX

June, 1943 to May, 1945

GENERAL RADIO COMPANY

CAMBRIDGE **MASSACHUSETTS**

U. S. A.



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THE COMMUNICATIONS RECEIVER AS A WAVE ANALYZER

● A GOOD MANY laboratory measurements would be considerably simplified if there were available a device to do at radio frequencies the job that the familiar

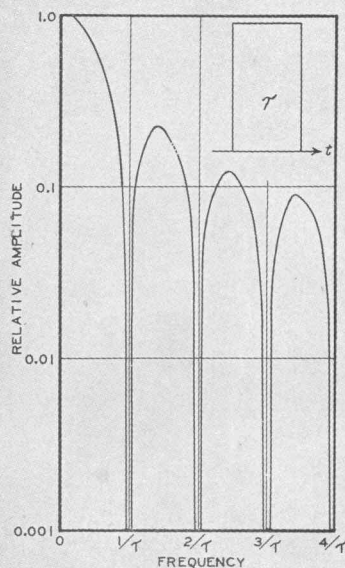
heterodyne type of wave analyzer does at audio frequencies.

Nearly every laboratory, however, has a communications type of radio receiver which can be quite easily adapted to the problem. While the input impedance may not be so high, and the output meter and the frequency dial calibration may not be so accurate as may be desired in many cases, such inadequacies can often be remedied by proper use and calibration. Standard-signal generators and frequency meters are readily used for this calibration.

The simplicity with which an elementary type of analysis can be made with a communications receiver is illustrated by its use for determining the length of short, periodically-repeated pulses of voltage. Harmonic generators¹ and Class C amplifiers often operate with such pulses of plate current. The adjustment for proper operation in these cases is usually well indicated by setting for the desired length of current flow, and a simple technique for the measurement of this length provides an additional tool in this adjustment.

¹F. R. Stansel, "Some Analyses of Wave Shapes Used in Harmonic Producers," *Bell System Technical Journal*, v. 20, n. 3, July, 1941, pp. 331-339.

FIGURE 1. Envelope of the frequency spectrum of a rectangular pulse.



The waveform that is generally obtained in a system of this type can be well represented by a short pulse that is a section of a sine wave. The angle of current flow, θ , is the usual measure of the length of such a pulse and is related to its length in seconds, τ , by the expression

$$\theta = 2\pi f\tau$$

where f is the frequency of the fundamental driving voltage.

In many cases the pulse is so flat on the top that it is better represented by a trapezoid with sides of finite slope. As a limiting case it can be considered rectangular, and, because the analysis and measurement for the rectangular shape is simple, it will be considered first.

The length of a rectangular pulse can be determined from the locations of the minima of intensity of the components in its frequency spectrum. For a truly rectangular pulse, the minima (zeros of the envelope) occur at integral multiples of the frequency whose period is equal to the length of the pulse. Thus a 10-microsecond pulse has a spectrum with minima of intensity at 100 kc, 200 kc, 300 kc, etc.; and a 1-micro second pulse has minima at 1 Mc, 2 Mc, 3 Mc, etc.

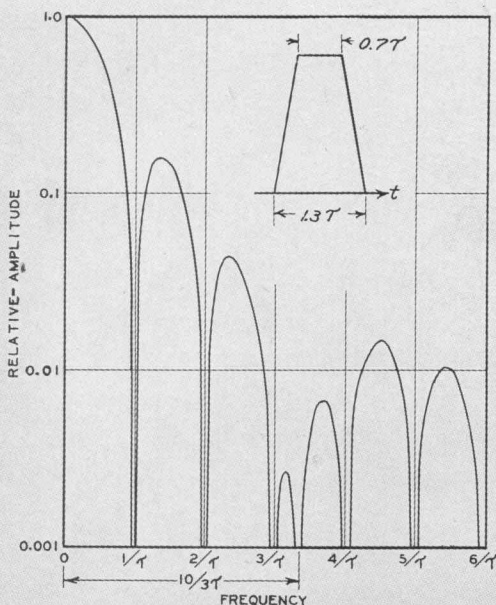
The Fourier analysis of a rectangular pulse² yields a frequency spectrum of the form shown in Figure 1. Here only the envelope of the amplitude of the components is plotted. A pulse that is periodically repeated has a line spectrum with discrete components at integral multiples of the frequency of repetition, and the actual minima will in general not occur at exactly the above values. Thus, if the repetition rate is 1100 cps for the 10-microsecond pulse, components can be found at 1100, 2200, 3300, etc., cps. The amplitude of these components will decrease gradually with increasing frequency until the first minimum at 100,100 cps, beyond which the amplitude increases to a secondary maximum, decreasing again to the second minimum at 200,200 cps. Succeeding minima occur at 300,300; 400,400; 500,500; 599,500; 699,600; etc., cps.

If a rectangular pulse of voltage is impressed on a communications receiver, the general nature of the spectrum can readily be observed on the output meter as the receiver is tuned over its range. Then the frequencies at which minima occur can be readily located. The simple relationship between the location of the minima of the frequency spectrum and the length of the pulse then makes possible the calculation of the length of the pulse.

Because of the relatively broad band of reception (except with the use of quartz-crystal filters) at audio-frequency repetition rates, the response may appear to follow the envelope of the spectrum instead of giving the pattern of discrete responses that one might at first

²The rectangular pulse is analyzed in many texts. The article by Stansel (see Footnote 1) considers all the wave shapes treated here.

FIGURE 2. Envelope of the frequency spectrum of a trapezoidal pulse. It is very similar to the envelope of Figure 1 but has a double set of zeros.



expect. This envelope response is generally advantageous for locating the minima, in that a smooth drop to a minimum as the receiver tuning is varied makes possible rapid location of each one. In addition, since the sensitivity of the receiver is not perfectly uniform over its range, the frequency spectrum may appear to be somewhat different from the expected, but usually the distortion is not troublesome.

One precaution worth observing is that the receiver should not be overloaded. If overloading occurs, the frequency spectrum may be altered sufficiently to make the measurements useless. However, except for very low repetition rates and impulses of extremely short duration, one can usually obtain adequate measurements of the frequencies of the minima with relatively low inputs. Extraneous signals may cause spurious responses to be obtained, and the apparatus therefore should be arranged to avoid pickup of unwanted signals.

Although the foregoing has been based on a pulse of truly rectangular shape, the procedure can also be used for pulses that are only approximately rectangular. In actual practice the sides of a pulse have a finite slope, and a trapezoidal pulse (actually of an isosceles-trapezoid shape) is a better approximation to the usual case.

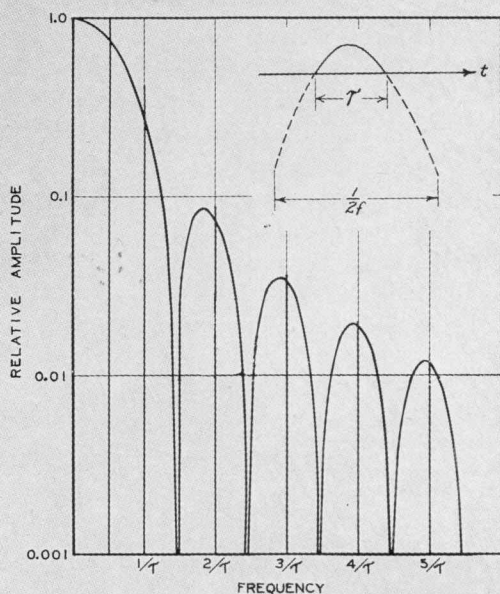
The envelope of the frequency spectrum of a pulse whose base is 86% longer

than its top is shown in Figure 2. Two sets of minima occur. Those in the first set are integral multiples of the fundamental frequency whose period is the average length of the pulse. This relationship is true in general for trapezoidal pulses, and it is normally the only one of interest.

The minima in the second set occur at integral multiples of the frequency whose period is one-half the difference in length of the base and the top. This latter difference is often small, with the result that only a few such minima occur for many minima of the set corresponding to the average length of the pulse. For the extreme case of the triangular wave shape the two sets are identical, and the average length is the only one obtained.

The pulse that is a section of a sinusoid has a series of zeros that are not so simply related to the length of the pulse. If the section is a complete half wave, the zeros occur at frequencies that are $3/2$, $5/2$, $7/2$, etc., multiples of the frequency whose period is equal to the length of the pulse. This case is the familiar one of the half-wave rectifier whose output has no components of the

FIGURE 3. Envelope of the frequency spectrum of a pulse that is a very small section of a sine wave. When the pulse is a half sine wave repeated at the sine-wave fundamental frequency, the relative amplitudes of the components of the frequency spectrum are as shown by the vertical lines in the plot. The amplitudes for intermediate pulse lengths will lie between the limiting values shown here.



3rd, 5th, 7th, etc., harmonic of the fundamental frequency.

As the pulse becomes a smaller section of a sine wave, these minima shift downward in frequency, reaching in the limit 1.430, 2.459, 3.471, 4.477, 5.482, 6.484, etc., times the frequency whose period is equal to the length of the pulse. The shift from the simpler values given above is small, becoming smaller with each succeeding minimum. The envelope for a very short pulse is shown in Figure 3.

It is interesting to notice that the av-

erage length, considered as the area divided by the maximum height, for a half sinusoid, is $2\tau/\pi$ and that for a section of a sine wave which represents only a small angle of flow is $2\tau/3$. Thus, the first minimum is approximately at the frequency whose period is equal to the average length of the pulse. However, each succeeding minimum is spaced from the one preceding it by approximately the frequency whose period is equal to the total length, τ .

— ARNOLD PETERSON

CONTINUOUS INTERPOLATION METHODS

● **THE FREQUENCY CALIBRATION METHODS** which have been outlined previously¹ are equivalent to point-by-point interpolation, where the frequency of an oscillator could be set to, or measured against, the nearest multiple of a low audio-frequency standard. We now consider methods for setting an oscillator to, or measuring the frequency of an oscillator at, any point in the frequency range.

The basis of continuous interpolation is quite generally a continuously variable oscillator, usually in the audio-frequency or low radio-frequency range. Such an oscillator is designed to be as stable as possible and to permit calibration on a greatly expanded scale. In general, the over-all accuracy of measurement depends upon the absolute accuracy (in cycles per second) of the interpolation oscillator, and for greatest absolute accuracy this requires that the oscillator be in the low-frequency ranges.

The problem is indicated in the very simple diagram of Figure 1. The fre-

quency to be measured, or the desired final value of frequency, is shown at f_x , lying anywhere between two known multiples of a standard frequency f_s , nf_s , and $(n+1)f_s$. Since all interpolation systems are based on the interpretation of this elementary diagram, it is worth considering in greater detail.

Firstly, assuming nf_s and $(n+1)f_s$ to exist as physical frequencies, that is, that voltages of these frequencies exist in usable magnitude, f_x can be evaluated in several ways, as follows:

(1) An auxiliary radio-frequency oscillator is used, generally called a "Heterodyne Frequency Meter." The dial settings of this oscillator are determined for nf_s , f_x , and $(n+1)f_s$ in turn;

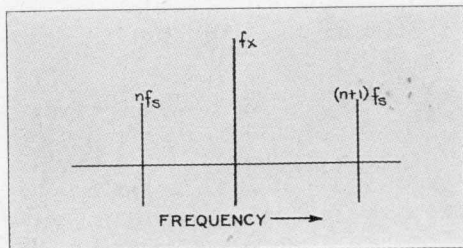


FIGURE 1. Basic Interpolation Diagram.

¹J. K. Clapp, "Calibration of Equipment in the Low and Medium Radio-Frequency Ranges, in Small Steps of Frequency," General Radio Experimenter, Vol. XVIII, No. 3, October, 1943.



f_x is then found by linear interpolation, as will be shown later.

(2) The frequencies f_x and nf_s , or f_x and $(n + 1)f_s$, are impressed on a detector (or receiver) and the *lower* beat frequency difference produced in the detector is obtained. This beat frequency ranges from 0 to $f_s/2$ cycles. The magnitude of the frequency difference is measured, and its sign determined, by methods considered below.

(3) The frequencies f_x and nf_s , or f_x and $(n + 1)f_s$, are impressed on a detector, and the *higher* beat frequency difference produced in the detector is obtained. This beat frequency ranges from $f_s/2$ to f_s cycles. The magnitude of this difference is measured, and its sign determined, by methods described later.

(4) If the standard frequency can be *raised*, by a smoothly acting fine frequency control, the harmonic nf_s can be slid along the frequency scale until it is brought exactly to the value f_x . The amount that nf_s is raised to accomplish this can be determined from the frequency control settings; the sign of the change is always positive.

Secondly, *assuming that nf_s and $(n + 1)f_s$ do not exist physically*, f_x can still be evaluated in terms of these multiples of the standard frequency f_s , as follows:

(5) The *difference* between f_x and the known frequency of an interpolation oscillator, f_i , is made equal to a multiple of the standard frequency f_s , using cathode-ray oscillograph patterns. The important point to notice is that this method permits measurements to be made at high multiples of a standard frequency, where the standard source has *no harmonics*. The multiples exist only as patterns on the oscilloscope screen.

Methods (1) and (2) are considered in the following paragraphs, where factors affecting the operation are detailed. Methods (3) and (4) will be discussed in a forthcoming issue of the *Experimenter*.

The methods are summarized in the accompanying diagrams, where the essential frequency relationships are illustrated at the top of the figure, followed by a block diagram below. The equations pertaining to each method are followed by a brief listing of the principal limitations and advantages.

Method 1: Direct Interpolation

This method would not be used where precise results are required, principally because of the basic fault that interpolation is accomplished by using only a very small portion of the working scale. To obtain fine interpolation, then, it is necessary to "magnify" this small portion of the scale which, in this case, can only be accomplished by mechanical means. These are subject to severe limitations, as the list indicates.

The method is still of value for rough work, or where measurements of moderate accuracy need to be made only occasionally.

Method 2: Direct Beating

This method is widely used and is capable of high precision. It has certain limitations, particularly as to the measurement of very high frequencies. Because of its general utility, however, the details below may be helpful.

If the frequency f_x is that of a remote transmitter, or one of several local transmitters or oscillators, radio frequency selectivity, as provided by the receiver in the diagram, is a practical necessity.

Where interference or noise conditions permit, the nuisance of having to tune the receiver to each value of f_x can be avoided by using an untuned detector for the receiver. Generally this results in some decrease in the upper limit of range for f_x because of the decreased sensitivity.

A compromise arrangement, which is useful if the unknown frequency is intermittent (as, for instance, that of a keyed transmitter), is to use an untuned detector with a local substitution oscillator. This oscillator is set to zero beat, or matched, with the unknown frequency. The fundamental frequency of the oscillator is then measured against the standard.

By using harmonics of the substitution oscillator to match the unknown frequency, the effective upper limit of measurement can be greatly increased. By choosing particular harmonics, which can be done easily whenever an approximate value for f_x is known, simple or decimal multipliers can be maintained for all measurements.

For example, when 10 kc standard frequency harmonics are used it is difficult to make measurements above 8 to 10 Mc. When a substitution oscillator is used, if the 10th harmonic is used for matching to f_x , then the final result is always 10 times the measured oscillator fundamental frequency, requiring no computation.

In common with other methods of measuring frequencies, some difficulty is encountered in measuring frequencies which are only a few cycles removed from a standard harmonic frequency, or which are very nearly halfway between two standard harmonic frequencies. The first

can be avoided by substituting a new value of standard frequency through the use of a 9 and 11 kc multivibrator, instead of 10 kc, which effectively transfers the small frequency difference to appear as that same difference from a multiple of 1 kc, where it is easily measured. This holds for all frequencies except multiples of 990 kc ($9 \times 10 \times 11$). A similar substitution can be made by using an off-set standard². A third possibility is to use an oscillating receiver, tuned so that a beat tone of about 1 kc is obtained against the standard harmonic frequency. The beat tone obtained against the unknown frequency will be only a few cycles different. The combination acts as a tone of about 1 kc, waxing and waning at a rate equal to the difference of the unknown and standard frequencies. The "flutter" can be matched with an interpolation oscillator using a cathode-ray oscilloscope³.

When the frequency to be measured lies nearly halfway between two standard frequency harmonics, the receiver selectivity can be used to favor the beat obtained with one of the two standard frequencies, at the expense of the other. In cases where the selectivity is not adequate, or where an untuned detector is used, the standard frequency can be doubled. In effect, this eliminates all odd-numbered harmonics from the original harmonic series, and only a single beat frequency difference remains to be measured.

In attempting to extend the measuring range toward higher frequencies, several problems arise. First, with a given standard frequency, say 10 kc, measurements at higher frequencies require correspondingly higher multiples

²J. K. Clapp, "Identification of Harmonics in a Harmonic Series," *General Radio Experimenter*, Vol. XVIII, No. 4, September, 1943.

³J. K. Clapp, "Using the Cathode-Ray Oscillograph in Frequency Comparisons," *General Radio Experimenter*, Vol. XVI, No. 7, December, 1941.



of the standard frequency. It is very difficult to get harmonics of usable amplitude much beyond the 1000th harmonic. Some improvement can be had by modulating a higher frequency multivibrator with the output of a lower frequency one. For example, substantially stronger 10 kc harmonics can be obtained by using a 50 kc multivibrator modulated by a 10 kc multivibrator. A control for the amount of modulation must be provided since the amplitudes of different multiples of 10 kc vary substantially with different amounts of modulation.

An accompanying difficulty in this case centers on identifying the standard frequencies. This becomes more and more difficult as the number of used harmonics is increased. Where the modulated multivibrator system is used, the

modulation can be reduced to zero and the nearest harmonic of the high frequency multivibrator can be identified. On reintroducing the modulation, the number of low-frequency harmonics passed over to reach the used harmonic can be counted.

If a higher standard frequency is used, to avoid the use of high multiples, other problems arise. Firstly, the pass-band of the receiver must be made wider, the higher the standard frequency. Secondly, the interpolating equipment must have a corresponding increase in range, usually accompanied by a corresponding loss in absolute accuracy. Since the interpolating equipment must be capable of measuring from 0 to $f_s/2$ cycles, any substantial increase in f_s means a great extension in the range of the interpolating equipment.

SUMMARY

Method 1 : Direct Interpolation

Using an oscillator (Heterodyne Frequency Meter), and assuming the calibration to be straight-line-frequency over the used portion, set for zero beat successively with the three frequencies nf_s , f_x , and $(n+1)f_s$, calling the respective dial readings θ_n , θ_x , θ_{n+1} . Then, from the figure

$$\frac{f_x - nf_s}{(n+1)f_s - nf_s} = \frac{\theta_x - \theta_n}{\theta_{n+1} - \theta_n} = \frac{f_x - nf_s}{f_s}$$

From which,

$$f_x = nf_s + \frac{\theta_x - \theta_n}{\theta_{n+1} - \theta_n} f_s$$

when the measurement is referred to the lower standard frequency harmonic, nf_s . Also,

$$\frac{(n+1)f_s - f_x}{(n+1)f_s - nf_s} = \frac{\theta_{n+1} - \theta_x}{\theta_{n+1} - \theta_n} = \frac{(n+1)f_s - f_x}{f_s}$$

From which

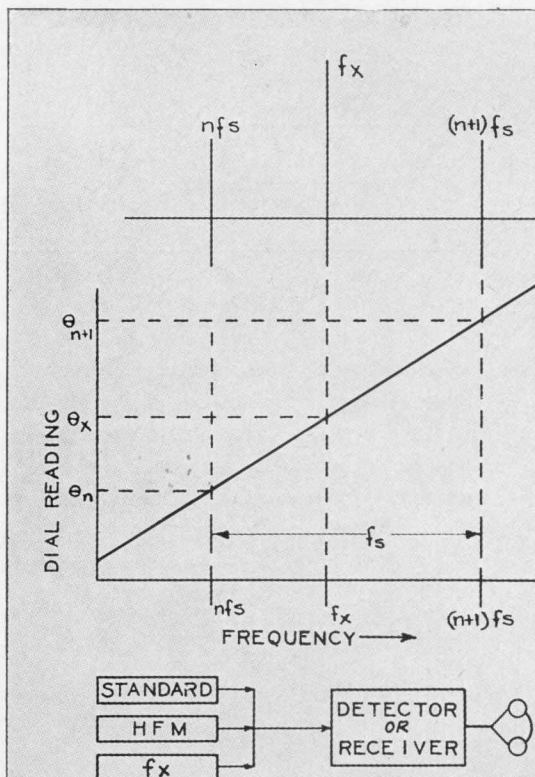
$$f_x = (n+1)f_s - \frac{\theta_{n+1} - \theta_x}{\theta_{n+1} - \theta_n} f_s$$

when the measurement is referred to the higher standard frequency harmonic, $(n+1)f_s$.

Principal Limitations of this method are:

(a) Linearity of calibration curve of Heterodyne Frequency Meter.

FIGURE 2. Diagrams illustrating method and equipment for direct interpolation.



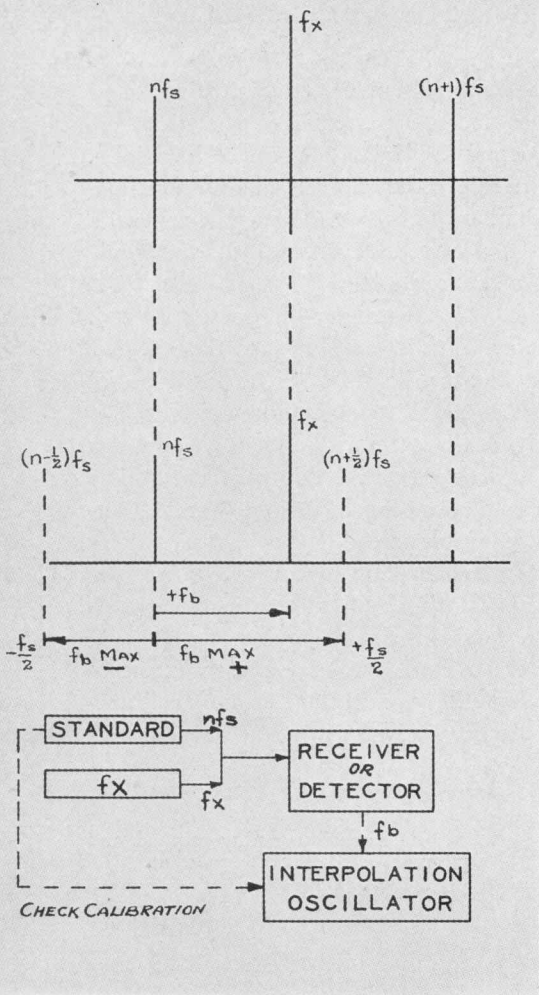


FIGURE 3. Diagrams illustrating method and equipment for direct beating.

- (b) Fineness of graduation of the scale.
- (c) Mechanical errors of drive, such as eccentricity, backlash, worm corrections, etc.
- (d) Accuracy of setting for zero beat.
- (e) Accuracy of calculations, or of reading charts, for solving equation.
- (f) Use of only a very small portion of the scale for making interpolations.

Principal Advantages are:

- (a) Simplicity of equipment.
- (b) Simplicity of operation.
- (c) Values of $n f_s$ and $(n + 1) f_s$ are known at once from heterodyne frequency meter calibration.

Method 2 : Direct Beating

The standard frequency $n f_s$ and the unknown frequency f_x are impressed on the receiver, which is tuned to accept both. The lower beat frequency difference $+f_b = f_x - n f_s$ (or $-f_b = n f_s - f_x$) is obtained in the receiver output and is measured by matching with the interpolation oscillator, covering a range from 0 to $f_s/2$ cycles.

Then $f_x = n f_s \pm f_b$.

Principal Limitations of this method are:

- (a) Standard frequency harmonics must be of usable magnitude in region around f_x .
- (b) The frequency of the used standard frequency harmonic, $n f_s$, must be determined.
- (c) The sign of the beat frequency, f_b , must be determined.
- (d) The over-all accuracy depends on the accuracy of the interpolation oscillator and how carefully it is matched to f_b .

Principal Advantages are:

- (a) Relatively high absolute accuracy is possible, since the entire range of the interpolating equipment is devoted to measuring f_b which is only a small part of the final result.
- (b) The absolute accuracy is the accuracy in cycles with which f_b can be measured and is constant over a very wide frequency range.

— J. K. CLAPP

(To be continued)

ERRATA

Readers who were mystified by the relation between Equations (11) and (10) on page 1 of the December *Experimenter* will be relieved to learn that the exponent ($\frac{1}{2}$) was omitted from Equation (10), which should read

$$R_{in} \approx R_0 \left[1 - \left(\frac{X_0 + X_T}{R_{in}} \right)^2 \right]^{1/2} \quad (10)$$

In addition, a factor of $\frac{1}{2}$ was omitted in the numerical expression for R_0 on page 2 which should read

$$R_0 = 71.2 \left[1 + \frac{1}{2} \left(\frac{2.1 + 10.3}{71.2} \right)^2 \right]$$

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THE

General Radio EXPERIMENTER

VOLUME XVIII No. 9

FEBRUARY, 1944

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

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MOTOR SPEED CONTROL WITH THE VARIAC

•AN IMPORTANT USE of the Variac is the control of speed on fractional-horsepower a-c motors. Successful use of the Variac in this application, however,

depends upon both the type of motor and the type of load, and it is important that this be kept in mind when considering the Variac as a speed control.

Repulsion and Series Motors

Repulsion and series motors are by far the easiest to control through voltage variation. The speed of each of these types is sensitive to voltage, and control can be obtained over a wide range of speed for practically any type of load. The curves of Figures 1 and 2 show typical speed-voltage characteristics for these types of motors with a belt-driven load. The repulsion type used was a $\frac{1}{4}$ hp and the series type a $\frac{1}{15}$ hp. The characteristics of these two motors are somewhat similar to those of the d-c series motor.

Induction Motors

In general, speed of an induction motor cannot be controlled by voltage. The motor tends to run at a speed approaching the synchronous value determined by the number of poles on the stator winding. Speed can be varied only by changing slip, and, for belt drives and other fairly constant loads, a reduction of voltage changes the slip only slightly until the breakdown point is reached and the motor stalls.

One outstanding exception, however, is found for fan loads, where the effective load varies greatly with the speed. Some types of induction motors can be made to operate satisfactorily for this service with Variac control, and stable operation can be had on a wide range of speeds. This is discussed below.

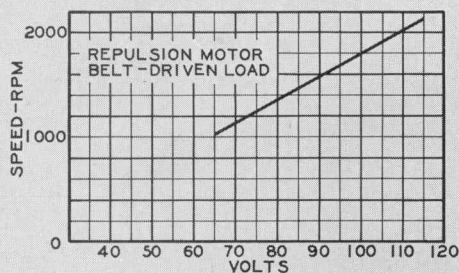


FIGURE 1.

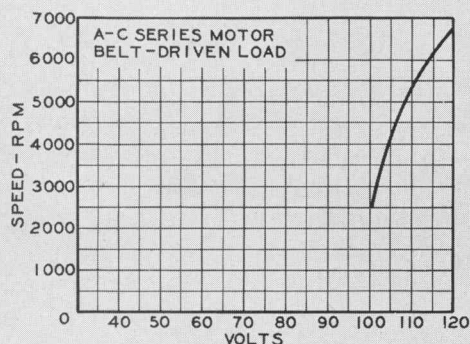


FIGURE 2.

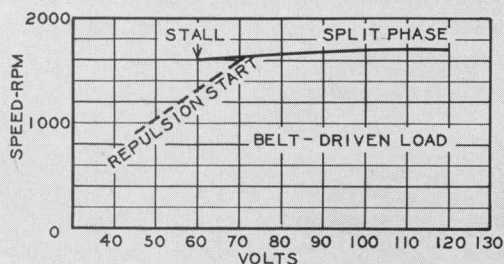
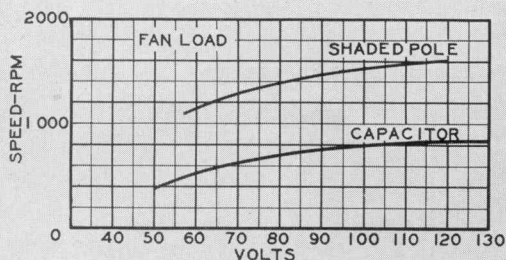


FIGURE 3 (above); FIGURE 4 (below).



Split Phase

The split phase motor with automatic cut-out, in particular, is a type that is not adapted to speed control by voltage variation.

This motor has an auxiliary winding spaced 90 electrical degrees from the main winding. Ordinarily, this winding is used only for starting and is disconnected by an automatic switch when running speed is attained. After the auxiliary winding is disconnected, the speed tends to approach synchronous speed, the slip depending mainly upon the load. Voltage variation has little effect upon speed. The speed of a split phase motor is usually changed by switching taps on the stator winding, which effectively changes the number of poles, and also provides two or more values of synchronous speed. The speed-voltage characteristic of a $\frac{1}{3}$ hp split phase motor controlled by a Variac is shown in Figure 3.

Repulsion-Start Induction Motors

The repulsion-start induction motor is provided with an inducing winding, commutator, and brushes, by means of which the rotor is brought up to running speed by operation as a repulsion motor. The commutator bars are then short-circuited by an automatic switch, and the motor runs as a single-phase induction motor. While the motor can be Variac-controlled during the starting period when it is running as a repulsion motor, or whenever the voltage is reduced to the point where the repulsion-start system cuts in, the range of speed variation is necessarily low, and, if the repulsion winding is not designed for continuous service, excessive heating may occur. Figure 3 shows a voltage-speed curve for a $\frac{1}{3}$ hp repulsion-start type of motor.



Capacitor and Capacitor-Start Motors

Split phase operation is often obtained by using a capacitor in series with the auxiliary winding. If the auxiliary winding is opened by an automatic switch when running speed is reached, the motor is called a capacitor-start motor. If the auxiliary winding is connected permanently, the motor is called a capacitor motor.

The capacitor-start type cannot be controlled by a Variac. The capacitor-type can be, particularly if the load is a fan or blower.

Speed Control for Fans and Blowers

The torque required to drive a fan or blower is a function of speed, and, consequently, stable operation can be had with induction motors operating at high values of slip. Capacitor motors and shaded pole motors are often used for this service.

With the shaded-pole motor, speed control over a considerable range can be obtained, as shown in Figure 4. For low speeds, however, it is often necessary to start at a fairly high voltage and then reduce the voltage to a lower value for running.

The capacitor motor can be controlled by varying the voltage on both the main and auxiliary windings simultaneously. Some capacitor motors, however, are so designed that a fixed voltage can be applied to the auxiliary winding and a variable voltage to the main winding. Figure 4 shows the speed-voltage characteristic obtained with a typical capacitor motor driving a squirrel-cage fan when the voltage on both windings is varied.

A split-phase motor can be controlled in the same way, if it is designed for operation with the auxiliary winding permanently connected. While a few of these are used for fan service, they are not as common as the shaded-pole and capacitor types.

Within each basic type of motor, the designer can vary many factors to produce a motor suited to any particular job. To cover the complete range of possibilities is considerably beyond the scope of this article. With any given type of motor the speed-voltage characteristic is, of course, dependent upon the torque-speed characteristic of the load as well as that of the motor itself. The speed of some types of motors is voltage-sensitive, however, and that of others is not, and this should be considered first when the use of a Variac for motor speed control is contemplated.

CONTINUOUS INTERPOLATION METHODS

PART II

● **LAST MONTH'S ARTICLE** outlined the various methods of continuous interpolation. Method 1, Direct Interpolation, and Method 2, Direct Beating, were considered in detail. This portion of the article discusses Methods 3, 4, and 5.

Method 3: Direct Beating

This method, utilizing the higher of the two beat frequencies, $f_x - nf_s$ or $(n + 1)f_s - f_x$, is adaptable to high frequency measurements because of the limitation of the interpolator range to $f_s/2$ to f_s . For example, with a standard

frequency of 1 Mc, the lower beat frequency (Method 1) ranges from 0 to 0.5 Mc; the higher (Method 2) ranges from 0.5 to 1.0 Mc.⁴

For frequencies of these magnitudes, a broadcast receiver can be used for an interpolator, readings being taken from the receiver calibration. For improved accuracy, the beat frequency can be compared with a lower standard frequency, the difference frequency in the receiver output being measured by an interpolation oscillator (as in Method 1).

In common with other methods, difficulties are encountered when f_x lies near a standard frequency, giving a difference frequency to the next higher or next lower harmonic which is near to the value of the standard frequency. Thus two frequencies are impressed on the receiver instead of one.

Method 4. Sliding Harmonics

To reduce the difficulties enumerated above and to simplify operation, this system is designed so that no beat difference frequency is used and the only beat utilized is solely for matching, — zero beat. The interpolator control operates on all frequencies simultaneously, so no selection of "lower" or "higher" beat frequency differences and no measurement of any beat frequency difference is required.

The system illustrated can be set up as follows: The standard consists of a 950 kc crystal oscillator, in terms of which all measurements are made and against which the interpolator can be calibrated or checked. The interpolator consists of a 50 to 60 kc variable frequency oscillator, straight line, direct reading calibration. The scale has 1000 divisions, giving 10 cycles per division.

The standard and interpolator outputs are impressed on a modulator and the upper side frequency is selected by a filter in the modulator output. This filtered output is of 1000 kc, when the interpolator is set at zero on the dial (50 kc) and can be varied up to 1010 kc at 1000 on the dial (60 kc), a total range of one percent. Since there are 1000 divisions, each division is 0.001 percent or 10 parts per million.

This output frequency is used to control one or more multivibrators. A one-megacycle multivibrator, with a special output amplifier, provides harmonics of usable magnitude up to 200 Mc. This harmonic output is utilized in connection with a heterodyne frequency meter having a range of 100 to 200 Mc, calibrated at each megacycle.

The system is particularly intended, as described, for the measurement of frequencies in the range from 100 to 1000 Mc (approximately). Through the addition of a second multivibrator, 0.1 Mc and the use of a Heterodyne Frequency Meter covering 10–20 Mc, measurements in the range from 10 to 100 Mc could be made.

Now, with the interpolator dial at zero, the output frequency ($f_s + f_i$) is exactly 1 Mc and the harmonic frequencies of the multivibrator are all multiples of 1 Mc. Thus the heterodyne frequency meter can be checked at every point on its scale. If the interpolator dial is set at 1000, each harmonic frequency is increased by 1 percent. At the 100th harmonic this means that the frequency has been raised from 100 Mc to just 101 Mc. Also, complete coverage over this interval has been obtained. For any higher harmonic, the 1-percent range exceeds the difference between the given harmonic and the next higher one. Consequently complete coverage of the entire range from 100 to 200 Mc is assured.

⁴"An U.H.F. Measuring Assembly," S. Sabaroff, PROC. I.R.E. 27-3, p. 208, March, 1939.



To make a measurement, the heterodyne frequency meter is set to zero beat against f_x , either on the fundamental or a harmonic. With the heterodyne frequency meter at this setting, the interpolator dial is advanced from zero until the first beat is heard between a multivibrator harmonic and the frequency meter. This beat is set to zero. The unknown frequency is then

$$f_x = nf_s (1 + \Delta) = nf_s + nf_s \Delta,$$

where Δ is given directly by the interpolator dial, on pointing off 5 decimal places. The used harmonic frequency of the multivibrator is identified at once from the dial of the frequency meter.

If the frequency being measured is slightly above 162 Mc and the interpolator dial reading is 276 divisions, the frequency is $162 + 162 \times 276 \times 10^{-5} = 162 + 0.447 = 162.447$ Mc.

Since the frequency change to move any harmonic up to the position of the next higher is always just one megacycle, the number of divisions on the interpolator dial necessary to accomplish this can be determined in advance and tabulated. The fraction of a megacycle involved in measuring any frequency is then [(observed number of divisions)/(divisions for 1 Mc)] $\times$ 1 Mc.

The question can be raised as to what would happen if the heterodyne frequency meter calibration were in error by 1 Mc. The calibration can be positively checked at both ends of the scale. With the interpolator dial at zero, set the heterodyne frequency meter to the point marked 101 Mc and set to zero beat. Leaving the heterodyne-frequency meter at this point, advance the interpolator dial to 1000 divisions. At this setting zero beat should again be obtained. If no beat is heard, or if the beat is heard before the dial has been advanced the full 1000 divisions, the heterodyne frequency meter is not set to 101 Mc but to

a point above, or below, this value respectively. A similar check can be obtained at 202 Mc.

Method 5. Cathode-Ray Oscilloscope

Through the use of a cathode-ray oscilloscope, interpolations can be made over an extended frequency range, *without the use of harmonics*. The working upper limit depends principally on the ability to open out the pattern.

The basic method is to combine the unknown frequency f_x with a second known and adjustable frequency, from an interpolating oscillator, f_i , so that the difference in frequency, $f_x - f_i$, is a multiple of a third, known and fixed frequency f_s .⁵

The setup is extremely simple. The unknown, f_x , and interpolator, f_i , oscillators are effectively connected in series to the vertical deflection plates of a cathode-ray oscilloscope. The standard frequency source, f_s , is connected to the horizontal deflection plates, and should be capable of supplying a large voltage so that high-frequency patterns can be opened out for easy viewing. The pattern produced by the two voltages in series, as viewed against a linear sweep, is indicated in the sketch, for unequal and equal amplitudes. This pattern appears similar to a modulated wave, but it differs in several important respects.⁶ The important feature, for present purposes, is that the frequency of the *envelope* is that of the *difference of the two applied frequencies*.

For illustration, assume that a standard frequency of 1000 cycles and an interpolation oscillator having a range of 11,000 to 11,500 cycles, calibrated at

⁵"An Interpolation Method for Setting Laboratory Oscillators," F. R. Stansel, Bell Labs. Record, 1940, 19, p. 98 (Nov. 1940).

⁶See, for example, "Applied Electronics," E. E. Staff, M.I.T., John Wiley & Sons, p. 699 et seq.

every cycle, are available, and that a laboratory oscillator is to be set to a frequency of 323,383 cycles.

The interpolator oscillator is set to 11,383 cycles by its calibration. The oscillator to be adjusted is then set to 323,000 cycles, as read from a previous calibration. This adjustment need not be precise — it is to insure that the final result will not be in error by one, or more, multiples of 1000 cycles.

The oscillator frequency is then increased from 323,000 cycles. The envelope of the combined wave on the cathode-ray tube is that of the difference of the two oscillator frequencies, but *no pattern appears until this difference is made a multiple of 1000 cycles*. This occurs when the oscillator frequency has been increased from 323,000 cycles to 323,383 cycles. At this frequency the difference of oscillator and interpolator frequencies is $323,383 - 11,383 = 312,000$ cycles, which is an integral multiple of the standard frequency of 1000 cycles, and the envelope pattern stands still. (The individual oscillations of the "carrier" wave would not be seen because of the high frequency.)

Actually two patterns appear, the "front" and the "back" traces, as sketched. These can be separated by phase shifting networks, but this is not usually necessary.

The error, *in cycles*, in making this adjustment is numerically equal to the error in cycles in setting the interpolator frequency. This error can be made very small, so that the method is capable of high accuracy.

If the oscillator is set at some frequency, to be measured, the interpolator is varied until the pattern appears and is

made stationary. The frequency increment is then read from the interpolator scale, the appropriate multiple of the standard frequency being identified from the oscillator calibration.

In the example an interpolator frequency of 11,383 cycles was used, but patterns can be obtained with any frequency which gives a multiple of the standard frequency when subtracted from or added to the desired frequency. For example, the frequency of 323,383 cycles gives a pattern for interpolation oscillator frequencies of either 11,383 ($11,000 + 383$) or 10,617 ($11,000 - 383$). The interpolator oscillator consequently needs to have a range of only one-half the standard frequency to obtain complete coverage.

The interpolator frequency does not have to be as low as 11,000 cycles, but may be of the order of the unknown frequency. A high interpolator frequency has the advantage of lowering the frequency of the envelope and hence the pattern is easier to spread out. The absolute accuracy of a high frequency interpolator is usually less, so high accuracy demands the lower frequency.

Using an interpolator of 11,000–11,500 cycles range, measurements can be made to within a cycle to above one megacycle. To extend the frequency range, a higher frequency interpolator, say 30–35 kc, can be used with a 10 kc standard, with a somewhat larger error.

For convenience the scale of the interpolator is marked only in frequency increment, 0–500 cycles for the range 11,000 to 11,500. A second scale marked 500–1000 cycles is convenient when the required increment is negative, avoiding subtraction.

—J. K. CLAPP



Summary

Method 3 : Direct Beating

The standard frequency nf_s and the unknown frequency f_x are impressed on a high frequency receiver, which is tuned to accept both. The higher beat frequency difference is obtained in the receiver output and is measured by tuning of the low frequency receiver (or by comparing with a suitable standard frequency in this low-frequency receiver, as previously described). The beat frequency difference lies between $f_s/2$ and f_s .

$$f_x = nf_s \pm f_b$$

Principal Limitations are:

- Standard frequency harmonics must be of usable magnitude in region around f_x .
- The frequency of the used harmonic nf_s must be determined.
- The sign of the beat frequency must be determined.
- The over-all accuracy depends on the calibration of the low frequency receiver.

Principal Advantages are:

- A high frequency standard can be used.
- The low frequency receiver, or interpolator, range is from $f_s/2$ to f_s , instead of from 0 to $f_s/2$ (Method 2); the interpolator is therefore more easily designed.
- Improved accuracy can be obtained by comparing f_b with a suitable frequency from the standard, in the low frequency receiver.

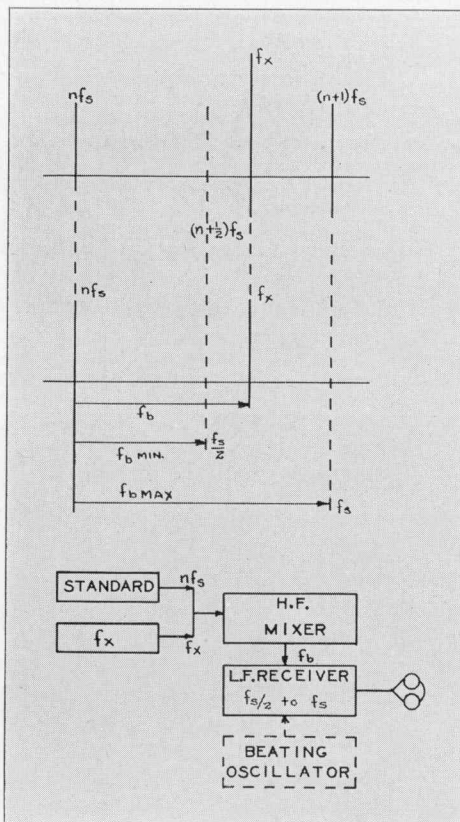


FIGURE 4. Interpolation and functional diagrams for Method 3.

Principal Limitations are:

- The variable frequency harmonics must be of usable magnitude in region around f_x .
- The frequency of the used harmonic must be determined.
- Result is on a fractional basis (not in cycles directly) unless added equipment is used.
- Accuracy of the order of 25 parts per million; can be greatly improved with added equipment.

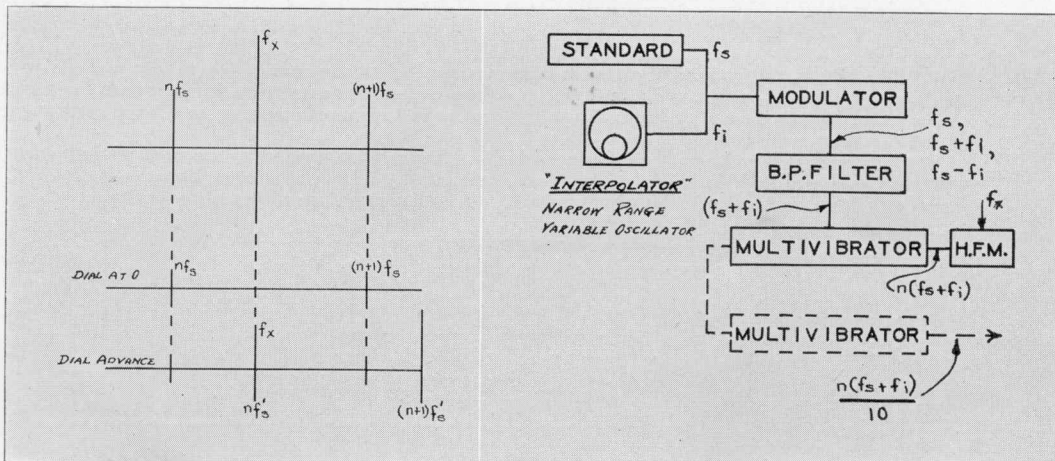
Method 4. Sliding Harmonics

The frequency f_x is impressed on a detector with the output of the variable harmonic standard. The frequency f_s is then increased to a new value f'_s such that $nf'_s = f_x$ (zero beat). From the standard controls, the change $nf'_s - nf_s$ is known. Then

$$\begin{aligned} f_x &= nf'_s = nf_s(1 + \Delta) \\ &= nf_s + nf_s\Delta \end{aligned}$$

where f_s is the known, fixed, standard frequency and Δ is the fractional change in f_s required to make $nf'_s = f_x$.

FIGURE 5. Interpolation and functional diagrams for Method 4.



Principal Advantages are:

- Only one zero beat setting is required for a measurement.
- A wide measurement range is covered by a narrow range oscillator.
- The full range of the variation control covers 1% or less of the measured frequency.
- The sign of the frequency increment is always positive.
- Wide-pass circuits, filters, etc., are avoided at f_x ; only a detector for setting zero beat is necessary.

Method 5. Cathode-Ray Oscilloscope

The three frequencies f_x , f_i , and f_s exist physically, where f_i = frequency of an interpolation oscillator, f_s = standard frequency.

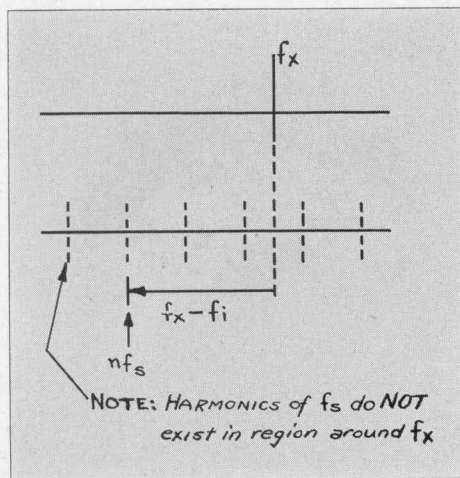
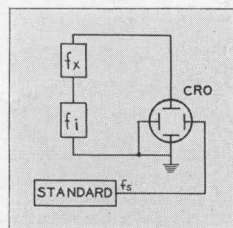


FIGURE 6. Functional block diagram for Method 5.



The used multiple $n f_s$ of the standard frequency does not exist physically; neither does the frequency difference $f_x - f_i$. These relationships are established by patterns on the oscilloscope screen.

The interpolator frequency is adjusted so that $f_x - f_i = n f_s$.

$$\text{Then } f_x = n f_s + f_i.$$

Principal Limitations are:

- The frequency f_x must be stable enough to observe a pattern.
- The used multiple $n f_s$ of the standard must be identified.
- The sign of the increment must be determined.
- Absolute accuracy depends on oscillator f_i and accuracy of setting.

Principal Advantages are:

- No harmonics are necessary.
- Extreme simplicity of equipment and operation.
- Wide ranges of f_x can be covered with suitable choice of f_s and f_i .

FIGURE 7. Interpolation diagram for Method 5.

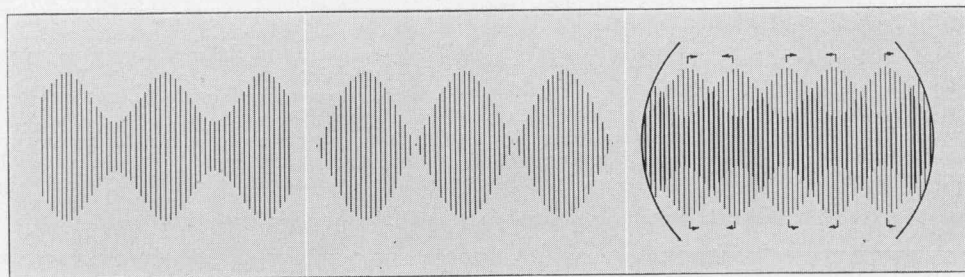


FIGURE 8. Patterns produced by two frequencies applied to one pair of plates of a cathode-ray oscillograph. Viewed against a linear sweep, two frequencies of unequal amplitude appear as shown at the left; with equal amplitudes as shown at the center. The envelope frequency is the difference of the two frequencies.

At the right is shown the pattern produced by two frequencies applied to the vertical deflecting plates, viewed against a sinusoidal sweep of large amplitude. When the pattern is nearly stationary, one envelope moves slowly in one direction, while the other moves slowly in the opposite direction, as indicated by the arrows.

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ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

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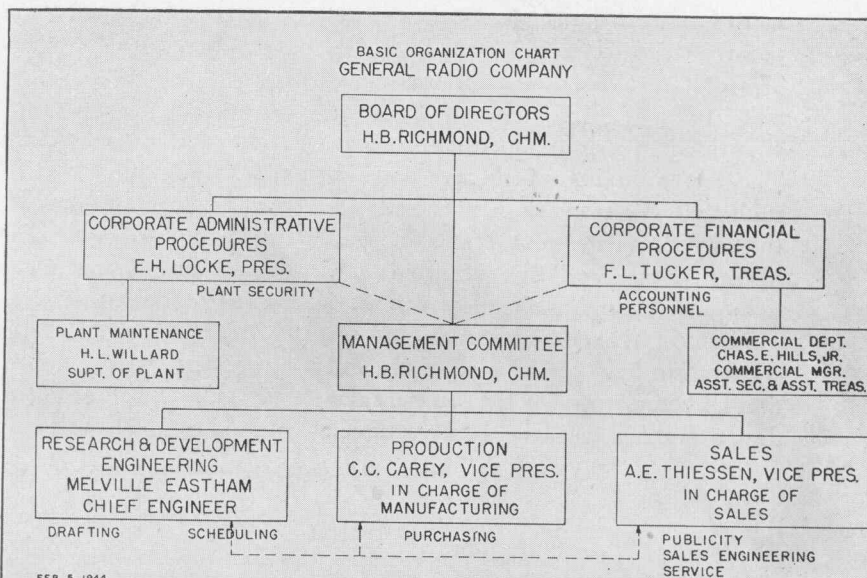
NEW CHAIRS — SAME PEOPLE

● ON JUNE 14, 1915, when Melville Eastham received from the Secretary of the Commonwealth of Massachusetts a charter to operate the General Radio Company he had not quite reached his thirtieth birthday. During the days of World War I, Mr. Eastham was joined

in the General Radio venture, first, by Henry S. Shaw, and later by Errol H. Locke. Shortly after returning from France in 1919 Harold B. Richmond also joined this group

Deflated from its violent war expansion, this new company in the fall of 1919 found itself in that twilight zone where either life or death could easily follow. Its eldest officer, Mr. Shaw, was not yet thirty-five, and the youngest of the group, Mr. Richmond, was still under thirty. Perhaps it is the recollection of the struggles of those early days when youth did not know defeat that have made this group of four so unani-

Functional organization chart of the management of the General Radio Company.





MELVILLE EASTHAM
Retiring President

mous in their decision that the active operation of the Company should be in the hands of young men.

It has further been the aim of the four principals of the Company to have it known as a group working together rather than the image of any individual. For this reason titles have not always been fully descriptive, while the active management of the Company stemmed from a group known as the Management Committee, made up of the officers and the chiefs of the operating divisions. The officers also felt that it was desirable to avoid making any sudden change in the over-all management. The age grouping of these officers is such that the transfer of responsibility to the younger men can best be accomplished if these officers simultaneously relinquish their functional positions.

Thus, at the annual meeting of the corporation held on February fifth, and at a Directors' meeting held on the same day, a number of changes in personnel,

and the revision of duties pertinent to the several positions, were made to bring about the above-mentioned desired results.

Messrs. Eastham, Locke, Richmond, and Shaw will continue as Directors and have been joined by Frank L. Tucker, formerly Comptroller, who was also elected Treasurer and Secretary of the Company. Mr. Eastham, formerly President, and Mr. Shaw, formerly Chairman of the Board, will hold no officership, although Mr. Eastham will temporarily continue in charge of research and development with title of Chief Engineer. Mr. Richmond was appointed Chairman of the Board of Directors and Chairman of the Management Committee. Mr. Locke, formerly Vice-President, was elected President.

New appointments, in addition to Mr. Tucker as Secretary and Treasurer, are Arthur E. Thiessen, formerly Commercial Engineering Manager, as Vice-President in Charge of Sales, and Charles C. Carey, formerly Superintendent, as Vice-President in Charge of Manufacturing.

A simplified sketch showing the new organization is given on page one. The Board of Directors now consists of Mr. Richmond as Chairman, Messrs. Eastham, Locke, Shaw, and Tucker. These same men with the addition of Messrs. Carey, Thiessen, and Charles T. Burke, Engineering Manager, comprise the Management Committee. Mr. Burke is also Chairman of the Scheduling Committee which is responsible for coordinating the activities of the three departments that report to the Management Committee.

Mr. C. E. Hills, Jr., who has been associated with the Company since 1922, and has been Commercial Manager since 1929, will continue in that



position, but in addition has been appointed Assistant Secretary and Assistant Treasurer of the Company.

Henry S. Shaw, the retiring Chairman of the Board, joined the Company in 1917 and was elected Treasurer the following year. He served in this capacity until 1926 when he became Chairman of the Board of Directors, which office he held continuously until February 5, 1944.

Mr. Shaw is a graduate of Harvard University in the Class of 1906 and before joining the General Radio Company was employed in the office of the treasurer of several New England textile mills. He has served as a director of a Boston trust company and as a trustee of a Boston savings bank, and is currently director of other business corporations. He has been interested in amateur radio since 1910, and it was this interest which was instrumental in bringing him to the General Radio Company during the first World War. His scientific memberships include:

- Fellow, Institute of Radio Engineers
- Fellow, American Academy of Arts and Sciences
- Fellow, American Association for the Advancement of Science
- Fellow, Acoustical Society of America
- Member, American Institute of Electrical Engineers
- Member, American Meteorological Society
- Member, American Geophysical Union.

Melville Eastham was one of the founders of the Clapp-Eastham Company in 1906. This company, whose name is familiar to all old-timers in the radio industry, first manufactured X-Ray apparatus, but started the manufacture of "wireless" equipment in 1908.

Mr. Eastham left the Clapp-Eastham Company in 1915 to become president of the General Radio Company, and he held this office until 1944. He has re-

cently spent nearly two years on leave of absence to the Radiation Laboratory at Massachusetts Institute of Technology, working on the design of military equipment.

His principal interest has always been in the design of instruments and components, and he is responsible for many of General Radio's small parts as well as some complex instruments.

One of his important interests outside the General Radio Company has been the Institute of Radio Engineers, in which organization he served many years as a director, and from 1927 to 1940 as treasurer. He was awarded the Institute Medal of Honor in 1937.

He holds, among others, the following memberships:

- Fellow, Institute of Radio Engineers
- Fellow, American Association for the Advancement of Science
- Member, American Institute of Electrical Engineers, Acoustical Society of America, American Physical Society, American Meteorological Society.



HENRY S. SHAW
Retiring Chairman



ERROL H. LOCKE
President

Harold B. Richmond, the incoming Chairman of the Board of Directors and Chairman of the Management Committee, joined the General Radio organization in 1919, becoming Secretary in 1921, Assistant Treasurer in 1924, and Treasurer since 1926.

He is a graduate of the Class of 1914, Massachusetts Institute of Technology, and for the past eleven years has been a member of the Corporation of that Institution. Since last year he has also been a member of the Corporation of Northeastern University. Mr. Richmond has served as Director and President of the Radio Manufacturers' Association, and as Vice-Chairman of the Board of the Scientific Apparatus Makers of America. He has been a director of a Cambridge trust company and is now a Corporator of the Home Savings Bank of Boston and also a director of the United Mutual Fire Insurance Company.

Before joining the General Radio

Company, Mr. Richmond had served with the Stone & Webster Management Association, the General Vehicle Company, and as a member of the electrical engineering staff of the Massachusetts Institute of Technology, which position was interrupted when he was called to active duty as a reserve officer in the Coast Artillery Corps in 1917. He is a Fellow of the American Institute of Electrical Engineers and of the Institute of Radio Engineers.

Errol H. Locke, our new President, joined the Company in 1918, becoming Vice-President in 1920, and holding that office until his election to the presidency on February fifth. Mr. Locke has been particularly interested in production and has devoted himself very largely to that phase of the Company's activities.

Mr. Locke is a graduate of Harvard College in the Class of 1913. Before joining the General Radio Company he was in the investment banking business.



HAROLD B. RICHMOND
Chairman of Board



His outside activities have largely centered around his native town of Lexington, Massachusetts, where he has served on various town committees and since 1939 has been one of its selectmen. He is a Director of the Lexington Trust Company, a Trustee of the Lexington Savings Bank, the President of the Lexington Home for Aged People, and Treasurer of the Lexington Historical Society.

Frank L. Tucker, Treasurer, received his B.S. degree in Electrical Engineering from the University of Texas in 1927, and his M.B.A. degree from the Harvard Graduate School of Business Administration in 1930. He was employed by the Scout and Lease Department of the Gulf Oil Company in 1928-29, and by Arthur D. Little, Inc., from 1930 to 1934.

He came to the General Radio Company as statistician in 1934 and was appointed Comptroller in 1937, where his activities have included many phases



ARTHUR E. THIESSEN
Vice-President for Sales

of Company financial procedure, including accounting, statistics, costs, and government regulations.

Mr. Tucker is a member of the National Association of Cost Accountants. He has been active in the Harvard Business School Alumni Association and the Greater Boston Community Fund.

Arthur E. Thiessen, Vice-President in Charge of Sales, graduated from Johns Hopkins University in 1926 with the degree of Bachelor of Engineering. After graduation he became a Member of the Technical Staff of the Bell Telephone Laboratories, leaving there in 1928 to join the engineering staff of the General Radio Company. For the next several years he was engaged in both development engineering and sales activities, becoming Commercial Engineering Manager in 1937.

Articles by Mr. Thiessen have appeared frequently in the *Experimenter* and in other technical publications. He



FRANK L. TUCKER
Treasurer



CHARLES C. CAREY
Vice-President for Manufacturing

is a Senior Member of the Institute of Radio Engineers, a Member of the American Institute of Electrical Engineers, and of the American Association for the Advancement of Science.

He is a Director of the Family Welfare Society of Cambridge and also a director of two local manufacturing concerns.

Charles C. Carey, Vice-President in Charge of Manufacturing, joined the General Radio organization as a winder in 1927 and became foreman of the winding department in 1928. He became assistant to Mr. Locke, who was then vice-president in charge of production, in 1931, in which capacity he supervised the production of special items and government contracts. While in this position he also reorganized the Company's production control system. In 1935 he was appointed production superintendent, holding that post until recently elected Vice-President.

Prior to his association with the General Radio Company, Mr. Carey was employed by the New England Telephone and Telegraph Company from 1922 to 1925, and by the Acme Apparatus Company from 1925 to 1927.

Charles E. Hills, Jr., Assistant Secretary and Assistant Treasurer, was graduated from Northeastern University in 1921 with the degree of Bachelor of Electrical Engineering. He came to General Radio in 1922, after a short period in the employ of the Boston Edison Company. After spending one year in the production department and five years in the sales department, he became Commercial Manager in 1929, which position he still holds.

Mr. Hills is a member of the National Office Managers' Association, the Export Managers' Club of New York, and the New England Export Club. He is an associate of the Institute of Radio Engineers.



CHARLES E. HILLS, JR.
Assistant Secretary and Assistant Treasurer



CHARLES T. BURKE
Management Committee

Charles T. Burke, Engineering Man-

ager, received his S.B. and S.M. degrees from the Massachusetts Institute of Technology in 1924. After graduation, he joined the engineering staff of the General Radio Company. Since that time his activities have included engineering, publicity, sales, and export. He was editor of the *Experimenter* from 1927 to 1929, and was appointed Engineering Manager in 1933. He has been a member of the Management Committee since 1939, and was appointed Chairman of the Scheduling Committee in 1943.

Mr. Burke is a Fellow of the Institute of Radio Engineers and a Member of the American Institute of Electrical Engineers. He has been active in municipal affairs in his home town of Watertown, Massachusetts, having served as a Town Meeting Member, and is at present Chairman of the Trustees of the Free Public Library of that town.

DOUBLE DUTY



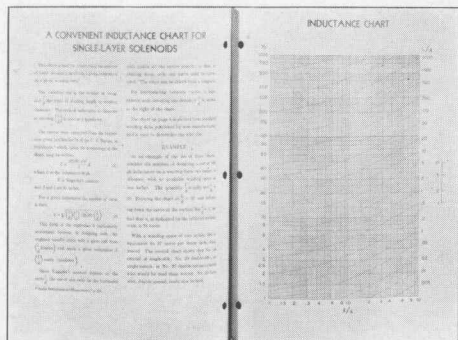
Supplementing the efforts of our 66 employees in the armed forces, 31 other General Radio employees contribute much of their time after working hours directly to military activities. Of this group, 23 are members of the U. S. Coast Guard Auxiliary, six others have joined the Massachusetts State Guard, and two women employees are members of the Women's Defense Corps.

NEW PUBLICATIONS

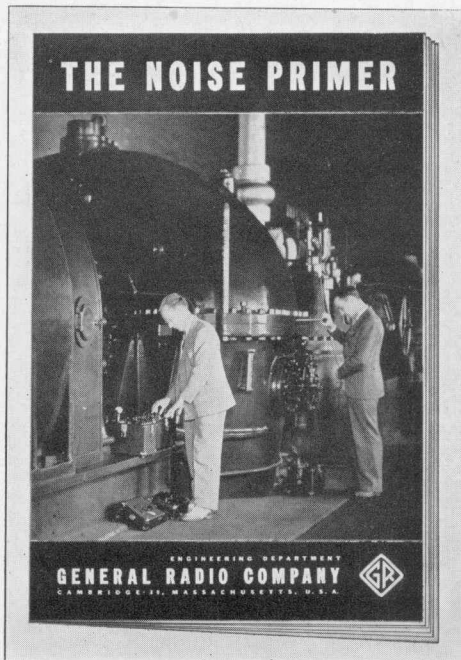
● **THE SERIES OF ARTICLES** by H. H. Scott, entitled "The Noise Primer," which ran serially in the *Experimenter* last year, is now available in booklet form.

The Noise Primer is the ABC of noise and vibration measurement. Standards, methods, and instruments are discussed in detail.

If you make noise and vibration measurements, or if you would like to learn more about the accepted methods of measurement, you will find this booklet valuable. A copy will be sent free on request.



● **THE INDUCTANCE CHART** originally published in the August, 1940, issue of the *Experimenter* and recently available in a 17 x 22-inch size for wall



mounting, has been reprinted in a 4-page, 8½ x 11-inch folder, punched for a standard 3-ring binder. This chart indicates the number of turns required for a given inductance in terms of the length and diameter of the winding form. An auxiliary chart determines the wire size to be used. If you can use this handy aid to the coil designer, write for a copy.

THE General Radio *EXPERIMENTER* is mailed without charge each month to engineers, scientists, technicians, and others interested in communication-frequency measurement and control problems. When sending requests for subscriptions and address-change notices, please supply the following information: name, company name, company address, type of business company is engaged in, and title or position of individual.

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A BRIDGE-CONTROLLED OSCILLATOR

● A BRIDGE-CONTROLLED OSCILLATOR, having particular advantages in use with low-frequency quartz crystals and low-frequency narrow-range tuned circuits, is shown schematically in Figure 1. The diagram indicates the arrangement for use with a 50- or 100-kc quartz crystal.

A single-stage high-gain tuned amplifier is followed by a triode used as a phase-inverter. The inverter is connected to the bridge input terminals; the bridge output terminals are connected back to the input of the high-gain amplifier.¹

First, consider the conditions for balance of the bridge; these are

$$R_1/R_2 = R_3/R_4$$

and

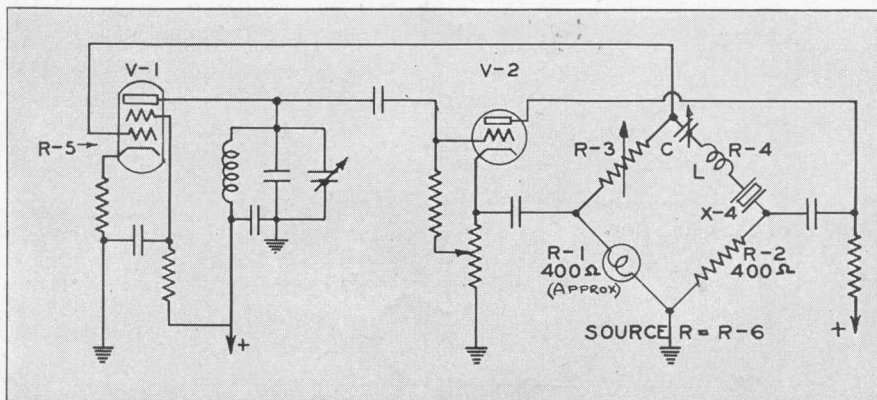
$$X_4 = 0.$$

At balance, the bridge output voltage is zero and the crystal reactance is zero, indicating that the crystal operates at exact series-resonance. Obviously no oscillations could be maintained under these conditions.

Next, if the ratio arm R-1 be given a value of resistance slightly less than that of the arm R-2, there will be a small output voltage from the bridge in the correct phase to maintain oscillations.

¹This is a substantially simplified modification of Meacham's oscillator. It is also readily adapted to operation over a narrow band of frequencies. "A Bridge-Stabilized Oscillator," L. A. Meacham, *Proc. I.R.E.*, Vol. 26, No. 10, Oct. 1938, p. 1278.

FIGURE 1. Schematic circuit diagram of the bridge-controlled oscillator.



This condition is not stable; the slightest change in bridge balance conditions or in the gain of the amplifier would cause the amplitude of oscillation to increase or decrease. If the value of resistance of the arm R-1 be made dependent on the voltage across it, then stable amplitude conditions will be maintained. In this case, the value of R-1 must increase if the voltage across R-1 increases. If R-1 is a tungsten filament lamp this condition is realizable. (In other arrangements of the bridge, a resistance whose value decreases with increasing voltage may be necessary. In such case a carbon filament lamp could be used.)

With a 115-volt 6-watt tungsten filament lamp operated at 0.6 to 3.0 volts, resistances from about 300 to 600 ohms will be obtained. In the region of 1.0 volt the resistance is near 400 ohms. Therefore, if $R_3 = R_4$ and if a voltage of approximately 2.0 volts is impressed on the bridge input terminals, a resistance balance will be reached.

As an oscillator, assuming first that the amplifier is tuned for zero phase-shift, the system will operate so that the lamp resistance is not quite equal to R-2. The equilibrium value will be such that the loss from the input terminals through the bridge to the output terminals is exactly equal to the gain from the bridge output terminals through the amplifier to the bridge input terminals. It is evident that very slight changes in the lamp resistance will very greatly affect the magnitude of the bridge output. A change in gain of the amplifier is then immediately compensated for by a very slight change in lamp resistance, which, in turn, is produced by an inappreciable change in the voltage across the bridge input terminals, or the level of operation of the oscillator.

The lamp-controlled bridge is therefore a form of automatic level (or "volume") control. In contrast to the usual forms of A.V.C. circuits it has several interesting and important properties. It has infinite cut-off, at balance, whereas A.V.C. circuits have restricted ranges. Also, since the lamp resistance cannot change materially over the time of one cycle of the oscillation frequency, the lamp-controlled bridge introduces no distortion or phase-shift as regulation takes place, with the consequent alterations of the frequency of oscillation.²

With conditions outlined above, the amplitude performance of the oscillator can be calculated completely since the system is entirely linear. With the bridge near balance, and with $R_3 = R_4 = 3000$ ohms (which would represent operation with a particular 50-kc plated quartz bar), the net load on each half of the phase-inverter would be slightly less than 400 ohms. The gain of the phase-inverter is:

$$A = \frac{E_{out}}{E_{in}} = \frac{2\mu}{r_P/R + \mu + 2}$$

where R is the net resistance of each half of the bridge load. The internal resistance of the phase-inverter stage is:

$$R_0 = \frac{r_P}{\mu/2 + 1}$$

With a tube such as the 6J5-G, $\mu = 20$, $r_P = 7600$ ohms and we find $A = 0.92$, $R_0 = 690$ ohms.

If we have nearly 2.0 volts across the bridge input terminals (to bring the bridge near balance), the input voltage to the phase-inverter stage will be $2 \div 0.92 = 2.17$ volts. If the Q of the tuned amplifier is 400 (which can easily be realized), then the input voltage

²"Constant Frequency Oscillators," F. B. Lewellyn, PROC. I.R.E., Vol. 19, p. 2063, December, 1931.



would be $2.17/400 = 0.0055$ volts. This voltage is obtained from the bridge output when the lamp resistance is approximately one per cent lower than the resistance of the fixed ratio arm, or 396.0 ohms. Each half of the bridge, consisting of 3000 ohms in parallel with 400 ohms, presents a load of 355 ohms to each half of the phase-inverter, or a total load of 710 ohms. With the internal resistance of the phase-inverter, R_0 , at 690 ohms, a very good power match is obtained from the inverter to the bridge.

The conditions outlined so far can be represented in a vector diagram as shown in Figure 2. To avoid confusion, the voltages across the upper and lower bridge arms, and the bridge output voltage, are shown on separate lines.

If next we consider that the amplifier phase-shift is not zero, then the crystal phase-angle must assume such a value that the phase-shift through the bridge is equal and opposite to the phase-shift in the amplifier, or the total phase-shift around the entire loop is zero. These conditions are outlined in the diagram of Figure 3.

From the diagram, even though it is not to scale, it is evident that a given amplifier phase-shift is compensated by the bridge with the crystal phase-shift much smaller than that of the bridge. Since the crystal phase-shift is brought about by a change in operating frequency, the small value of phase-shift means a correspondingly small change in frequency.

Finally, in practice, we require a means of adjusting the operating frequency over a small range around the true crystal series-resonant frequency. This is accomplished by placing a series-tuned L-C circuit in series with the crystal. When this circuit is tuned so that its reactance is zero at the crystal

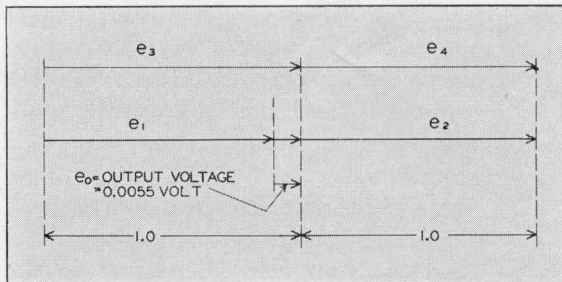


FIGURE 2. Vector diagram for operation with zero phase-shift in amplifier.

frequency, and the amplifier is adjusted by test for zero phase-shift, then the operating frequency is the series-resonant frequency of the crystal. If the capacitance of the tuned circuit is made larger than the resonant value, the operating frequency is reduced; if smaller, it is increased.

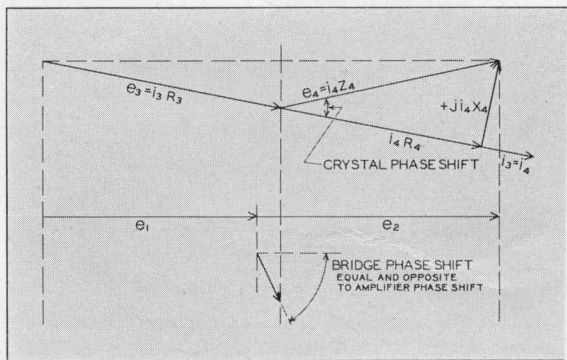
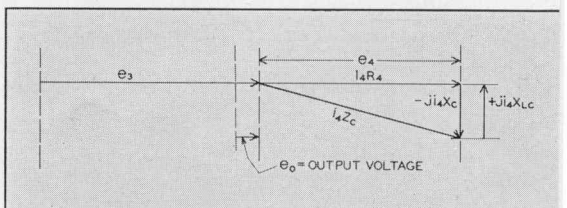


FIGURE 3. Vector diagram for operation with amplifier phase-shift equal and opposite to bridge phase-shift.

These conditions are shown in Figure 4, where the *net* reactance of the series tuned L-C circuit is X_{LC} and the *net* reactance of the crystal is X_C .

In actual operation, the resistance of the crystal is generally unknown, so that

FIGURE 4. Vector diagram for operation with reactance added in series with the crystal.



the adjustment of R_3 to equal R_4 cannot be made in advance. Consequently, it is convenient to operate the system so that $e_1 = e_2$, which can be checked experimentally, adjusting R_3 to obtain this condition. Since the resistances of the arms R_3 and R_4 are substantially higher than those of the ratio arms R_1 and R_2 , considerable latitude in the values of R_3 and R_4 is possible without substantially altering the voltage developed across the ratio arms. Consequently, a rudimentary vacuum-tube voltmeter can be built into the system indicating the voltage at the grid of the phase-inverter. Such a voltmeter is also necessary to indicate when oscillations take place. Since the system is linear, plate- or grid-current meters are of no value for this purpose. A particular reading of this voltmeter then corresponds to the condition that $e_1 = e_2$. Whatever the value of the crystal resistance R_4 , within reasonable limits for similar crystals, R_3 is adjusted to obtain the proper reading of the voltmeter.³ When this is done, the value of R_3 exceeds the value of R_4 by a very small amount, the amount being inversely dependent on the gain of the amplifier.

Operation by this method opens up the possibility of measuring the characteristics of the crystal. From the above, the value of R_3 slightly exceeds the re-

³As in the Types 675-P and 690-D Piezo-Electric Oscillators.

sistance of the crystal plus the resistance of the inductor. For comparative purposes these need not be separated.

To determine the crystal reactance, we adjust the L-C circuit to introduce known values of reactance X_1 and X_2 . Then:

$$\begin{aligned} X_1 + X_C &= 0 = X_1 + 2\Delta_1 X_0 \\ X_2 + X_C &= 0 = X_2 + 2\Delta_2 X_0 \end{aligned}$$

Where Δ_1 and Δ_2 are the resulting frequency deviations and X_0 is the resonant reactance of the crystal. Then

$$X_0 = \frac{X_2 - X_1}{2(\Delta_2 - \Delta_1)}$$

Using this method, the resistance of a particular type of 50-kc quartz bar was 2700 ohms, its resonant reactance was 228×10^6 ohms. The "Q" of the bar, X_0/R , was 84,500.

The method is very useful in comparing the performance of bars in production. The resonant reactances of such bars are all the same, since the reactance depends on the cut and dimensions. The effective resistance of the bars is also determined by the cut and dimensions, but *in addition* depends upon many other factors such as the type and condition of the mounting device, supersonic radiation and reflections. Where identical bars are being processed, these factors can be controlled so that the observed resistances are closely the same.

J. K. CLAPP

THE EFFECT OF HUMIDITY ON ELECTRICAL MEASUREMENTS*

● AS SUMMER APPROACHES, we prepare for the usual crop of complaints that come in regarding the erratic behavior of electrical measuring equipment, particularly precision impedance bridges. The various stories con-

flict. Sometimes the bridge seems to read high, sometimes low; the balance may drift badly or suddenly jump to a new value; it may show increasing errors as the frequency is lowered, or the balance may shift with applied voltage. At first glance it seems that there must be as

*Reprinted from the August, 1943, issue of the *Experimenter*.



many causes as kinds of error. Actually, there is just one cause, high relative humidity.

High relative humidity affects insulation in two ways. If the insulation is porous, moisture will be absorbed into the volume of the material, while if moisture wets the surface, a thin film of water is formed covering the whole surface. Water distributed throughout the interior of an insulator produces interfacial polarization which causes an increase in capacitance, dissipation factor, and volume conductivity. The amounts of these increases vary with the relative humidity and inversely with the frequency.¹ At 100% relative humidity and a frequency of 60 cycles, increases as much as 50% in capacitance, of a million-fold in conductivity, and up to a dissipation factor of 1.0, are quite possible for such porous materials as filled and laminated thermo-setting plastics, many thermoplastics and natural fibers like cotton, wool, and silk. The rate at which a porous material absorbs water and the tenacity with which it holds it depend greatly on the cross section of the pores. When these approach molecular dimensions, as in silica gel, which is a silicate having microscopic pores produced by suitable heat treatment, the material acts as a desiccant, and the absorbed water can be removed only by heating above the boiling point. Mica and some ceramics act in this manner. Only quartz and most glasses, some steatites, polystyrene, and a few other polymers are free from volume absorption and the accompanying deterioration of dielectric properties.

The formation of a surface film of water on an insulator is determined by the ease with which water wets the surface,

which in turn is measured by the contact angle between the surface and a drop of water on the surface.² Most of the porous materials that show large volume absorption also wet very easily. A microscopic roughness of the surface helps film formation. Quartz, glass, and steatite also wet easily. Only wax, polystyrene, and some other polymers successfully prevent the formation of a continuous film. The condition of the surface is also important. Dust and particularly acid perspiration from handling greatly aid wetting. The conductivity of even a thin film is enormous. Merely breathing on the surface of a good insulator like quartz will lower the insulation resistance between terminals spaced $\frac{3}{4}$ inch apart from above 10 MΩ to below 1 MΩ. A film so formed will vanish rapidly if the surface is chemically clean and the relative humidity low. On a dirty surface, however, the film persists and can be removed only by thorough cleaning or by heating.

Because there are no rigid stable insulators which are unaffected by moisture, it is customary to impregnate them or at least to coat their surfaces with one of the water-repellent substances such as wax, polystyrene, or the newer silicon resins. Any of these materials operates successfully on the non-porous insulators, such as glass and steatite, so long as perfect adhesion is maintained. Large changes in temperature, particularly toward freezing, will produce cracking and chipping of the surface material because of the differences in temperature coefficients of linear expansion, and because most coatings, the waxes in particular, become brittle at the lower temperatures. Any moisture film which

¹R. F. Field, "Frequency Characteristics of Decade Condensers," *General Radio Experimenter*, Vol. XVII, No. 3, Oct., 1942, pp. 1-7.

²On the well waxed surface of an automobile during a rain, large drops stand with a small re-entrant angle, the area of contact being smaller than the maximum section of the drop.

then forms between the insulation and the coating persists and can be removed only by complete cleaning of the surface or by heating.

On porous materials a thin protective coating is of no value because even the waxes are themselves somewhat porous. Such a coating decreases the rate at which moisture penetrates to the inner material, but continued exposure to high relative humidity will eventually result in the same equilibrium conditions. For reasonable success, wax coatings must be heavy, the result of multiple dippings, and of the order of 0.1 inch.

In all General Radio instruments great care has been taken to provide adequate protection against high relative humidity. All solid dielectric condensers are hermetically sealed or heavily waxed. All high-valued resistors are waxed or similarly protected. All steatite insulation is protected by a surface coating by the manufacturer. Mica-filled phenolic or polystyrene is used as insulation for mounting all high impedances. Wires for cables are insulated with rubber or a synthetic, with a lacquered identifying braid. These precautions are sufficient to allow normal operation under 90% relative humidity at 90° F. of all instruments except the 0.1% impedance bridges. An even more severe test occurs when, at 90% relative humidity, the temperature fluctuates sufficiently to reach the dew point and cause direct moisture condensation. Under these conditions the operation of an instrument may not meet catalog specifications. If power is dissipated inside the instrument, the heat generated will quickly evaporate the conducting moisture films. Otherwise some time must elapse to allow natural evaporation. A 40-watt lamp or other resistive load maintained inside the case will usually prevent condensa-

tion. It will have, however, little effect on moisture absorption.

The TYPE 716-B Capacitance Bridge is probably as greatly affected by moisture as any of our instruments. All steatite insulated terminals are wax coated, the input transformer is wax sealed, and the bridge wiring is open bus. Only the TYPE 722 Precision Condenser, used on the capacitance standard, is affected by moisture, and that only in its dielectric losses, not in its capacitance. Its own dissipation factor is defined by its figure of merit $F = DC = 0.04 \mu\mu f$, corresponding to a dissipation factor of the steatite stator support of 0.004. At about 60% relative humidity, moisture absorption through the wax coating on these bars causes their dissipation factor to rise. A tenfold increase at 90% relative humidity must be expected. This will produce a negative error in the direct reading of dissipation factor exactly equal to the increase in dissipation factor of the precision condenser. No error will appear in parallel substitution measurements.

Another dielectric loss occurs in aluminum-plate air condensers under high humidity conditions from the absorption of moisture by the aluminum oxide on the surface of all the plates. In its dry state aluminum oxide has a small dissipation factor and imparts to the whole condenser a dissipation factor of only 0.0000001, since its contribution is proportional to the ratio of the thickness of the oxide film, about 10 millionths of an inch, to the plate spacing of 30 mils. When exposed to moisture the dissipation factor of the oxide is an exponential function of the relative humidity,³ increasing a decade of dissipation factor for every 15% rise in relative humidity.

³A. V. Astin, "Nature of Energy Losses in Air Capacitors at Low Frequencies," *Journal of Research of the National Bureau of Standards*, Vol. 22, No. 6, June, 1939, pp. 673-695.



At 90% the air condenser has a dissipation factor of about 0.01. Since the dielectric loss occurs on the surfaces of all the plates, the air condenser behaves like a variable solid dielectric condenser, and its dissipation factor does not cancel out even in parallel substitution measurements. As before, the error in the bridge reading is negative.

Both of these kinds of moisture absorption in steatite and in aluminum oxide are troublesome when the relative humidity stays at 60% throughout the day, since at this level only a half day is required to attain equilibrium. A relative humidity of 40% causes no appreciable error, and even a rise to 60% followed by a drop back to 40% within six hours will cause little trouble. The time needed to attain equilibrium increases with the relative humidity and is at least three days at 90%.

Every laboratory should be equipped with some type of hygrometer in order that the possibility of errors in bridge measurements may be anticipated. The ordinary hair hygrometer is very useful in spite of its large errors because it is of the indicating type. A wet and dry bulb hygrometer should be available as a check under extreme conditions. Regular readings of relative humidity during the summer months are fully as important as are those of temperature, for the units being measured are in many

cases more liable to be affected by high humidity than the measuring equipment itself. This fact indicates that the bridge readings of dissipation factor may appear to be high or low dependent upon the relative rates at which the unknown condenser and the standard condenser in the bridge change their dissipation factors with humidity. Without a hygrometer this situation cannot be definitely recognized except as the reading of dissipation factor becomes ridiculously low or actually negative. A certain instability of bridge balance will appear at relative humidities above perhaps 70%, as indicated by a more or less steady drift of both capacitance and dissipation factor balance points.

There is little that can be done with existing measuring instruments to eliminate this type of error, short of air conditioning. This should preferably apply to the entire room containing measuring equipment. Then the unknown unit is measured under standard conditions. It is also possible to dry out the measuring instrument by placing a desiccant, such as silica gel, inside its case. However, the amount of moisture which can seep through the joints between panel and case and around the control shaft is amazing. Unless unusual care is taken, it will be necessary to renew the desiccant each working day whenever the relative humidity is above 70%.

— ROBERT F. FIELD

SCHEDULED DELIVERY DATES

● **MOST OF OUR CUSTOMERS** who have to deal directly with procurement are familiar with the scheduling system under which electronic test equipment is now delivered. To our engineering customers whose principal interest is "when do I get it" rather than a detailed explanation of delays in deliv-

eries, several words to clarify the predicament of General Radio seem to be in order.

Early in the war it became apparent that there would be severe shortages of manpower and materials and that test instrument manufacturers, sooner or later, would develop large shortage lists.



The materials shortage was due not only to a scarcity of certain basic raw materials and the necessary manpower to process them, but also to an even greater shortage of a considerable number of finished components which instrument contractors purchase from subcontractors.

At first manufacturers attempted to establish the delivery sequence of orders from the urgency statements of the customer. The priority system set up at the beginning did not cover test equipment and some other tools of production. Often the axle which squeaked the loudest got the most oil, since the manufacturer had no inside knowledge of the basic war plan nor any information as to how an order for a particular instrument fitted into the war picture. It became necessary for closer supervision by the governmental agency which had been set up to determine the priority of orders for test equipment. This function is now administered by the War Production Board, in collaboration with the procurement agencies, which decides the relative importance of orders placed by the Army, Navy, Maritime Commission, the British, Russian and Chinese Governments, and all other branches of the United Nations war effort.

For some time the sequence of shipments to these agencies has been decided solely by the tactical urgency of the war problem involved. Sub-divisions of each agency set the urgency of orders within the agency.

When anyone places an order for electronic test equipment it is now required that very complete supporting information be supplied to the WPB as to the tactical urgency in order that the proper delivery schedule may be established. This schedule is based upon two things:

the production capacity of the manufacturer (already expanded to the limit) and the urgency of the war problem involved.

After a delivery date has been set by WPB, the customer and the supplier are notified. The order then takes its place in the manufacturer's delivery schedule. The manufacturer makes every effort to meet the delivery date and his efforts are supplemented by those of the Army-Navy Electronics Production Agency, which is in close touch with changes in urgency and with any shortages and bottlenecks which may develop. All changes in scheduled deliveries are controlled by the WPB, and become necessary principally from two causes:

- (1) From frequent changes in the war plan, the urgency standing of many orders changes from day to day.
- (2) Due to shortages of raw materials, finished components, and manpower, the manufacturer's production schedule lags.

When it becomes necessary for WPB to change a delivery date, the manufacturer has no prior knowledge of that fact, nor any information as to the reason for the change. He is required by law to fill all orders in the sequence decided upon by the WPB schedules. Appeals directly to the manufacturer to reshuffle deliveries accordingly cannot be acted upon.

The present scheduling system at times results in repeated disappointments to our customers. Its whole purpose, however, is to supply war materials when and where they are most urgently needed. This, after all, seems to be the basic aim of war production.

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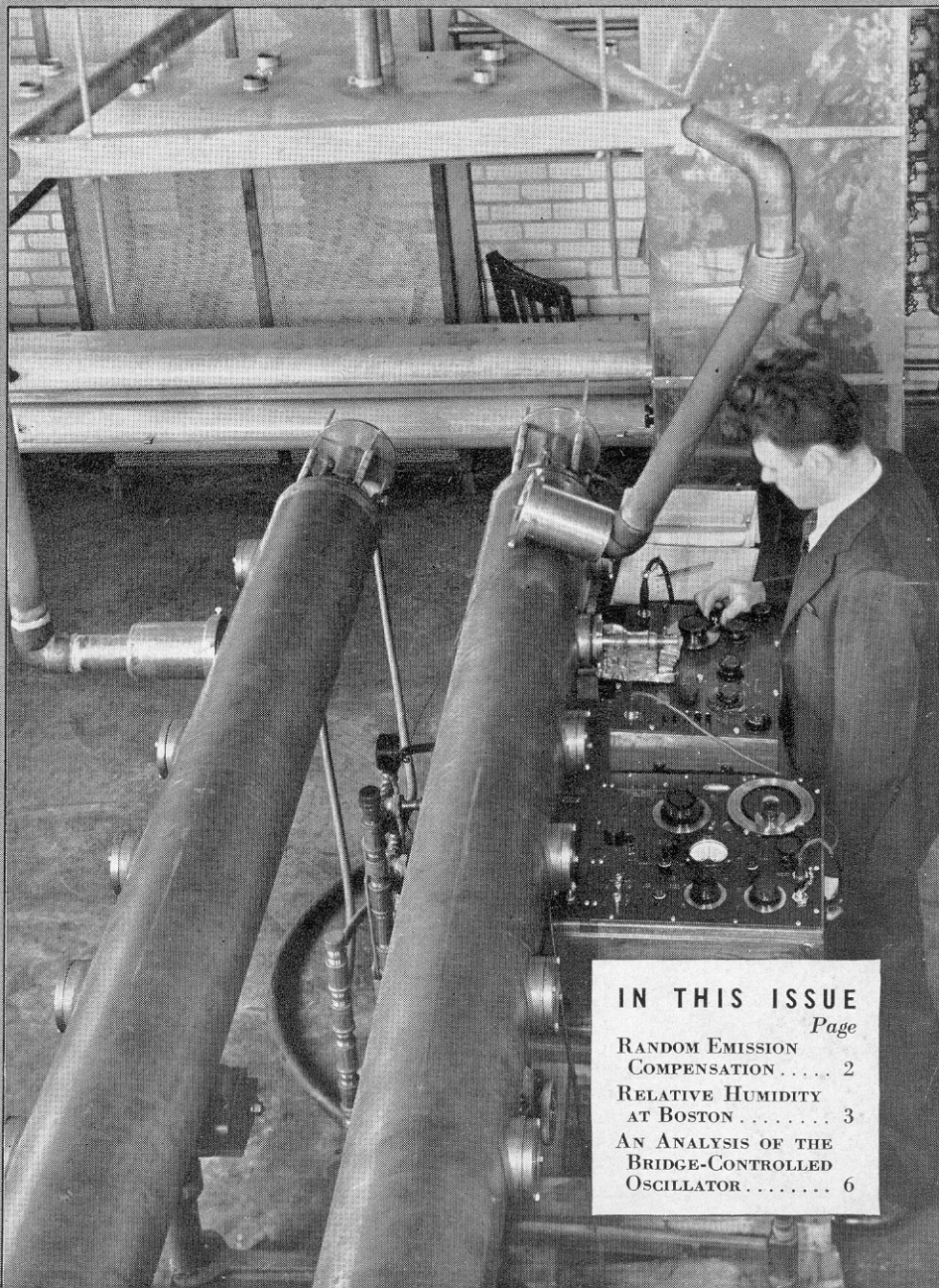
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VOLUME XVIII No. 12

MAY, 1944

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS



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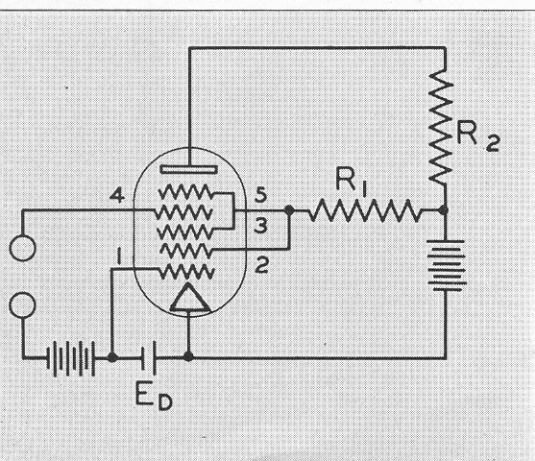
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RANDOM EMISSION COMPENSATION

● OWING TO ITS AMPLIFYING PROPERTIES and its high input resistance, the vacuum tube is ideally suited to the problem of measuring d-c potentials. Inherently stable circuits employing degeneration and a closely regulated power source are required if the vacuum-tube voltmeter is to be used for measuring small potentials. As a further refinement, it is necessary to shield magnetically and electrostatically the most sensitive tube or tubes of the circuit; all leads entering the shielded compartment must be carefully filtered; shock mounting is needed to avoid microphonics.

When these precautions have been taken, the sensitivity may be increased to the point where irregularities in the electron stream are the dominant cause of unwanted fluctuations in the output. These irregularities cannot be eliminated. They are due to the very nature of electron emission: electrons are freed from the cathode or filament in a random fashion; they push and jolt one another in a mad scramble to escape the heated emitter very much like popcorn on a hot stove and it is only because of their vast numbers that in normal usage they present a seeming constant flow.

FIGURE 1. Schematic diagram showing tube with grids connected to compensate for random emission.



Since the random fluctuations cannot be eliminated, the only recourse is to devise a compensating mechanism and this immediately suggests the use of a multi-electrode tube wherein several avenues are available for getting at the electron stream.

Pentagrid-converter-type tubes are supplied with several grids interposed between the cathode and the plate. Such a tube can be connected as shown in Figure 1 at no sacrifice in gain, and, if the operating potentials are properly chosen, fluctuations in emission will produce no change in the output voltage. The mechanism of compensation operates as follows: an increase in electronic emission causes more current to flow through R_1 ; the voltage at grids No. 2, No. 3, and No. 5 is reduced because of the voltage drop in R_1 . Since grid No. 5 is between the control grid (No. 4) and the plate, the current flow at the plate is affected by two compensating effects: the increase in electronic emission and the decrease in grid No. 5 voltage. As a net result, the current flow at the plate remains unchanged in spite of random fluctuations in the electronic stream.

Figure 2 illustrates the degree of compensation that may be expected. To simulate random fluctuations the filament voltage of a 1D7-G type tube was changed over what is really an enormous range, considered in terms of normal emission fluctuation. Since we are interested in the ability of the device to measure small voltages, the change in output, due to filament voltage changes, was corrected by introducing a voltage at the control grid. This "equivalent" voltage is plotted in the figure and the

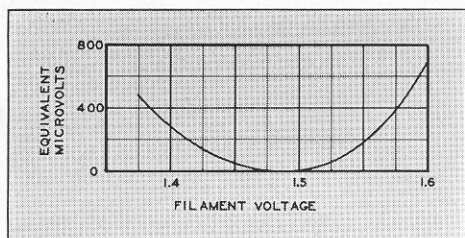


FIGURE 2. Plot showing range over which compensation is effective.

slope of the curve is the criterion of the effectiveness of compensation. The tube of course should be operated at the inflection point of this curve. At the time these data were taken, it was also ascertained that similar results could be obtained with a heater-type of tube (6A8), and it can be expected that other types of pentagrid-converter tubes would yield equally satisfactory results.

High stability can often be attained by using additional stages of amplification and introducing practically com-

plete degeneration over the entire amplifier. But any fluctuation in the voltages that exist in the path common to the control grids (contact potentials, etc.) is not degenerated out because the source of these voltages is not included in the degenerative loop.

Fortunately, when the circuit of Figure 1 is adjusted to compensate for changes in emission, it is equally balanced for changes in any voltage (E_D) which is common to both control grids. In other words the transconductance of the two grids (No. 1 and No. 4) to the plate are equal but of opposite polarity.

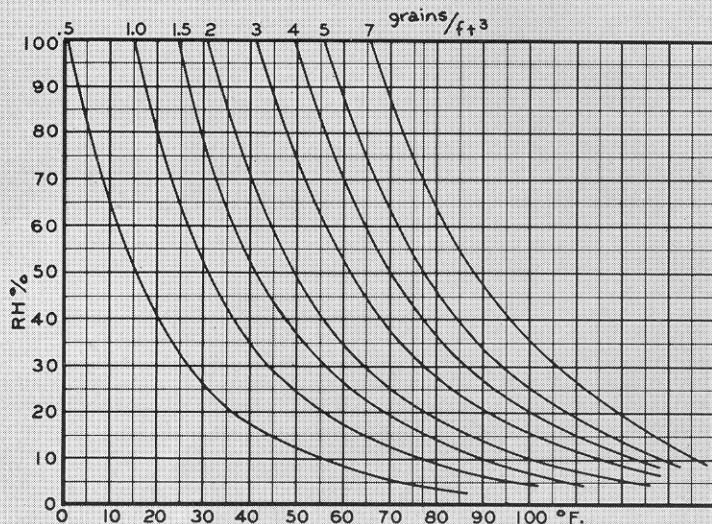
Consequently, the circuit of Figure 1 not only compensates for fluctuations in emission, but it can also be used to supplement the over-all degeneration and the heater supply regulator in the attempt to achieve the high stability so essential in measuring small d-c potentials and currents. — A. G. BOUSQUET

RELATIVE HUMIDITY AT BOSTON

● IN THE ARTICLE on "The Effect of Humidity on Electrical Measurements," reprinted in the *General Radio Experimenter* last month, mention was made of the fact that, whenever the rela-

tive humidity exceeds 60% for a day, aluminum plate condensers have appreciable losses, and when the relative humidity passes 70% the balance of a capacitance bridge becomes unstable.

FIGURE 1. Curves showing the relation between relative humidity and temperature for various densities of aqueous vapor expressed in grains per cubic foot.



The question naturally arises, how often during the year do these conditions occur? This of course depends on the location of the apparatus. In New England the humidity indoors is high only during the summer months of July and August, while in the tropics the humidity is always high.

For the last two years a record has been kept of the relative humidity in a room whose temperature is kept at 25°C. except when the outdoor temperature is higher. The relative humidity remains below 30% for the six winter months, November through April, and rises to an average of 60% in July. A value of 60% is exceeded once in May and October, about once a week in June and September, and almost continually in July and August. A value of 70% is exceeded about once a week in July and August.

The relation between the relative humidity indoors and outdoors is determined by the temperatures of the two locations. Assuming that the weight of water in a unit volume remains con-

stant, the change of relative humidity with temperature is given in Figure 1. As an example, assume that conditions outdoors are 70% RH at 40°F. Then entering the chart at this point and following down the curve of 2 grains per cubic foot gives 20% RH at 77°F. (25°C.).

Values of relative humidity and temperature are observed by the U. S. Weather Bureau at many points in the United States and published monthly in Weather Bureau Reports and daily in some newspapers. The monthly averages for Boston are given in Figure 2 for the average over a period of years and also for the year 1943, which appears to be a reasonably normal year. Average relative humidity is essentially constant at 72% with only a slight rise in August and September. This condition is mainly due to the fact that New England has an approximately constant monthly precipitation of about 3 inches. Proximity to the ocean is a contributing factor. This fact makes average relative humidity at constant temperature solely a

FIGURE 2. Curves of average monthly relative humidity and temperature at Boston. Note that the relative humidity at 77°F., calculated from outdoor values, agrees well with that observed in the constant-temperature room.

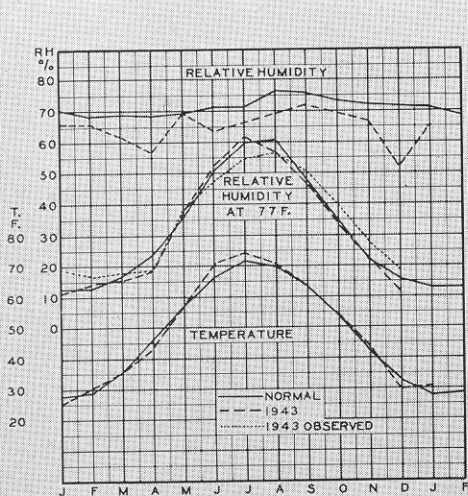
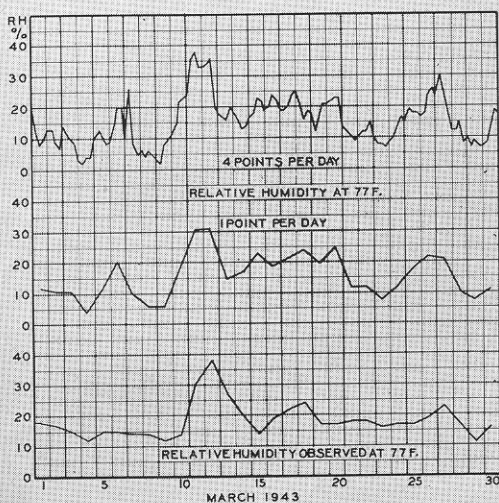


FIGURE 3. Curves of observed and calculated relative humidity at 77°F. for March, 1943, at Boston. The relative smoothness of the lower curve compared to the middle one illustrates the filtering and storage action of a large building.



function of outdoor temperature. The curve of relative humidity at 77°F. (25°C.), obtained by means of the curves of Figure 1, follows exactly the curve of temperature. An average value of 60% can occur only during the summer when the outdoor temperature is at least 70°F. The average values observed in the constant temperature room are also plotted in Figure 2. Their agreement with the calculated values justifies that calculation. There is an interesting time lag between the calculated and observed curves which can perhaps be explained as indicating a storage of moisture in the building. This would reduce the humidity in the early summer and increase it in the early winter.

Daily values of relative humidity and temperature show much greater variations than monthly values, as is well shown in Figure 3 for March, 1943. The diurnal changes are such that temperature tends to have a maximum value and relative humidity a minimum value shortly after noon. The wide fluctuations in relative humidity are caused partly by the temperature changes and partly by changes in the water content of the air. This is shown by the upper curve of Figure 4, which gives the relative humidity at the constant temperature of 77°F. (25°C.). The fluctuations have a longer period than before. The major changes still appear even when the daily average values are plotted as shown in the middle curve of Figure 4. The lower curve of Figure 4 shows the relative humidity as observed in the constant temperature room. Here the

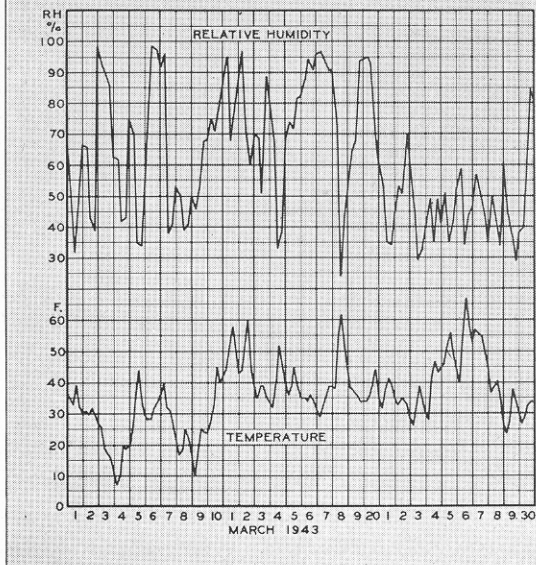


FIGURE 4. Daily values of relative humidity of temperature for March, 1943, at Boston. Over short periods of time relative humidity drops when temperature rises, as would be expected for constant water content of the air.

fluctuations are smaller and of still longer period. This is caused by the filtering and storage action of the building.

This comparison of observed values of indoor relative humidity with values calculated from outdoor data furnished by the U. S. Weather Bureau indicates that, whenever values of relative humidity are given in the daily newspapers, it is possible with the lag of only one day to determine indoor relative humidity without depending upon a hair hygrometer or wet and dry bulb thermometers. In many cases this method will be found satisfactory for indicating when aluminum plate condensers will begin to introduce errors in dissipation factor in capacitance bridge measurements and when the bridge balance itself will become unstable. — ROBERT F. FIELD

AN ANALYSIS OF THE BRIDGE-CONTROLLED OSCILLATOR

● **THE BRIDGE-CONTROLLED OSCILLATOR***, particularly as used with quartz bars in frequency standards, has a high degree of inherent stability of frequency. The principal properties of this circuit can be analyzed as follows:

Referring to Figure 1, the impedances involved are those of the bridge, R_1 , R_2 , R_3 , R_4 , and $+jX_4$ (the effective resistance and the net reactance of the quartz bar, operating very near to series resonance), the resistance of the source R_6 , and the resistance R_5 of the bridge output load, which is assumed to be very high compared with all of the preceding resistances.

For the bridge, the attenuation factor can be written:†

$$\beta = \frac{E_{out}}{E_{in}} = \frac{AR_4 - jBX_4}{MR_4 + jNX_4} \quad (1)$$

where

$$A = R_5(R_2R_3 - R_1R_4) \quad (2)$$

$$B = R_1R_4R_5 \quad (3)$$

*J. K. Clapp, "A Bridge-Controlled Oscillator," *General Radio Experimenter*, April, 1944.

†K. S. Johnson, "Transmission Circuits for Telephonic Communication," pp. 284-285.

$$\begin{aligned} M &= (R_1 + R_2)(R_3R_4 + R_5R_6) \\ &+ (R_3 + R_4)(R_1R_2 + R_5R_6) \\ &+ (R_5 + R_6)(R_1R_4 + R_2R_3) \\ &+ R_5(R_1R_3 + R_2R_4) \\ &+ R_6(R_1R_2 + R_3R_4) \end{aligned} \quad (4)$$

and

$$\begin{aligned} N &= R_4(R_1 + R_3 + R_5)(R_2 + R_6) \\ &+ R_1R_4(R_3 + R_5) \end{aligned} \quad (5)$$

For the system, the condition for oscillation is

$$\mu\beta = 1 \quad (6)$$

If there is a phase-shift in the amplifier, we can write*

$$\mu = \mu_1 + j\mu_2 \quad (7)$$

Inserting (1) and (7) in (6), we obtain:

$$(\mu_1 + j\mu_2) \frac{AR_4 - jBX_4}{MR_4 + jNX_4} = 1 \quad (8)$$

and, separating reals and imaginaries:

$$\mu_1AR_4 + \mu_2BX_4 - MR_4 = 0 \quad (9)$$

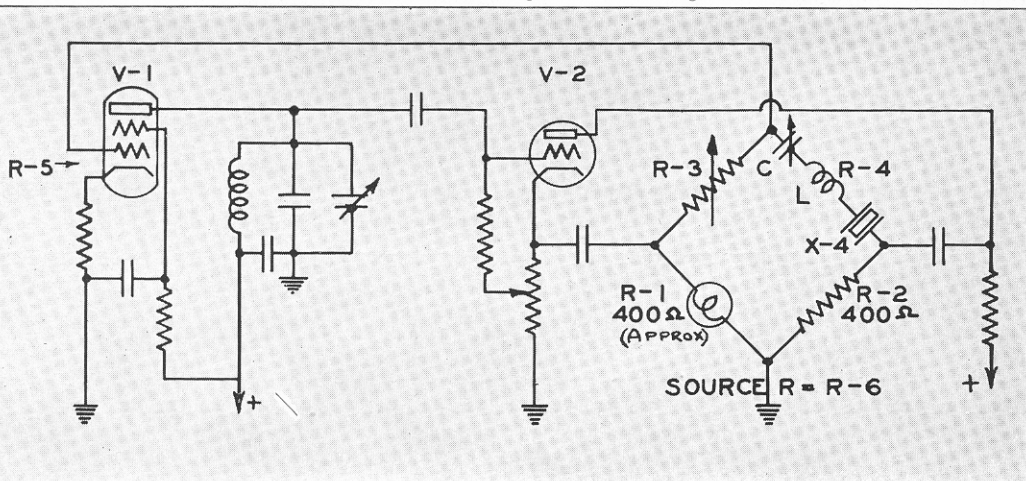
$$\mu_2AR_4 - (\mu_1B + N)X_4 = 0 \quad (10)$$

If there is no amplifier phase-shift, $\mu_2 = 0$, so that

$$\mu_1 = M/A = \mu_a \quad (11)$$

*L. A. Meacham, "The Bridge-Stabilized Oscillator," *Proc. I.R.E.*, Vol. 26, No. 10, p. 1278, October, 1938.

FIGURE 1. Schematic circuit diagram of the bridge-controlled oscillator.





and

$$X_4 = 0 \quad (12)$$

where μ_a is the magnitude of the amplification of the amplifier, from bridge output to bridge input. The frequency is then independent of the circuit except the crystal and the crystal must operate at exact series resonance.

If μ_2 is very small, so that θ , the phase angle of the amplifier, is small, from (9) and (10) we obtain

$$\begin{aligned} X_4 &= \frac{\mu_2 A R_4}{\mu_1 B + N} = \frac{M R_4}{\mu_1 B + N} \cdot \frac{\mu_2}{\mu_1} \\ &= \frac{M R_4}{\mu_a B + N} \tan^{-1} \frac{\mu_2}{\mu_1} \\ &= \frac{M R_4 \theta}{\mu_a B + N} \end{aligned} \quad (13)$$

since θ is small and $\tan \theta \cong \theta$.

For the crystal, in the region near series resonance, we can express the frequency deviation from resonance as a fraction of the resonant frequency:

$$\frac{f - f_0}{f_0} = \frac{X_4}{2\omega L_4} = \frac{X_4}{2Q R_4} \quad (14)$$

where X_4 is the net crystal reactance, that is, the difference of the inductive and capacitive reactances, and $Q = \omega_0 L_4 / R_4$.

Substituting (13) in (14) we obtain:

$$\frac{f - f_0}{f_0} = \frac{1}{2Q} \cdot \frac{M\theta}{\mu_a B + N} \quad (15)$$

We can now find the effects of changes in the amplifier on the frequency. For changes in phase only (μ_a remaining constant),

$$\left. \frac{df}{f_0} \right|_{\theta} = \frac{M d\theta}{2Q(\mu_a B + N)} \quad (16)$$

For changes in amplification only (phase

remaining constant):

$$\left. \frac{df}{f_0} \right|_{\mu_a} = - \frac{B M \theta d\mu_a}{2Q(\mu_a B + N)^2} \quad (17)$$

For the oscillator previously described, we have:

$$\begin{aligned} R_1 &= 400(1 - K) \text{ ohms} = R(1 - K) \\ \text{where } K &= \text{fraction that } R_1 \text{ is lower than } R_2. \\ R_2 &= 400 \text{ ohms} = R \\ R_3 &= R_4 = 2800 \text{ ohms} \\ &= 7R \text{ (for a particular 50 kc bar)} \\ R_6 &= 800 \text{ ohms} = 2R \text{ (very nearly)} \\ R_5 &= 1 \text{ megohm} = 2500 R \\ Q &= 100,000 \\ \mu_a &= 300 \quad \Delta\mu_a = \pm 30 \text{ (10\%)} \\ \theta &= 0 \quad \Delta\theta = \pm 0.1 \text{ radian.} \end{aligned}$$

For the bridge constants we get, in terms of R = resistance of the fixed ratio arm:

$$\begin{aligned} A &= 17500 R^3 K \text{ where } R_1 = R(1 - K) \\ B &= 17500 R^3 \\ M &= 150240 R^3 \\ N &= 70217 R^3 \end{aligned}$$

Inserting these in (16) and (17), we obtain:

$$\begin{aligned} \left. \frac{df}{f_0} \right|_{\theta} &= \frac{4.29 d\theta}{Q(\mu_a + 4.01)} \\ &= +1.5 \times 10^{-8} \end{aligned} \quad (18)$$

for

$$\Delta\theta = +0.1 \text{ and } \mu_a = 300 - 30 = 270,$$

and

$$\begin{aligned} \left. \frac{df}{f_0} \right|_{\mu_a} &= - \frac{4.29 \theta d\mu_a}{Q(\mu_a + 4.01)^2} \\ &= +1.7 \times 10^{-9} \end{aligned} \quad (19)$$

for $\Delta\mu_a = -30$ and $\theta = +0.1$; this reduces to zero if there is no amplifier phase-shift ($\theta = 0$).



These results indicate that a change of amplification of 10% produces only 2 parts in a billion change in frequency in the presence of an amplifier phase-shift of 0.1 radian (6 degrees). If the amplifier phase-shift is made smaller, this change is correspondingly reduced. Changes in phase of the amplifier will produce 2 parts in one hundred million change in frequency, for a phase variation of 0.1 radian (6 degrees), with the

gain at its subnormal value of 270.

The results also show that for best performance the phase-shift of the amplifier should be kept as small as possible, the gain should be high, and the Q of the crystal should be high. The values given above are readily realizable in practice; in fact, substantially higher gains and higher Q 's can easily be obtained, resulting in better performance than that outlined here.

— J. K. CLAPP

COVER PHOTOGRAPH

● **NUCLEAR DISINTEGRATION**, or atom-smashing, has been made possible largely by the development of the cyclotron, which produces high-velocity particles without the use of correspondingly high voltages. The acceleration necessary to produce the desired velocities is imparted to the ions by radio-frequency fields. The production of the radio-frequency energy and its transfer to the dees, or accelerating electrodes, is mainly a radio engineering problem.

The photograph on the front cover, taken at the Massachusetts Institute of Technology, shows the TYPE 821-A

Twin-T Impedance-Measuring Circuit and the TYPE 605-B Standard-Signal Generator being used to measure the admittance of the transmission lines which couple the radio-frequency-oscillator to the dees.

Rapid progress is being made in cyclotron development in many laboratories, and designs are changed frequently. The lines shown in the cover photograph are part of an older model. An interesting paper on the present M. I. T. cyclotron appears in the *Journal of Applied Physics*, Volume 15, Nos. 1 and 2, January and February, 1944.

THE General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in communication-frequency measurement and control problems. When sending requests for subscriptions and address-change notices, please supply the following information: name, company name, company address, type of business company is engaged in, and title of position of individual.

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THE

General Radio EXPERIMENTER

VOLUME XIX Nos. 1 and 2

JUNE and JULY, 1944

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

NEW ENTRANCE—SAME LOCATION

Also IN THIS ISSUE

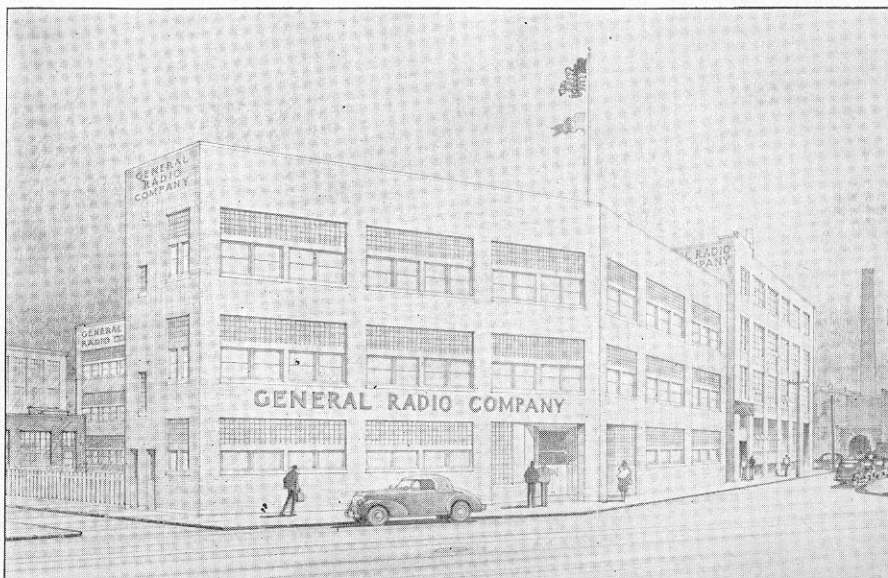
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● VISITORS are now being welcomed at a new front door — 275 Massachusetts Avenue, instead of 30 State Street. This does not mean that we have strayed far afield. In fact, in the twenty nine years of the Company's existence we have had only three addresses and they have all been within a circle not over five hundred feet in diameter. What has taken place is that, as we have grown, the "main entrance" has been moved to suit the changing conditions.

"entrance" has been moved to suit the changing conditions.

Our offices and research laboratories were formerly located in our main factory building. Gradual growth, quite aside from war demands, had made us consider the advisability of constructing a new building devoted only to research laboratories and administrative offices. War expansion made this imperative.

Architect's drawing of front of new building.



Last year a building adjacent to our present factory became available. This building, which adds thirty-eight thousand square feet to our plant, has been completely reconditioned and is now occupied entirely by our laboratories and general offices. The space vacated will be devoted to manufacturing.

War conditions prevent us from extending a general invitation to our friends to inspect these new laboratories, but we hope, as soon as war restrictions are removed, that many of our friends will visit us.

To those familiar with our old location we would call attention to the fact that the new entrance is on the side of our buildings almost diagonally opposite the former State Street entrance. The factory and offices now occupy most of the space in the block bounded by Massachusetts Avenue, Front Street, Windsor Street, State Street, and Village Street.

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In reconditioning the building for office use, considerable attention has been paid to lighting and to ventilation. Liberal use of glass block in addition to normal window space takes full advantage of available daylight, while the interior lighting is fluorescent. A central circulating air system provides adequate ventilation for each individual office. In planning the system, provision was made for adding humidifying and cooling equipment after the war when it again becomes generally available.

Space has been allotted with a view to achieving the most efficient layout. Departments that are functionally related have been placed as near to one another as is possible. Thus the first floor is occupied by the Sales Engineering, Commercial, Service, and Publicity Departments; the second floor contains administrative offices, production office, and drafting room; while the third is devoted entirely to engineering, research, and the model shop.

The new engineering laboratories are

View of the interior of an engineering office designed for two men.





among the more interesting features of the new building. Our development and design engineers have been working under difficult conditions for some time. Scattered over two floors, they were necessarily grouped according to the problems on which they were working, with a resultant overcrowding of some of the laboratories. In the new quarters, adequate space has been provided, and the size of each room depends upon the number of men scheduled to occupy it.

Bench and desk space is provided for each occupant.

Grouping the entire department on a single floor permits a more efficient use of available equipment and facilitates consultation between individuals and between groups. An engineering library, a standards room, and a model shop complete the present facilities. Later, when equipment becomes available, an instrument and stock room will be added.

MEETING THE WARTIME DEMAND

● **IN WAR** as in peace, the test equipment maker provides some of the vital tools of engineering and production. The nature of his product changes little in wartime; the volume of his production increases tremendously.

Long before Pearl Harbor it became evident to producers that test equipment production far in excess of previous levels would be needed to implement the war production program. Some procurement agencies recognized this; some did not. The production and materials agencies of the government had to be convinced and, because of this, much early production was held up for lack of materials and priority. All this was inevitable, in view of the rapid increase in

national production of direct combat material. The bottlenecks were quickly broken, but test equipment production was delayed and expansion seriously hampered.

The General Radio Company manufactures hundreds of different items. It is the largest manufacturer of most of them and the only manufacturer of many. Their uses cover many kinds of industry — radio, telephone, telegraph, chemical, automotive, ordnance, electrical manufacturing, petroleum, textiles, aircraft, and transportation, to name a few.

In wartime, signal generators, vacuum-tube voltmeters, and oscillators are

(Continued on page 4)

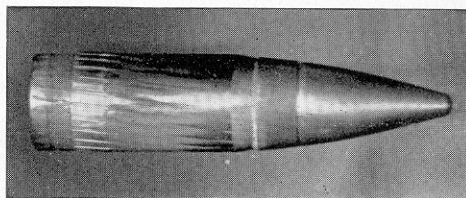
OUR FOURTH ARMY-NAVY "E"

● **ON JUNE 14, 1944**, the General Radio Company was awarded for the fourth time the Army-Navy "E" for excellence in the production of war material. We are proud to be one of the few in our industry to receive the third star.



needed for testing combat radio and radar equipment used in aircraft, in ships, and on the ground; crystal manufacturers need frequency standards and frequency measuring equipment to calibrate the quantities of quartz crystals used in radio equipment; manufacturers of component parts need bridges, as do insulation manufacturers and wire and cable companies to meet stringent Army-Navy test specifications; stroboscopes and vibration meters are used in the design and test of motors, engines, vehicles, and airplanes, and in fundamental research on mechanical and hydraulic problems. All these vital war industries need test equipment in a volume commensurate with their own increased production.

Industrial laboratories, and those of the Army and Navy as well, have expanded to take care of the greatly increased research and development program necessary to produce the many new types of weapons that modern warfare requires. These must be equipped with electronic measuring equipment, as must the military service and repair depots all over the world that keep com-



(Photo courtesy
Inspection Board of United Kingdom and Canada.)

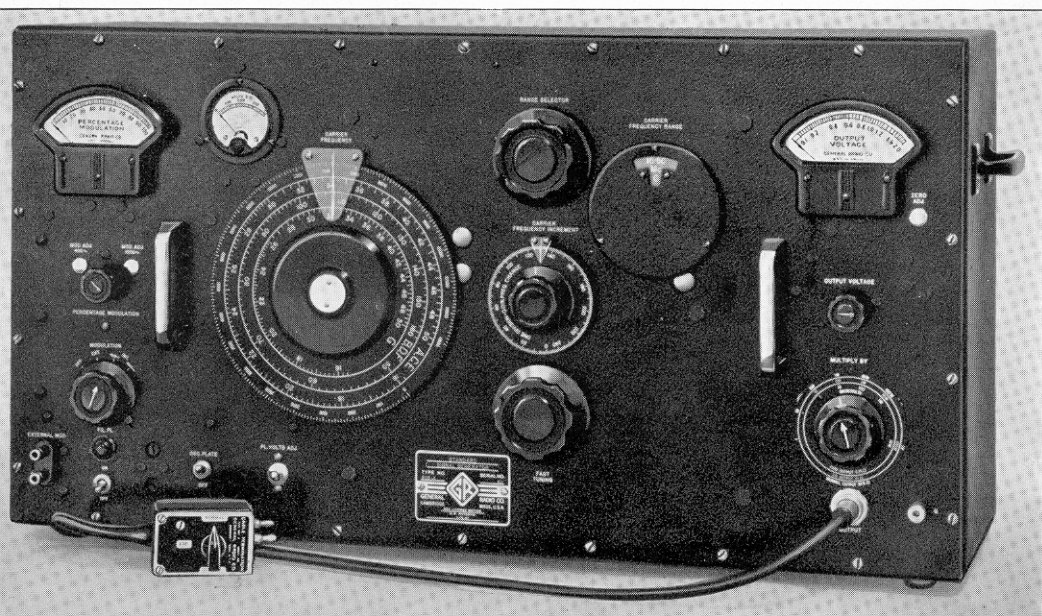
bat equipment in operating condition. High-speed photograph of a projectile in flight, taken with General Radio stroboscopic equipment. The exposure was of the order of 1 or 2 microseconds.

bat equipment in operating condition.

A further portion of General Radio's output goes to the armed forces in the form of specialized equipment for service and combat use. Supplied in moderate quantities in peacetime, these products are now in continuous large-scale production. Their importance cannot be discussed now, but will be an interesting postwar story.

Quality parts have always been an important part of General Radio's output. Wartime emphasis on precision and ruggedness has greatly increased the demand for these. Radio equipment for tanks, planes, and ships requires large

Radio manufacturers use the TYPE 805-A Standard-Signal Generator for testing and calibrating military receivers.





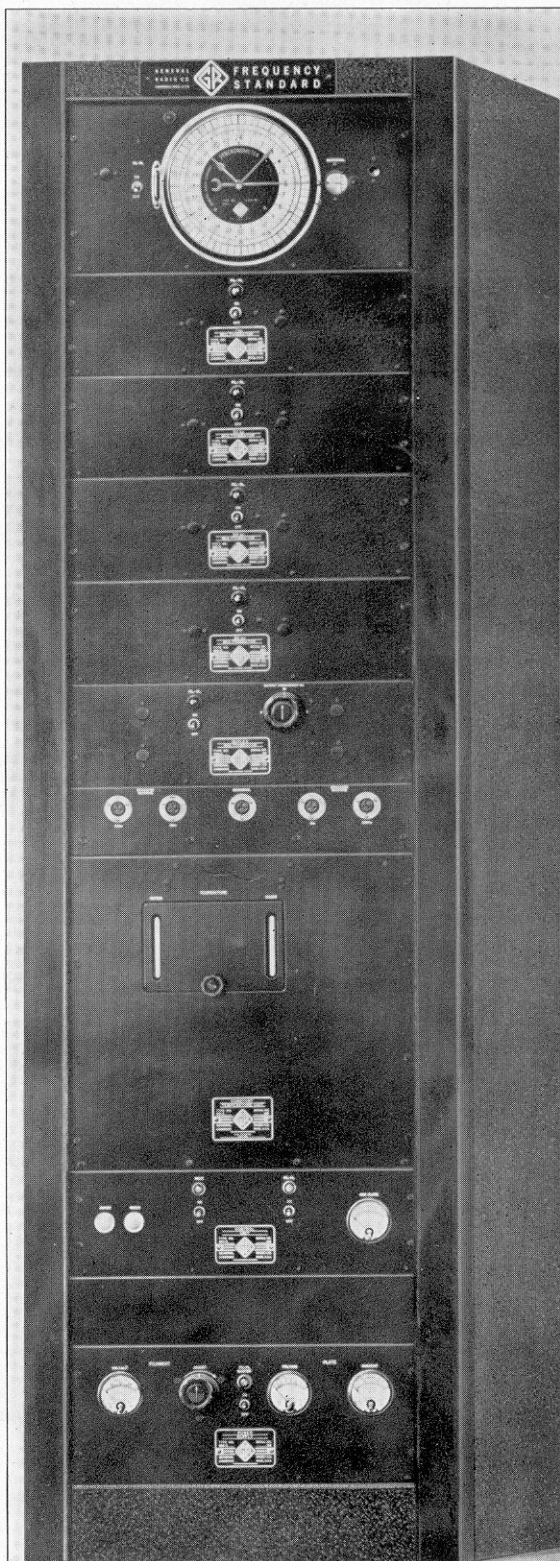
quantities of Variacs, rheostats, precision condensers, knobs and dials, and plugs and jacks — on all of these the rate of production has been stepped up many times.

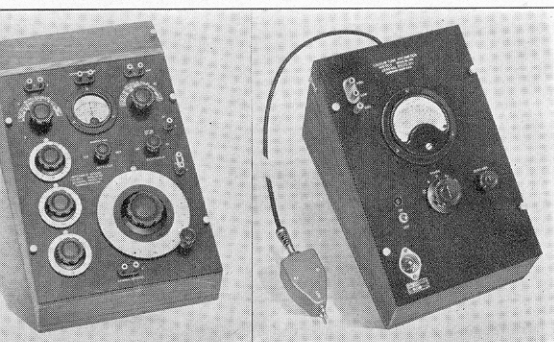
Even before Pearl Harbor, the demand was heavy; we expanded to meet it, and have been expanding ever since. Through greatly increased personnel, the leasing of outside space, and widespread subcontracting, we have been able to meet most of the demand. When quantities ordered reach the point where production facilities are not sufficient in our plant, designs and full manufacturing information are turned over without charge to other manufacturers who have free facilities and are equipped to do the job and to maintain the quality of the product.

Owing to the unprecedented demand, a good part of which was not anticipated at the beginning of the war, there is currently a shortage of some kinds of test equipment, particularly of the higher quality types. All producers of this class of material are heavily loaded but are doing their best to alleviate the shortage. Although the time now required for delivery may be several months, this actually represents a very short production cycle.

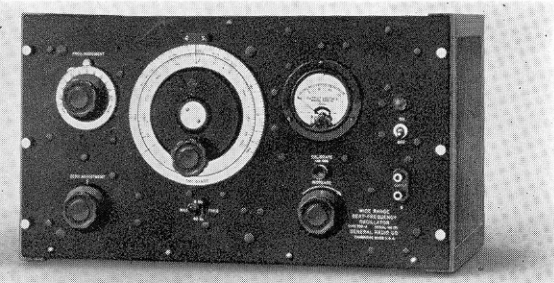
The contribution of our engineering department to the war effort has been important. Many specialized items of test equipment, for which the demand was small but the need urgent, have been designed and produced in our model shop. The experience and background of our engineers have been applied to highly specialized development projects where no manufacture by General Radio beyond a sample or working model was contemplated. In some cases no other manufacturer had the facilities to carry out the contract.

Standards of frequency, like the primary standard shown here, are necessary for the calibration of quartz crystals,

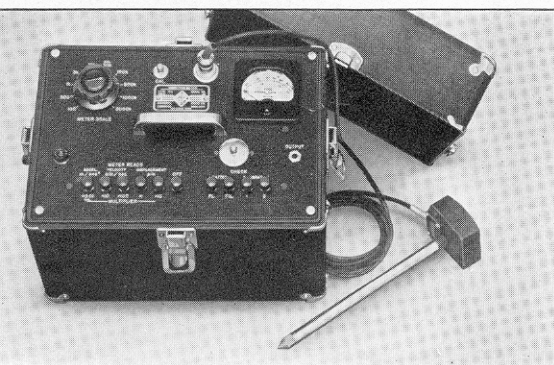




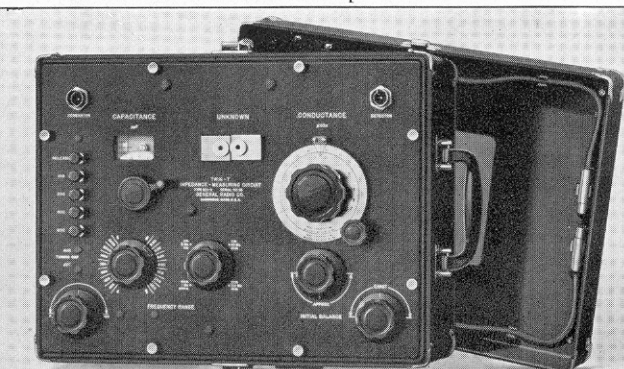
Two instruments used in military and industrial laboratories—left, TYPE 650-A Impedance Bridge and right, TYPE 726-A Vacuum-Tube Voltmeter.



Another important test instrument is the TYPE 700-A Wide-Range Beat-Frequency Oscillator.



The TYPE 761-A Vibration Meter is used by builders of mechanical equipment to measure the magnitude of vibration as a step in its elimination.



Many of these developments have been completed. Of some items only a sample was made, of others a production lot, and of still others complete working drawings were supplied in order that other manufacturers could produce them in quantity.

Among the specialized small quantity items produced by General Radio for war use are receivers, signal generators, pulse generators, stroboscopes, frequency meters, and oscillators. Although these specific types are not generally available now, the experience gained in designing them will eventually mean new and better instruments.

Our engineering department is in close contact with several of the large laboratories engaged in research on new weapons. General Radio engineers have devoted much time to their assistance, sometimes as consultants, occasionally as part-time staff members, always without compensation to the individuals or to the Company. In addition, special parts and assemblies have been made for these laboratories in our model shop.

(Left, below) Cables, transmission lines, and circuit elements are held to rigid specifications for military radio equipment. The TYPE 821-A Twin-T Impedance Measuring Network is used to measure reactance, resistance, power factor, characteristic impedance, and other quantities at radio frequencies.

(Right, below) Both the military services and manufacturers of military equipment use Strobotac for testing, for speed measurement, and for tachometer calibration.





Company executives have given freely of their time to the supervision of outside war research problems and in serving on advisory committees set up by war agencies.

In spite of substitute materials and inexperienced personnel, the general level of quality of our instruments has been maintained. Service difficulties have been held at a minimum. Wartime use puts a severe strain on equipment designed for peacetime service. This problem was recognized at an early date, and facilities were set up for handling an un-

precedented volume of repairs and service correspondence. Service manuals were made available, repair parts were stocked, and repair facilities were expanded. Seldom do urgent repair jobs take more than one or two weeks in our plant, unless delayed by parts or repair work which must be done outside our plant by the original manufacturer. Hundreds of repairs each month go through our Service Department, but the proportion of repairs to total output is being steadily reduced.

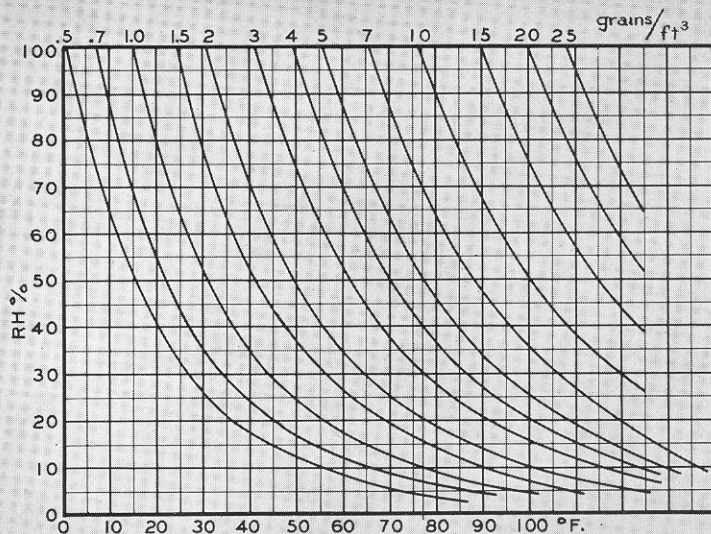
HEAT CAN CONQUER HUMIDITY

● **WE HAVE BEEN TAKEN** somewhat to task for the pessimistic note on which the article on "The Effect of Humidity on Electrical Measurements," reprinted in the April General Radio *Experimenter*, ended, when it stated that "little can be done to eliminate this type of error (from humidity), short of air conditioning." Two palliatives were, however, mentioned, both applicable to the measuring equipment, namely, placing a desiccant or a source of heat inside the instrument. The former decreases the relative humidity, and also the absolute humidity, by removing the moisture, while the latter decreases only the rel-

ative humidity, leaving the total moisture content unchanged.

While the addition of heat has been recommended for preventing moisture condensation when in a fluctuating ambient the measuring instrument is below the dew point, the effectiveness of a rise in temperature of 10 to 20 degrees in reducing relative humidity has not been appreciated. The change in relative humidity with temperature, assuming that the weight of water in a unit volume remains constant, is shown in Figure 1, first shown in the article on "Relative Humidity at Boston" in last month's General Radio *Experimenter* and ampli-

FIGURE 1. Curves showing the relation between relative humidity and temperature for various densities of aqueous vapor expressed in grains per cubic foot. Plotted from Table XII, Psychrometric Tables, by C. F. Marvin, Weather Bureau, U. S. Department of Agriculture, Government Printing Office, Washington, D. C., 1937.



fied here to apply to the higher temperatures. At high relative humidities only a moderate rise in temperature is required to decrease the relative humidity by a factor of two. Suppose that ambient conditions are 91% RH at 80°F. Then entering the chart at this point and following down the curve of 10 grains per cubic foot, a rise of 18°F. (10°C.) decreases the relative humidity by 37% RH to 54% RH, while a rise of 36°F. (20°C.) produces a drop of 58% RH, to 33% RH. Since the curves of constant water content have about the same shape over a considerable temperature range, these figures are representative. Thus the application of enough heat to raise the temperature of the air inside an instrument 20°C. above ambient will reduce even 100% RH to the safe value of 36% RH. Since the safe temperature rise of component parts of an instrument is usually 40°C., a rise of 20°C. will be perfectly safe and will not impair accuracy.

Considerable care should be taken in choosing the source of heat. The heating element should be completely shielded with a grounded shield to prevent 60-cycle induced voltage from appearing in the measuring circuit, which, unless filtered out by a selective detector, would produce errors when the measurements are being made at 60 cycles.

The heater should also operate at a low temperature, preferably not much more than 40°C. above its surroundings.

A high temperature source, such as a lamp, will heat surrounding objects directly by radiation, while what is wanted is that the air itself be heated. The addition of a fan to circulate the air and to direct it onto the heater might be desirable in a large instrument. Heater and fan could be combined, as in a hair drier. The fan motor must be of the induction type, without a commutator, to prevent radio interference.

The power needed to maintain the instrument and its contained air at an elevated temperature depends mainly on the area and surface condition of the instrument case. A value of 0.04 watts per square inch of exposed surface will raise that surface about 10°C. above ambient and provide an inside air temperature 10°C. higher both for a rack mounted instrument with a nickel-plated dust cover and for a copper-lined walnut cabinet. The total power for the TYPE 716-B Capacitance Bridge with a surface area of 1126 square inches is 45 watts, and for the TYPE 722-D Precision Condenser with an area of 425 square inches is 17 watts. It will be somewhat difficult to prevent the heater from rising more than 40°C. above its surroundings, for it can dissipate only about 0.3 watts per square inch of surface for that temperature difference, or 145 sq. in. for 45 watts and 55 sq. in. for 17 watts. Forced circulation would be very desirable for the larger heater.

— ROBERT F. FIELD

NOTE: In the article by Mr. Field entitled "Relative Humidity at Boston," in the May issue, the figure numbers for FIGURE 3 and FIGURE 4 were transposed.

—EDITOR

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NOTES ON THE DESIGN OF TEMPERATURE CONTROL UNITS

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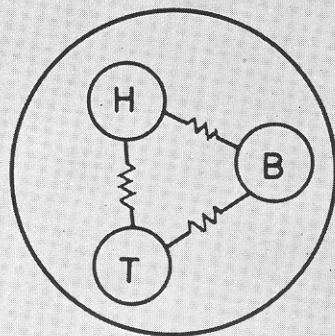
● THESE NOTES deal with a type of temperature control unit often used to control the temperature of a quartz bar or plate, for use as a frequency standard or frequency monitor. The general principles discussed are applicable to other types of units if account is taken of differences in the thermal properties of the materials or medium involved.

Referring to Figure 1, the conditions of the problem can be visualized as three objects, a heater *H*, a controlled space within a metal box *B*, and a thermostat *T*, all enclosed in some form of thermally insulating box. These three objects each have heat capacities; they are interconnected by thermal impedances, along which drops in temperature occur, dependent on the rate of heat flow and on the values of the thermal impedances.¹ Each of the three objects is also connected to the ambient, surrounding the entire unit, by a thermal impedance.

To reduce the fluctuations in average heat flow, the magnitude of which depends on the sensitivity of the thermostat and thermal impedance between the thermostat and heater, it is evident that the thermostat should be as sensitive as possible and the thermal impedance between heater and thermostat should be low.

¹O. M. Hovgaard, "Application of Quartz Plates to Radio Transmitters," *Proc. I. R. E.*, Vol. 20, No. 5, May, 1932, p. 767.

FIGURE 1. Schematic representation of temperature control system composed of a heater, *H*, a metal box, *B*, and a thermostat, *T*, in an insulating container.



To reduce the fluctuations in temperature in the controlled space, the impedance between the controlled space and the thermostat should appear as a low-pass thermal filter, to remove the variations caused by the operating temperature differential of the thermostat. In an elementary form, the metal wall of the box serves this purpose. The filtering action can be increased to almost any desired degree through the use of multi-layer walls and through grading or tapering the distribution of materials forming the wall.²

The remaining thermal impedances should be as high as possible. That of the heater to the ambient determines the power loss, and the cost of operation, of the unit; the initial cost of the insulating container can therefore be fairly high.

The heat capacities of the heater and thermostat should be low, so that small increments in heat energy supplied result in immediate changes in temperature at the thermostat.

Because of practical considerations, such as the use of standard forms of re-

sistors for the heater, mountings for the thermostat, and materials for the unit, the thermal impedances which should be negligibly small can only be brought to effective minimum values. It is the purpose of the following discussion to indicate how the performance can be improved by effectively equalizing temperature drops and by effectively regulating the magnitudes of impedances by electrical rather than mechanical means.³

In order for the thermostat to operate, the amount of heat energy released in the unit, per thermostat cycle, less the energy lost during the input period must be, at least:

$$\Delta H = (W_i - W_o)pq = MS\Delta T \quad (1)$$

where:

W_i = Rate of heat energy input
= Heater watts

W_o = Rate of heat energy output
= Loss watts

p = Thermostat period, seconds

q = Fraction of the period that heat is supplied

pq = "ON" time, seconds

M = Mass associated with thermostat

S = Specific heat of M

and

ΔT = Temperature differential required to operate thermostat.

The heat energy released per thermostat cycle is:

$$H_{in} = W_i pq \text{ watt seconds} \quad (2)$$

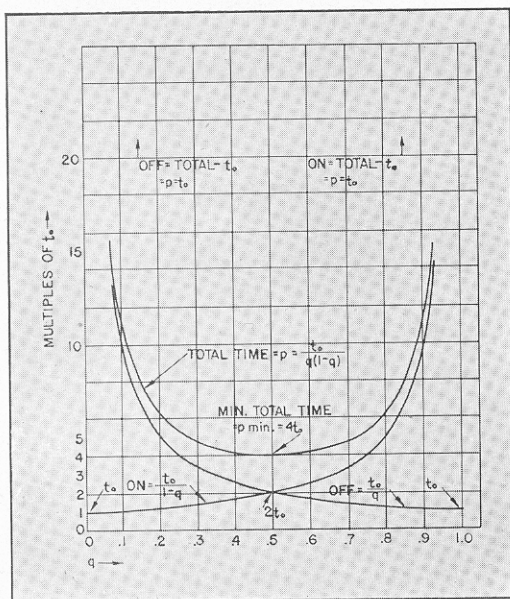
The heat energy lost per thermostat cycle must be the same, or else the temperature of the unit would change:

$$H_{out} = W_o pq \text{ watt seconds} \quad (3)$$

² W. A. Marrison, "Thermostat Design for Frequency Standards," *Proc. I. R. E.*, Vol. 16, No. 7, July, 1928, p. 976. This article contains useful data on thermal properties of common substances likely to be used in constructing units of this description.

³ V. J. Andrew, "The Design of Temperature Control Apparatus for Piezo Oscillators," *Review of Scientific Instruments*, Vol. 3, No. 7, July, 1932, p. 341.

FIGURE 2. Performance of thermostat under conditions of no "overshooting" ($\Delta H = \text{constant}$) and constant heater power ($W_i = \text{constant}$) for varying ambient temperature.





The energy lost during the "ON" time, or the fraction q of the total period, p , is:

$$qH_{\text{out}} = W_i pq^2 \text{ watt seconds} \quad (4)$$

The net heat energy input, per thermostat cycle, is then

$$\Delta H = W_i pq - W_i pq^2 \text{ watt sec.} \quad (5)$$

$$= W_i pq(1 - q) \text{ watt sec.} \quad (6)$$

$$= (W_i - W_0)pq \quad (1)$$

From the last two equations:

$$q = \frac{W_0}{W_i} \quad (7)$$

The loss from the unit can be expressed in terms of its conductivity:

$$W_0 = J(T - T_a) \text{ watts} \quad (8)$$

where:

J = conductivity of the unit to the ambient in watts per degree C.

T = temperature of unit, °C.

T_a = ambient temperature, °C.

For a unit approximately 9 inches cube, with balsa wood walls 1 inch thick, $J = 0.37$ approximately.

The conductivity can be obtained approximately by calculation from $J = KA/d$, where K = conductivity of the wall material, A is the average of inner and outer areas, and d is the thickness of the walls. Usually it is more satisfactory to determine W_0 by observing the operation of the thermostat under known heater power (Equation 7).

Since the thermostat operates between the same "ON" and "OFF" temperatures, ΔT (Equation 1) is constant, and consequently ΔH is constant. This implies that there is no "overshooting," that is, that ΔH is no greater than that necessary to operate the thermostat. We can now predict the operation of the thermostat provided that we determine the heat energy increment, ΔH , and the rate of heat loss of the unit, W_0 .

Our obvious interest would be to determine the performance for constant

power input and varying ambient temperatures. It will be very useful to determine the performance for constant ambient temperature and varying power input for two reasons; first, observations are more easily carried out and, second, "overshooting" is readily disclosed. As will be brought out later, the performance of a unit may be limited as much by "overshooting" as by large changes in ambient temperature.

Case I—Constant Power Input; Varying Ambient

Since ΔH is constant and W_i is constant, we have

$$\Delta H/W_i = pq(1 - q) = t_0, \text{ a constant.} \quad (9)$$

The operation of the thermostat is shown in Figure 2, where the significance of the constant t_0 is evident.

From (9), we can write

$$\text{"ON" time} = pq = t_0(1 - q)$$

$$\text{"OFF" time} = p(1 - q) = t_0/q \quad (10)$$

$$\text{and PERIOD} = p = t_0/q(1 - q)$$

From (9), it is seen that, for an increased input power, t_0 is decreased, and that the thermostat operation is speeded up.

Case II—Constant Ambient; Varying Power Input

In principle, this case is covered by finding a value of t_0 for each value of input power, W_i , from Equation 9. For each value of t_0 , a set of curves like those of Figure 2 is obtained, but each set is to a different scale. On each set of curves a single operating point is found corresponding to the given fixed ambient temperature, that is, to the value of W_0 . The final over-all characteristics would then be the curves connecting these operating points.

Practically, it is not necessary to plot the curves corresponding to each value of t_0 . Instead, the operating points can

be found as in the first case, except that there are different values of t_0 for each point. For the fixed ambient temperature, W_0 is constant. Therefore, since the energy increment ΔH is constant, the heat energy lost during the "OFF" time is constant, or

$$W_0 p(1 - q) = \text{constant} \quad (11)$$

$$\begin{aligned} \text{the "OFF" time} &= p(1 - q) \\ &= \text{constant} \end{aligned} \quad (12)$$

and $\Delta H = (W_i - W_0)pq = \text{constant}$ from which it is evident that, as W_i is increased, the "ON" time, pq , must decrease.

It is important to note Equation 12; if, as the input power is increased, the "OFF" time increases, it is positive evidence of "overshooting."

Results for this second case are shown in Figure 3, on the basis that $\Delta H = 95$ watt seconds and $W_0 = 13$ watts. As the power input is increased, note that the period decreases and that the "OFF" time remains constant.

In Figure 4 are shown the performance curves of a given temperature control unit. This unit was constructed with the thermostat mounted on the outer face of the box, B, with the heaters disposed around all six faces of the box,

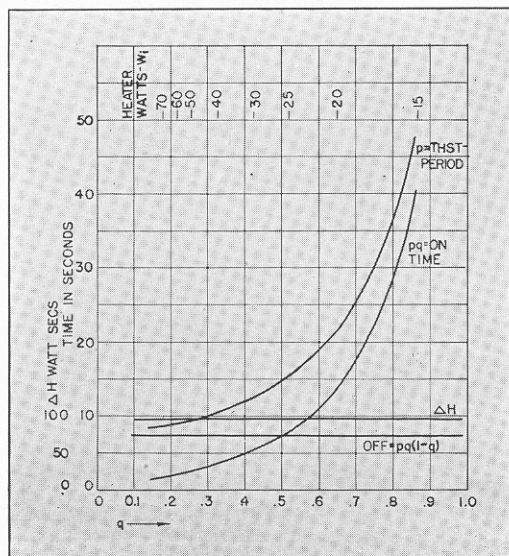
but not in direct thermal contact with the walls of the box.⁴ This construction insures that the average controlled space temperature will not depart appreciably from the thermostat operating temperature. The variations in controlled space temperature, as shown in Figure 4, are not large, and for many purposes would not be troublesome. The remainder of this discussion is devoted to reducing these small variations to the greatest possible degree, without materially altering the mechanical design of the unit.

It is evident that the curves are not of the form expected from Figure 3. The period decreases at first, as the heater power is increased, then passes through a minimum and finally increases again. The "OFF" time increases, with increasing power, following the energy increment curve; the energy increment is not constant. *The average controlled space temperature rises as the heater power is increased.*

This performance is largely due to a high thermal impedance between the heaters and the thermostat. The average controlled space temperature, being a function of power input, will change with any factor causing a change in power input; that is, *either ambient temperature changes or line voltage changes.* It will also be noticed that, because of the delay in temperature rise at the thermostat (causing the energy increment to be too large), the "OFF" time must increase to allow this extra heat energy to escape. The "ON" and "OFF" times are consequently very long.

⁴ J. K. Clapp, "Temperature Control for Frequency Standards," *Proc. I. R. E.*, Vol. 18, No. 12, December, 1930, pp. 2003.

FIGURE 3. Performance of thermostat under conditions of no "overshooting" ($\Delta H = \text{constant}$) and constant ambient temperature for varying heater power, W_i . ΔH assumed as 95 watt seconds; W_0 assumed as 13 watts.





This unit was modified by dismantling the thermostat from the metal box (enclosing the controlled space) and supporting it in such a position that the controlled temperature was about 63.5°C . (instead of 60.0°C ., the thermostat temperature). A thermostat heater, of the lowest practicable thermal capacity, was then wound directly over the thermostat bulb. The thermostat heater power was then adjusted so that the average controlled space temperature was close to 60.0°C . The thermostat heater was connected in parallel with the main heaters.

This arrangement overcomes the high thermal impedance, between the main heater and thermostat, by using a proportional amount of power close to the thermostat. (The significance of the differences in controlled space temperature will be discussed later.)

The performance of this arrangement is shown in Figure 5. Immediately we notice that the energy increment and "OFF" time curves are nearly horizontal, rising slightly only at the highest input powers. The period of the cycle decreases continuously as the input power is increased. All times have been greatly reduced. The average controlled space temperature is constant, within the limits of accuracy of measurement. Comparison of Figure 5 with Figure 3 shows how closely the performance follows the theory.

"Overshooting" has been practically eliminated at all but the highest input powers. It must be borne in mind that, for any given construction, if excessive input power is used, "overshooting" can always be made to occur.

FIGURE 5. Performance curves of a given temperature control unit, modified by reducing the thermal impedance between thermostat and heaters, at constant ambient temperature, for varying heater power, W_i . Compare with Figures 3 and 4.

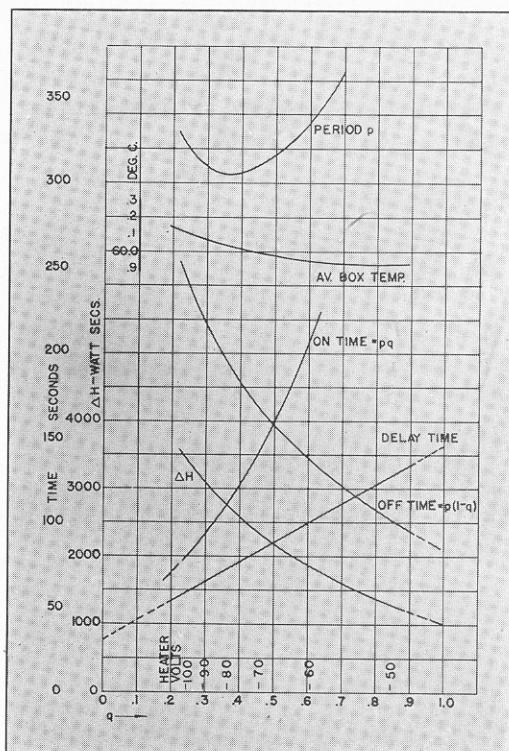
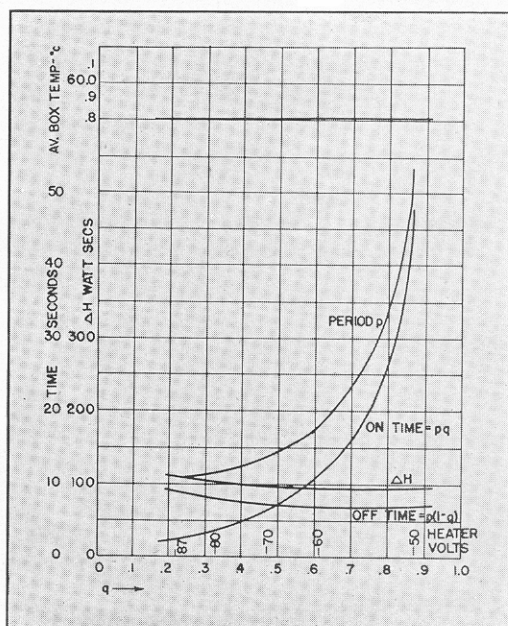


FIGURE 4. Performance curves of a given temperature control unit, at constant ambient temperature, for varying heater power, W_i . Compare with Figure 3.



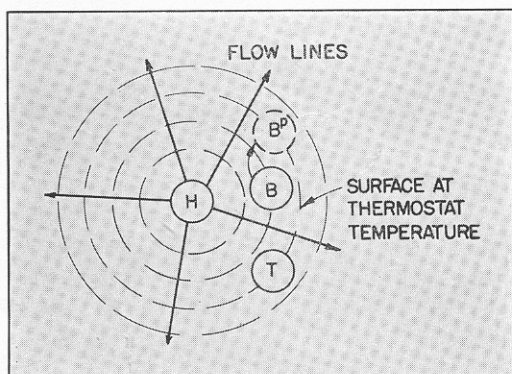


FIGURE 6. Schematic representation of heat flow from a heater, H , and equal temperature shells surrounding the heater.

Having established a means of preventing "overshooting," we next consider the gradient conditions in the unit. Turning to Figure 6, a series of constant temperature surfaces, at every point perpendicular to the lines of heat flow, can be considered as surrounding the heater. (In an actual unit, such surfaces would be very complex, since there are generally several heater units.) The thermostat is on one such surface, corresponding to the operating temperature of the thermostat. If the metal box, B , is inside of, outside of, or on the thermostat surface, the gradient conditions can be represented by Figures 7, 8, and 9 respectively.

In Figure 7a, the box is assumed to be on a constant temperature shell lying

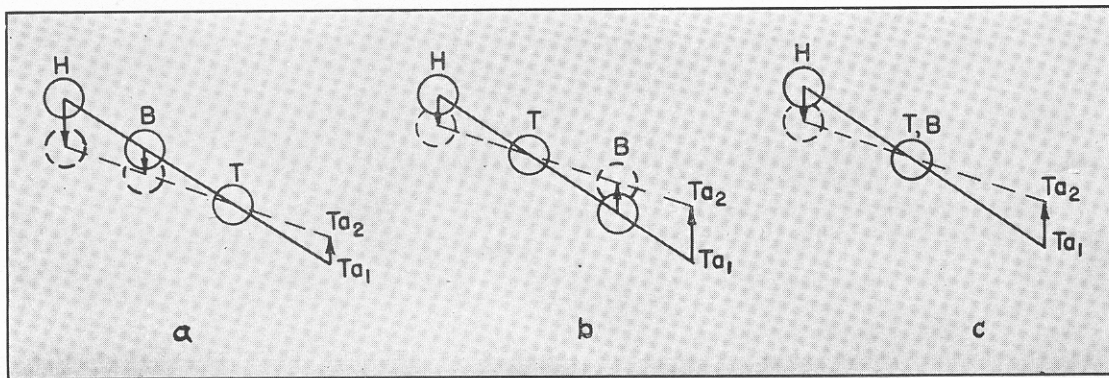
inside of the thermostat shell as at B , Figure 6. At a given ambient temperature, T_{a1} , a certain amount of heat flows through the thermostat shell. The product of this heat flow and the impedance between the shell and ambient is the temperature drop from the shell to the ambient, $T - T_{a1}$. Now if the ambient changes to a higher value T_{a2} , the heat flow must be reduced, since the impedance does not change. The average heater temperature must therefore be reduced, with an accompanying proportional reduction in temperature of the box, B . In this case, as the ambient temperature *rises*, the controlled space temperature *falls*.

Similar considerations apply to Figure 7b, except that, as the ambient *rises*, the controlled space temperature also *rises*.

If the box, B , is brought to the temperature of the thermostat shell as at B' , Figure 6, or vice versa, there should be no change in controlled space temperature with changes in ambient (Figure 7c). This can be accomplished by moving the objects in the assembly relative to each other, or by effectively moving them through the use of a thermostat heater.

Final adjustment of the thermostat heater power is then made on the basis of reducing the changes in controlled space temperature, with ambient changes, to zero. If the thermostat

FIGURE 7 (a, b, c). Schematic representation of gradient conditions, with change of ambient temperature, for various relative positions of heater, H , controlled space, B , and thermostat, T .





heater power is too small, conditions correspond to Figure 7a; if too large, to Figure 7b; and if correct, to Figure 7c. Measurement of the controlled-space temperature for two or more ambient temperatures and for different thermostat heater powers will disclose the power for which $dT/dT_a = 0$. The small readjustment required to bring this about does not appreciably alter the performance shown in Figure 5.

Tests on the unit described in connection with Figure 5 gave the following results:

Thermostat Heater

Power — Watts	dT/dT_a
0.090	+0.018
0.064	+0.0022
0.045	-0.016

Elimination of "overshooting" and proper adjustment of the thermostat heater power reduced the changes in controlled space temperature by nearly 100 to 1. With the thermostat heater power at 0.064 watts, changes in ambient are reduced by 0.0022 or to 1/450th within the controlled space.

—J. K. CLAPP

STANDARD FREQUENCY BROADCASTS

● **A NEW SCHEDULE** for broadcasts of standard frequencies and standard time intervals was announced by the National Bureau of Standards on May 22, 1944.

"This service comprises the broadcasting of standard frequencies and standard time intervals from the Bureau's radio station WWV near Washington, D. C. It is continuous at all times day and night, from 10-kilowatt radio transmitters except on 2500 kilocycles per second where 1 kilowatt is used. The services include: (1) standard radio frequencies, (2) standard time intervals accurately synchronized with basic time signals, (3) standard audio frequencies, (4) standard musical pitch, 440 cycles per second, corresponding to A above middle C.

"The standard frequency broadcast service makes widely available the national standard of frequency, which is of value in scientific and other measurements requiring an accurate frequency. Any desired frequency may be measured in terms of any one of the standard frequencies, either audio or radio. This may be done by the aid of harmonics

and beats, with one or more auxiliary oscillators.

"At least three radio carrier frequencies are on the air at all times, to insure reliable coverage of the United States and other parts of the world. The radio frequencies are:

2.5 megacycles (= 2500 kilocycles
= 2,500,000 cycles) per second,
broadcast from 7:00 P.M. to 9:00
A.M., EWT (2300 to 1300 GMT).

5 megacycles (= 5000 kilocycles
= 5,000,000 cycles) per second,
broadcast continuously day and
night.

10 megacycles (= 10,000 kilocycles
= 10,000,000 cycles) per second,
broadcast continuously day and
night.

15 megacycles (= 15,000 kilocycles
= 15,000,000 cycles) per second,
broadcast from 7:00 A.M. to 7:00
P.M., EWT (1100 to 2300 GMT).

"Two standard audio frequencies, 440 cycles per second and 4000 cycles per second, are broadcast on the radio carrier frequencies. Both are broadcast continuously on 10 and 15 megacycles. Both are on the 5 megacycles in the



daytime, but only the 440 is on the 5 megacycles from 7:00 P.M. to 7:00 A.M., EWT. Only the 440 is on the 2.5 megacycles.

"The 440 cycles per second is the standard musical pitch, A above middle C; the 4000 cycles per second is a useful standard audio frequency for laboratory measurements.

"In addition there is on all carrier frequencies a pulse of 0.005-second duration which occurs at intervals of precisely one second. The pulse consists of 5 cycles, each of 0.001-second duration, and is heard as a faint tick when listening to the broadcast; it provides a useful standard of time interval, for purposes of physical measurements, and may be used as an accurate time signal. On the 59th second of every minute the pulse is omitted.

"The audio frequencies are interrupted precisely on the hour and each five minutes thereafter; after an interval of precisely one minute they are resumed. This one-minute interval is provided in order to give the station announcement and to afford an interval for the checking of radio-frequency measurements free from the presence of the audio frequencies. The announcement is the station call letters (WWV) in telegraphic code (dots and dashes),

except at the hour and half hour when a detailed announcement is given by voice.

"The accuracy of all the frequencies, radio and audio, as transmitted, is better than a part in 10,000,000. Transmission effects in the medium (Doppler effect, etc.) may result at times in slight fluctuations in the audio frequencies as received; the average frequency received is, however, as accurate as that transmitted. The time interval marked by the pulse every second is accurate to 0.00001 second. The 1-minute, 4-minute, and 5-minute intervals, synchronized with the seconds pulses and marked by the beginning or ending of the periods when the audio frequencies are off, are accurate to a part in 10,000,000.

"The beginnings of the periods when the audio frequencies are off are so synchronized with the basic time service of the U. S. Naval Observatory that they mark accurately the hour and the successive 5-minute periods.

"Of the radio frequencies on the air at a given time, the lowest provides service to short distances, and the highest to great distances. Reliable reception is in general possible at all times throughout the United States and the North Atlantic Ocean, and fair reception throughout the world."

THE General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in communication-frequency measurement and control problems. When sending requests for subscriptions and address-change notices, please supply the following information: name, company name, company address, type of business company is engaged in, and title of position of individual.

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A METHOD FOR MEASURING SMALL DIRECT CAPACITANCES

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• A PROBLEM recently encountered in connection with the design of a shielded bridge transformer was that of measuring direct capacitances of the order of $0.01 \mu\text{f}$. The measurement could not be made on available bridges because of the small magnitude of the capacitances and the relatively large terminal admittances associated with them. A parallel-null method, using the TYPE 546 Microvoltage and the TYPE

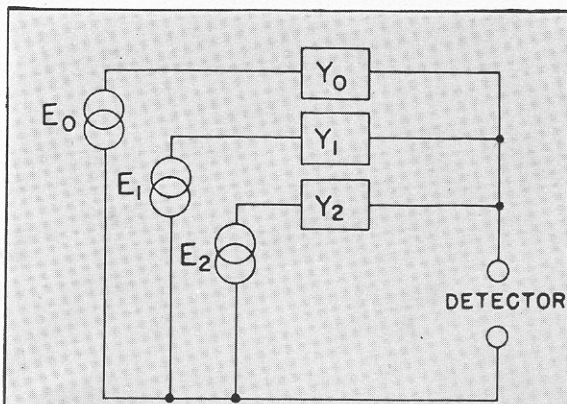
736-A Wave Analyzer, was found satisfactory and is described below.

Theory of the Method

The method is basically a parallel-transmission null method such as is used in the TYPE 561-D Vacuum-Tube Bridge. It resembles both bridge and parallel-T methods, but differs from both in several important respects. It also resembles the so-called "charging-current" methods commonly used for the measurement of interelectrode capacitances, but differs fundamentally from such methods in that it utilizes a null balance rather than a calibrated deflection to determine the unknown admittance.

In Figure 1 are shown three voltages feeding through three separate admittances

FIGURE 1. Basic diagram of the parallel-transmission null method. The relation between the voltages and admittances for zero voltage across the DETECTOR terminals is given in the text.



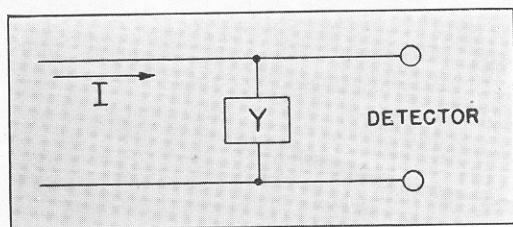


FIGURE 2. The circuit of Figure 1 can, by the application of Thevenin's Theorem, be reduced to the simple system shown.

to a common null detector.

The condition of balance can be derived most easily by reducing the system, as seen at the detector terminals, to a constant current generator shunted by its own internal admittance. The equivalent system is shown in Figure 2.

The total current I is simply the sum of the short-circuit currents produced at the detector by the individual voltages, i.e.:

$$I = E_0 Y_0 + E_1 Y_1 + E_2 Y_2 \\ = \sum E_n Y_n \quad (1)$$

The shunt admittance Y is the admittance at the detector terminals with the voltage sources short-circuited.

$$Y = Y_0 + Y_1 + Y_2 = \sum Y_n \quad (2)$$

If the current $\sum E_n Y_n$ is zero, the voltage across the detector terminals will be zero. Thus the condition of balance is

$$\sum E_n Y_n = E_0 Y_0 + E_1 Y_1 \\ + E_2 Y_2 = 0 \quad (3)$$

If the voltages E_1 and E_2 are 180° out of phase with E_0 , we have

$$Y_0 = \frac{E_1}{E_0} Y_1 + \frac{E_2}{E_0} Y_2 \quad (4)$$

Thus if Y_0 is an admittance $G_x + jB_x$ it can be balanced by adjusting E_1 in series with a susceptance B_1 and E_2 in series with the conductance G_2 . For example, consider the circuit of Figure 3.

The equations of balance for the circuit shown are:

$$C_x = \frac{E_1}{E_0} C_1 \\ G_x = \frac{E_2}{E_0} G_2 \quad (5)$$

Inspection of the circuit and the equations of balance reveals several important features:

(a) Direct admittance is measured, as admittances to ground of the terminals of the unknown are thrown across the generator and detector, where they have no effect on the balance. Likewise, admittance to ground of the leads to the unknown does not affect the measurement.

(b) The two balance conditions are completely independent.

(c) The voltages E_1 and E_2 can be varied, by means of attenuators or voltage dividers, from zero to maximum value. This gives the equivalent of capacitors and resistors variable continuously over extremely wide ranges.

(d) Pure resistance or pure capacitance can be measured equally well by setting E_1 or E_2 to zero as required.

The equations of balance given above are of course derived for idealized conditions which can only be approximated in practice. The factors which have been neglected are:

(a) The dissipation factor of the "standard" capacitor.

(b) The shunt capacitance of the conductance G_2 .

(c) Phase shift between the voltage E_0 and the voltages E_1 and E_2 (at audio frequencies caused largely by the leakage reactance of the transformer).

(d) The internal output impedance of the attenuators supplying E_1 and E_2 .



The effect of these factors is mostly to introduce errors into the measurement of small quadrature components of admittance. For example, entirely erroneous results may be obtained in attempting to measure the *conductance* of a good capacitor, if these factors are not taken into account.

Sensitivity of Balance

The practical usefulness of any null balance method of measurement depends to a large extent on the sensitivity required of the detector, that is, on the magnitude of the change in output voltage for a given unbalance of the measuring circuit. The sensitivity of the system described can be analyzed as follows:

The *open-circuit* voltage appearing across the detector terminals of Figure 2 is given by

$$e' = \frac{I}{Y} = \frac{\sum E_n Y_n}{\sum Y_n} \quad (6)$$

The *change* in open circuit output voltage for a small change in the admittances Y_0 is obtained by differentiation of (6) and can be written as

$$\Delta e' = \frac{E_0 \sum Y_n - \sum E_n Y_n}{(\sum Y_n)^2} \Delta Y_0$$

$$= \frac{\Delta Y_0}{\sum Y_n} \left\{ E_0 - \frac{\sum E_n Y_n}{\sum Y_n} \right\} \quad (7)$$

But at balance $\sum E_n Y_n = 0$ and Equation 7 reduces to

$$\Delta e' = E_0 \frac{\Delta Y_0}{\sum Y_n} \quad (8)$$

The actual voltage appearing across a detector having an admittance Y_d becomes

$$\Delta e = E_0 \frac{\Delta Y_0}{\sum Y_n + Y_d} \quad (9)$$

In a circuit arranged for the measurement of small admittances, the circuit admittance $\sum Y_n$ will normally be small compared to the detector admittance Y_d under which circumstance Equation 9 reduces to

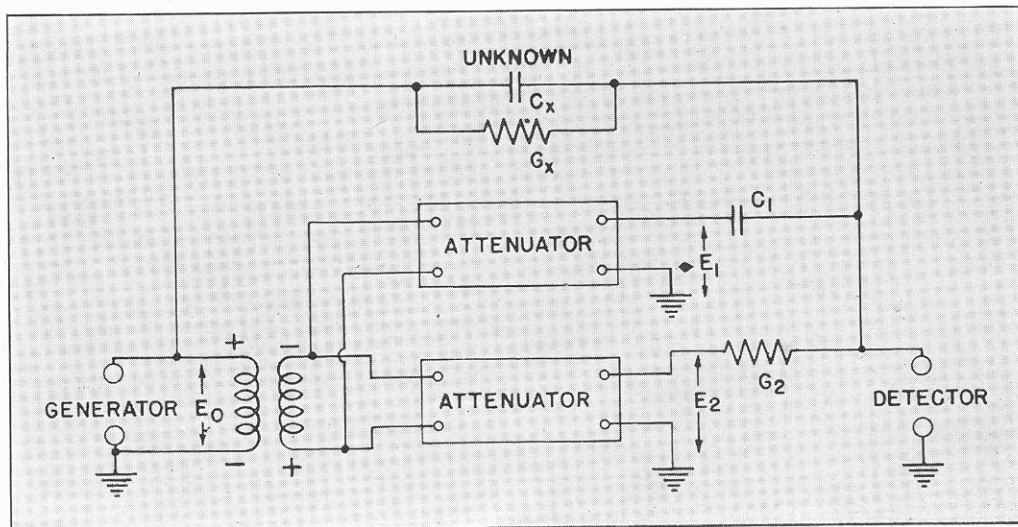
$$\Delta e = E_0 \Delta Y_0 Z_d \quad (10)$$

Equation 10 states that the output voltage for a circuit unbalance ΔY_0 is simply the voltage developed across the detector by the unbalance current.

With particular reference to Figure 3 the minimum capacitance ΔC_x which can be *detected* will be

$$\Delta C_x = \frac{\Delta e}{E_0 \times \omega \times Z_d} \quad (11)$$

FIGURE 3. Showing a practical circuit for the measurement of capacitance and conductance. As indicated, the transformer is connected to produce the phase reversal required for balance.



where E_0 is the applied voltage, Z_d the input impedance of the detector, and Δe the minimum detectable output voltage. Using the TYPE 736-A Wave Analyzer, which has an input impedance of 1 megohm and can easily detect 10 microvolts we have, using an input of 10 volts at 16 kilocycles,

$$\Delta C_x \approx \frac{10 \times 10^{-6}}{10 \times 6.28 \times 16 \times 10^3 \times 10^6} \text{ farads} \\ \approx 10^{-5} \mu\mu\text{f}$$

Even allowing for considerable reduction of the detector input impedance by the shunt capacitance of circuit and leads, there is evidently adequate sensitivity for the measurement of very minute capacitance.

Detector Selectivity

As with any null balance method, selectivity in the detector is necessary in order to obtain the most satisfactory results. In the absence of selectivity the true balance may be masked by power-supply hum and amplifier noise. Also, the presence of harmonics in the signal source may obscure the fundamental balance or produce a false balance, if the conditions for balance depend upon frequency. For the circuit described, frequency does not explicitly enter the equations of balance, but, as a practical

matter, either or both components of a complex admittance will vary with frequency, and the balance for the fundamental will not coincide with the balance for the harmonics.

EXAMPLES OF MEASUREMENTS

Transformer Capacitance

In Figure 4 is shown a schematic diagram of the shielded transformer mentioned earlier. Each winding is carefully shielded, with one side of each winding connected to its shield. In addition, there is a third shield, between the two winding shields, connected to the metal case which serves as a complete outer shield. The capacitance of particular interest is that between the primary lead and the center shield, as indicated in the sketch. In Figure 5 is drawn the reduced circuit showing this capacitance and its associated terminal impedances.

The measurement was made using a TYPE 546-A Microvoltage¹ to adjust the current through the standard capacitor. A sufficiently sharp balance could be obtained in this case without providing a quadrature balance. It was found necessary to surround the transformer with a grounded shield and to use shielded connections in order to reduce the direct capacitance of the measuring circuit to a value small compared to that being measured.²

The standard capacitance used was a 100 $\mu\mu\text{f}$ capacitor of the "postage stamp" variety. With the TYPE 546-A Microvoltage, an input of 7.4 volts was required to produce an output of one volt. Accordingly, the unknown capacitance from Equation 5 is given by

¹This model has a transformer built in. With the current model, TYPE 546-B, an external transformer must be used to obtain the phase reversal.

²If desired, an additional out-of-phase current path can be provided, to cancel out the effect of the initial capacitance.

FIGURE 4. Schematic diagram of doubly-shielded transformer. As this transformer is used, the outer shield is connected to the junction of the ratio arms of a Schering bridge. The capacitance C_T produces an error in dissipation factor reading under certain conditions.

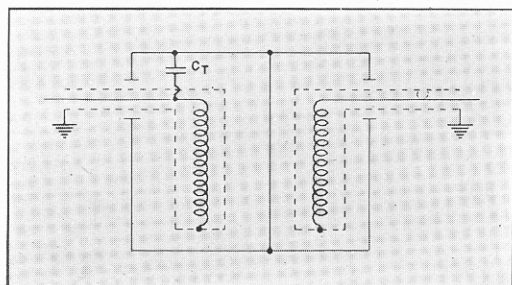
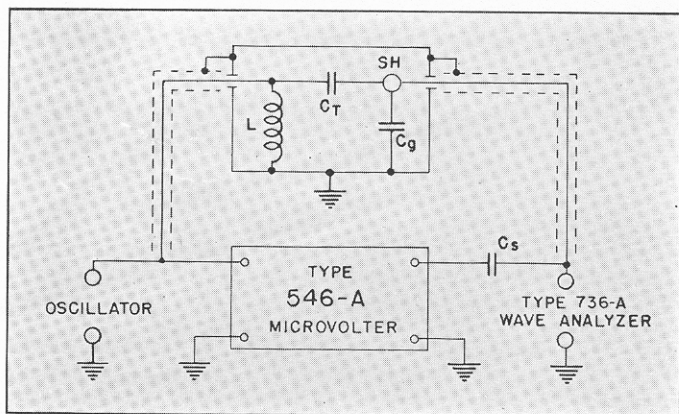




FIGURE 5. The circuit of Figure 4 reduced to show the capacitance C_T and its terminal impedances. The point marked SH represents the outer shield of the transformer. The capacitance C_g is the capacitance of this shield to ground, while L represents the inductance of the primary winding of the transformer. The entire transformer is enclosed in a grounded shield.

Shown also are the connections of the measuring circuit.



$$C_x = E \frac{C_s}{7.4}$$

where E is the reading of the microvoltage dial at balance, in volts. By proper choice of the value of C_s or by adjusting the voltage applied to the microvoltage meter (with a resistive voltage divider), the dial can be made direct reading in μf .

No independent checks of accuracy were available for very small capacitances, but at $5 \mu\text{f}$ (as measured against a TYPE 722-D Precision Condenser) the correct answer was obtained within the limit of accuracy of the calibration of the microvoltage meter dial. Also, the capacitance between primary shield and secondary shield (of the order of $1 \mu\text{f}$) was measured by the method described. An inde-

pendent determination with the transformer installed in the bridge checked within $0.01 \mu\text{f}$.

Interelectrode Capacitance

The type of measurement described obviously bears considerable similarity to the measurement of the interelectrode capacitance of vacuum tubes. A few measurements on representative tubes indicate that satisfactory results can be obtained in most cases. With properly designed, shielded sockets,³ this method should give accurate results for the plate-grid capacitance of screen-grid tubes.

—IVAN G. EASTON

³See, for instance, "The Dependence of Interelectrode Capacitance on Shielding" — Leonard T. Pockman, *Proc. I.R.E.*, Vol. 32, No. 2, February, 1944, pps. 91-98.

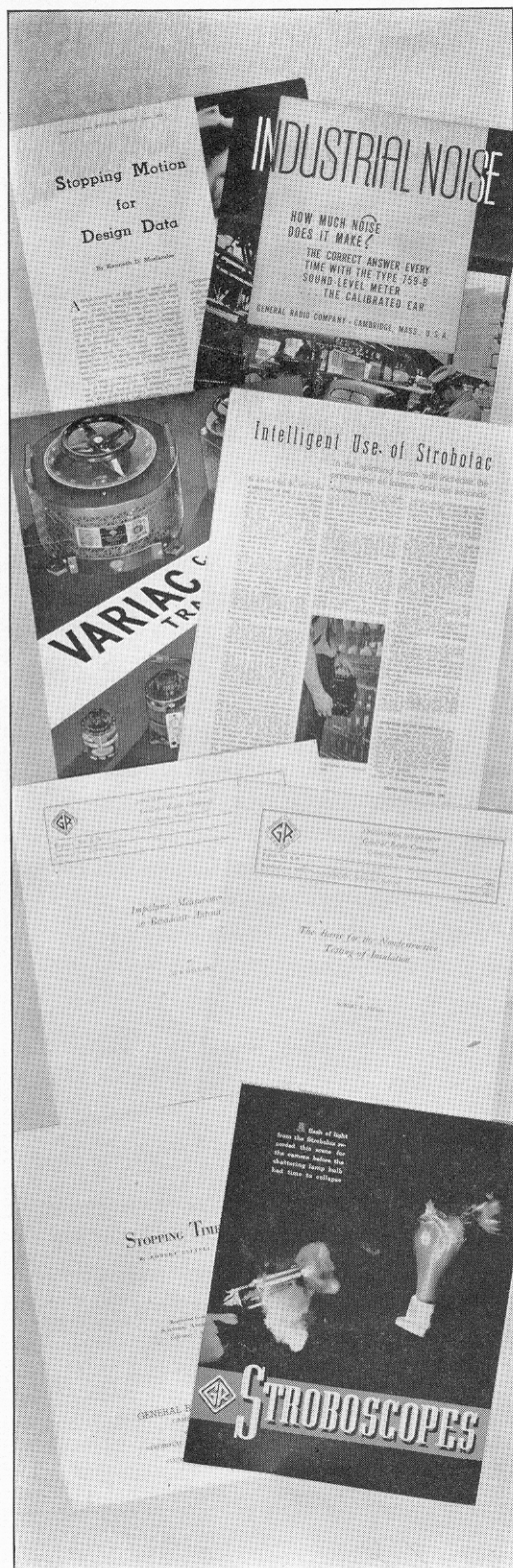
TECHNICAL AND PRODUCT LITERATURE

● **WE HAVE A NUMBER** of publications available that engineers may find useful. These include bulletins describing General Radio products, their applications, and technical literature and charts.

Eyes for Industry—A handbook of stroboscopic techniques. This booklet outlines the principles on which General

Radio stroboscopes operate and describes typical uses. Qualitative and quantitative measurements are discussed, as well as high-speed, single-flash photography.

Stroboscopes—A folder describing the Strobotac and Strobolux and illustrating typical uses. Ask for Form 420-D.



Stopping Motion for Design Data—by Kenneth D. Moslander. Reprint of an article on the use of the stroboscope by design engineers. This article was originally published in *MACHINE DESIGN* for June, 1941.

Intelligent Use of the Strobotac—by Raymond W. Mitchem. Reprint of an article from the October, 1942, issue of *TEXTILE WORLD*. This article discusses the practical application of the Strobotac in the spinning room and shows how it helps increase production without increasing the cost of the finished yarn.

Stopping Time—by Robert Littell. Reprint of an article on stroboscopes from the February, 1940, issue of *THE SCIENTIFIC AMERICAN*. This article was subsequently published in *READERS' DIGEST*.

The Noise Primer—The ABC of noise and vibration measurements. Discusses the characteristics of noise and vibration and the principles underlying their measurement. General Radio instruments for measuring and analyzing noise and vibration are described and details of their use are given.

Industrial Noise—A folder describing the General Radio TYPE 759-B Sound-Level Meter and its use in Noise Measurement.

Variacs—Folder describing the complete line of General Radio Variac autotransformers.

Heat Control Units—by John A. Riddick. Reprinted from *INDUSTRIAL AND ENGINEERING CHEMISTRY*, Analytical Edition, April 15, 1940. Describes a heat control system, using Variacs, on a fractionating column in the chemical engineering laboratory.



Impedance Bridges Assembled from Laboratory Parts—Reprint of a series of articles from the *EXPERIMENTER*. Discusses how audio-frequency impedance bridges can be assembled in the laboratory. Circuits and sources of error are covered, and examples of measurements are given.

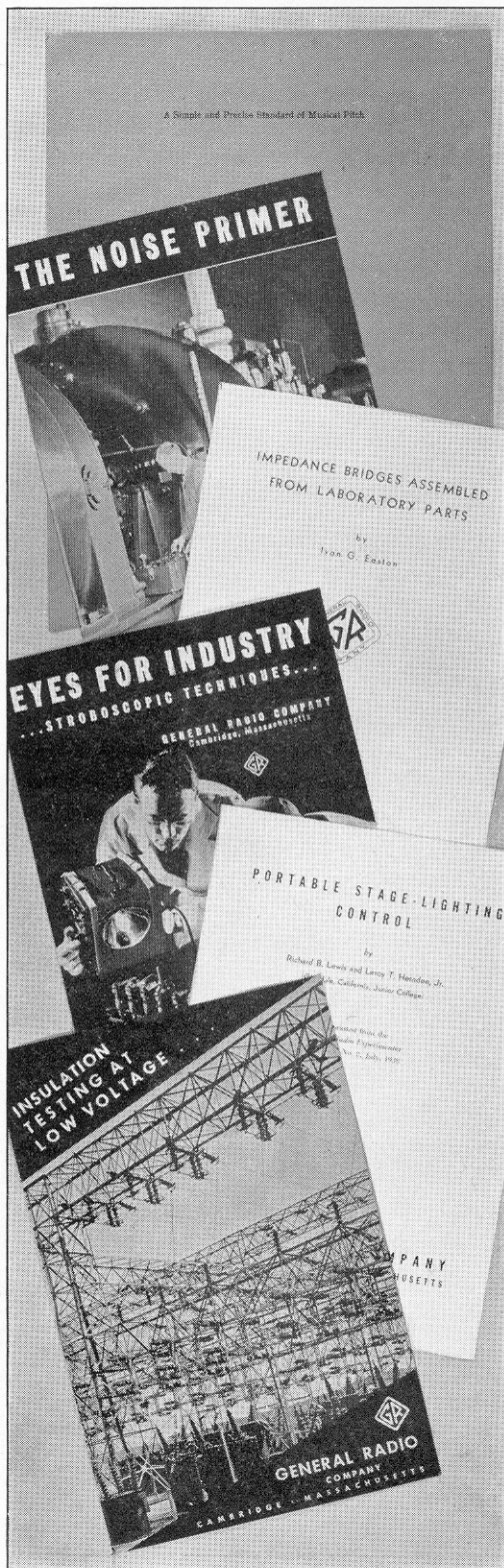
Insulation Testing at Low Voltage—Booklet describing the use of the TYPE 740-BG Capacitance Test Bridge in power factor tests on bushings and insulators.

Portable Stage Lighting Control—by R. B. Lewis and L. T. Herndon, Jr. Reprint of an article from the *EXPERIMENTER*, describing the portable stage lighting equipment, with Variac control, built at the Glendale, California, Junior College.

A Simple and Precise Standard of Musical Pitch—by Horatio W. Lamson. Reprint of an article in the July, 1935, issue of *THE JOURNAL OF THE ACOUSTICAL SOCIETY OF AMERICA*. Describes the use of a 440-cycle electrically-driven tuning fork as a standard of musical pitch.

The Basis for the Non-Destructive Testing of Insulation—by R. F. Field. Reprint of a paper from *TRANSACTIONS OF THE A. I. E. E.* for September, 1941. An analysis of the physical nature of dielectrics, the mechanism of deterioration and breakdown, with an outline of non-destructive methods of measurement.

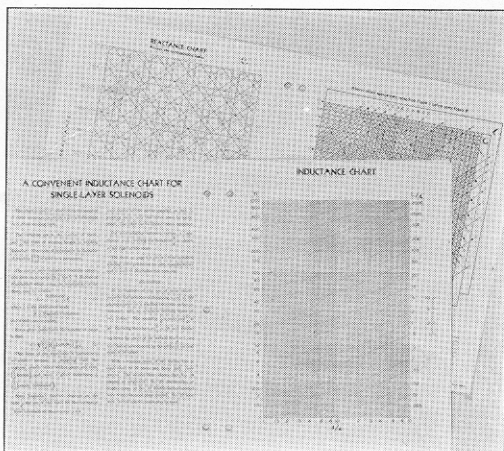
Impedance Measurements on Broadcast Antennas—by D. B. Sinclair. Reprinted from *COMMUNICATIONS*, June and July, 1939. A summary of methods of measuring antennas and lines at broadcast frequencies, with a discussion of sources of error.





Reactance Chart—A wide-range, easy-to-use chart relating reactance, capacitance, inductance, and resonant frequency. Available in $8\frac{1}{2}'' \times 11''$ size, punched for a three-ring binder and also in larger size for wall mounting.

Inductance Chart—A handy chart for determining the number of turns and the size of wire necessary to obtain a given inductance on a given winding form. Available in two sizes, for notebook and wall mounting.



SINCLAIR APPOINTED ASSISTANT CHIEF ENGINEER

Dr. Donald B. Sinclair of the General Radio engineering staff has been appointed Assistant Chief Engineer and will be in charge of circuit development.

Dr. Sinclair was born in Winnipeg, Manitoba, in 1910. He attended the University of Manitoba from 1926 to 1929. He received the degree of S.B. from the Massachusetts Institute of Technology in 1931, S.M. in 1932, and Sc.D. in 1935. Prior to joining the General Radio engineering staff in 1936, he was a Research Assistant at M.I.T. from 1932 to 1935, and a Research Associate in 1935 and 1936. Dr. Sinclair is a member of Sigma Xi and a Fellow of the Institute of Radio Engineers.

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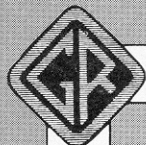
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THE

General Radio EXPERIMENTER



VOLUME XIX No. 5

OCTOBER, 1944

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

THE BUTTERFLY CIRCUIT*

● AT HIGH RADIO FREQUENCIES transmission lines have been used for a long time to perform the functions of resonant circuits. In the region near resonance, the curve of line impedance as a function of frequency has the same shape as the resonance curve of a circuit with lumped constants. The equivalent Q depends upon the frequency, upon the configuration of the line, and upon the materials used. For a concentric line, for instance, with optimum ratio of outside to inside diameter of 3.6, with copper conductors and air dielectric, Q is approximately $0.1D\sqrt{f}$ where D is the inside diameter of the outer conductor in inches and f the frequency in cycles per second. Such a line has approximately 75 ohms characteristic impedance and, for an inside diameter of 1 inch, has a Q of 1000 at 100 Mc. Off resonance, transmission lines are reactances, but they change somewhat differently with frequency than do coils and condensers.

While transmission lines have been used at moderately high frequencies mainly to obtain superior performance as expressed by high values of Q , they have seemed to be the only practical means of tuning at still higher frequencies. It has been assumed that conventional tuning devices were of little use at frequencies over a few hundred megacycles, and so concentric lines have generally been adopted. Theoretically,

*Patent applied for.

FIGURE 1. 55-400 Mc Wavemeter. Adaptation of conventional tuning condenser.

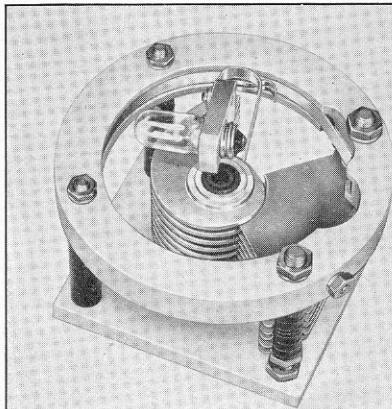
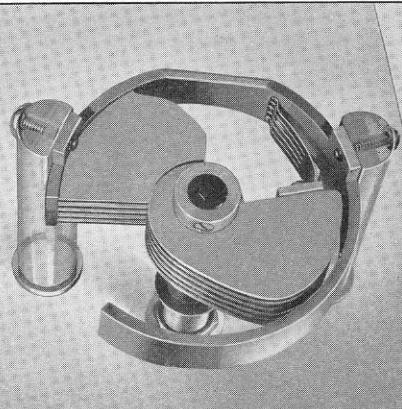


FIGURE 2. Contact type tuning unit for 60-660 Mc.



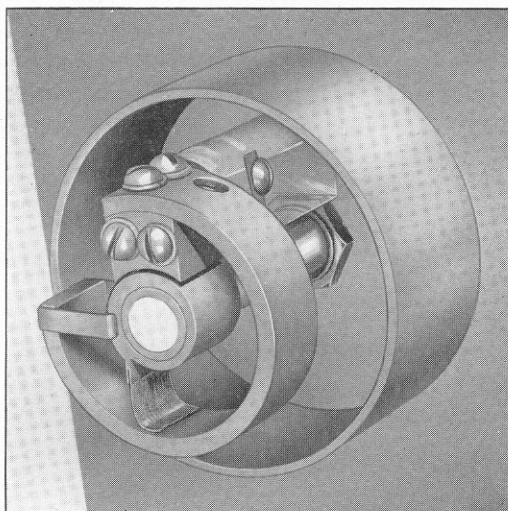
these lines are very suitable for the intended purpose, but in practice their performance has not been completely satisfactory. This is due entirely to mechanical difficulties, but thus far no simple solution to them has been found. The problem involved is merely that of changing the length of a line by accurately known amounts without introducing losses and erratic variations. In some applications the changes required are small, and actual contact with the line conductors can be avoided, but sliding contacts are required if the tuning range is large. In addition, the motion is ordinarily derived from a lead screw of considerable length, which is difficult to manufacture with sufficient accuracy.

It seemed desirable, therefore, to investigate again the possibility of circuits with lumped parameters, where tuning over wide frequency ranges could be achieved by rotating one member with respect to another and to see if sliding contacts could be eliminated.

Contact-Type Tuning Units

The first step of this development can be seen in the wavemeter shown in Figure 1. A conventional tuning condenser

FIGURE 3. 400-1600 Mc. Wavemeter. End of stator band connected to rotor.



and a single-turn inductor consisting of a metal band fastened to a ceramic ring are mounted coaxially. The stator of the condenser is connected to one end of the metal band, and a metal brush on the condenser rotor shaft makes contact with the band. The band covers an arc of 270° , corresponding to the rotation of the condenser, and the length of the band between rotor and stator increases with increase of capacitance. The tuned circuit consists of a capacitance variable from 12 to 85 μmf and an inductance variable from 14 to 99 cm. It covers the range from 55 to 400 Mc. Resonance is indicated by an incandescent lamp or a small glow tube carried on the rotor arm. A transparent housing and a calibrated dial complete the wavemeter. The range could be extended to lower frequencies by adding tuning-condenser capacitance, but the top frequency is limited by the minimum capacitance of the condenser and by the relatively large inductance of the connections to the metal band.

To increase the top frequency obtainable in a circuit of this kind, the tuning condenser must be replaced by a more suitable structure, and the connections to the band must be shortened. The new circuit shown in Figure 2 has little resemblance to the wavemeter of Figure 1, but the basic structure is still the same. The capacitance of this unit, which has an inside band diameter of $2\frac{3}{4}$ inches, varies between 7.4 and 118 μmf and the inductance between 7.8 and 59 cm., giving a frequency range of 60 to 660 Mc. The law of inductance variation is determined by the design, but the frequency distribution can be chosen by shaping the condenser plates. The plates shown in Figure 2 give a straight-line logarithmic frequency scale, which is always desirable for an instrument covering a wide frequency range.

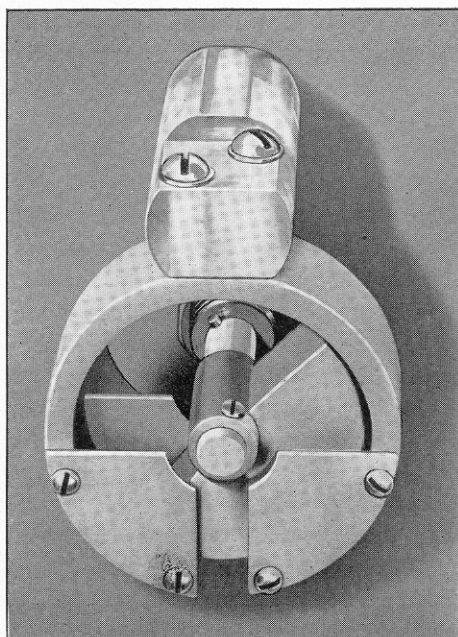


FIGURE 4. Tuning unit without contacts for 400–1200 Mc.

The unused part of the inductance band, projecting beyond the point of contact, is closely coupled to the tuned circuit and resonates at a high frequency. The end of the band is usually connected to the rotor to keep the resonant frequency outside the range of the tuned circuit. This connection can be seen in Figure 3 on a 400 to 1600 Mc wavemeter of somewhat different design, in which the capacitance is varied by an eccentric hub of the rotor.

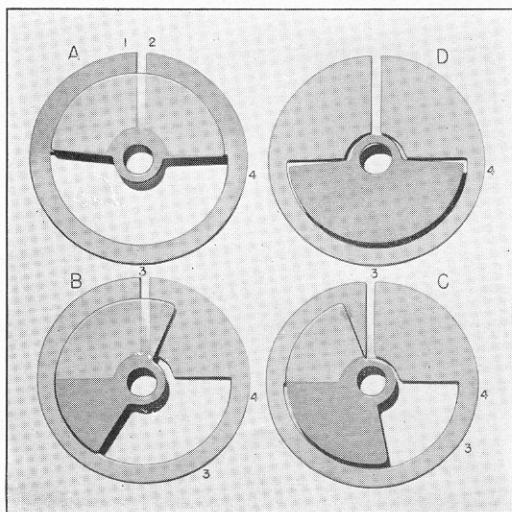
These wide-range circuits are very useful for many purposes, but they still have sliding contacts which require a certain amount of care to insure low losses and trouble-free operation. To avoid these difficulties, the contact spring can be replaced by a capacitive contact member, but suppression of resonance effects by shorting the end of the band to the rotor is then no longer effective. Consequently, the top frequency of the circuit has to be lowered, and, since the bottom frequency

is raised owing to the series capacitance of the contact, the tuning range is reduced considerably.

Tuning Units Without Contacts

A different approach to the problem is shown in Figure 4. In all the circuits discussed so far the highest resonance impedance was developed between a point on the rotor and a point on the stator. In the new circuit, the semi-butterfly circuit, so-called because of its relation to the butterfly circuit shown later in Figure 9, the highest impedance appears between two points on the stator. The plates of the semi-butterfly circuit are shown in four different positions in Figure 5. The stator consists of a semi-circular band connecting two pairs of sectorially shaped plates. To change from the lowest frequency position shown under *A* to the highest frequency position shown under *D*, the rotor is turned through 180°. In position *A* the capacitance between points 1 and 2 on the two stator plates is at maximum. In position *B* the capacitance is almost at minimum and changes only slightly until position *D* is reached. The inductance

FIGURE 5. Stator and rotor plates of semi-butterfly circuit. Capacitance and inductance between points 1 and 2 on the stator are at maximum in *A* and at minimum in *D*.



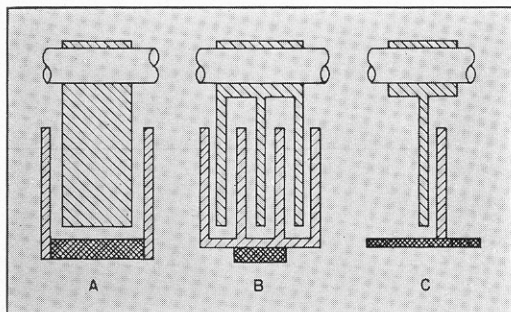
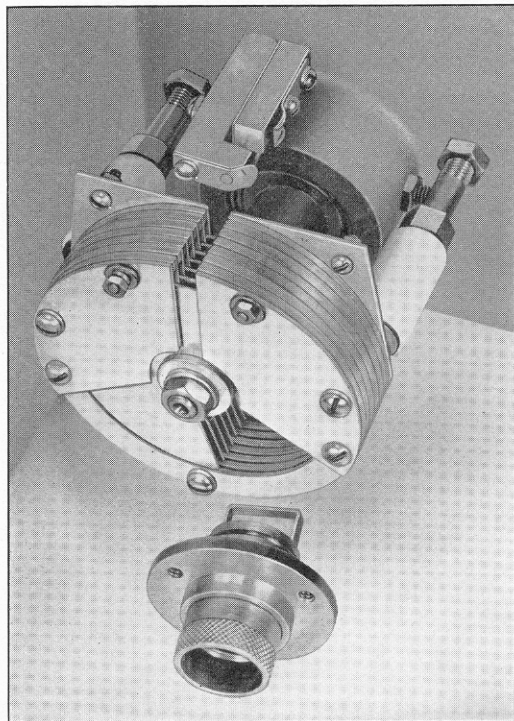


FIGURE 6. Cross section of butterfly circuit with different types of design. The double-cross-hatched section represents the inductance band.

between points 1 and 2 is likewise at maximum in position *A* and changes to minimum over position *B* and *C* to position *D*. The change in inductance of the semicircular band is brought about as the rotor gradually fills up the opening of the inductance loop. In the final position *D*, lines of magnetic flux are restricted to the small clearance between rotor and stator.

FIGURE 7. 100–500 Mc Oscillator. Inductive coupling to the load is varied by rotation of the coupling loop in lower part of the picture.



In this design, wide tuning range with simultaneous change of lumped capacitance and inductance is obtained by rotation of a member that does not require any electrical connections. The rotor shaft can be made of insulating material, and the flow of radio frequency currents through the bearings of this shaft is easily avoided. A cross section of the unit of Figure 4 is shown schematically in Figure 6*A*. The rotor is a solid block. The stator carries a pair of plates on each end of a circular band. The cross section of the band corresponds to the cross-hatched area. The resonant frequency in closed position, corresponding to Figure 5*A*, is changed most effectively by a change in size. Capacitance and inductance increase, and frequency decreases, approximately with the square of the diameter. The resonant frequency in the open position, corresponding to Figure 5*D*, does not vary as rapidly, and the tuning range, therefore, increases almost proportionately to the diameter. At any given diameter the frequency range can be shifted by variations in design. Figure 6*B* shows a unit in which capacitance and inductance have been increased by the addition of plates and by reducing the width of the band. Figure 6*C* is the corresponding high-frequency unit with lowered capacitance and inductance.

The law of inductance variation again is determined by the geometry of the design. In general, the inductance variation will be considerably smaller than in the circuits with sliding contacts and will be further reduced if the rotor plates are shaped to provide a predetermined frequency distribution. The loss of tuning range, the price that has to be paid for a desirable frequency distribution, is, therefore, much higher in the butterfly circuit than in a conventional tuning condenser.



A further modification that yields a lower L/C ratio is used in the 100 to 500 Mc oscillator shown in Figure 7. This circuit uses 120° sections for the stator plates and a 240° section for the rotor. The rotor turns through 120° .

It is important to keep in mind that the tuning devices described so far are two-terminal impedances only. These two terminals are designated 1 and 2 in Figure 5. No point on any of the circuits maintains electrically its position relative to these terminals over the tuning range, in the sense, for instance, of a tap on a coil with respect to the ends of the winding. In Figure 5, the point 3, midway electrically between the terminals, is symmetrically located only in the two extreme positions *A* and *D* and shifts toward terminal 2 for intermediate settings. The potential of the rotor is midway between points 1 and 2 in position *A* and *D* but shifts toward terminal 1 for intermediate settings. This fact is an important consideration in all applications.

Any point of the circuit can be connected to ground, but it is convenient at times not to make any definite connections. This might be desirable in an oscillator circuit, for instance, where plate and grid of a tube are connected to the two terminals and where the cathode is grounded.

The optimum method of coupling a load to the circuit depends upon the ground arrangement. If one terminal is grounded, electrostatic coupling is most convenient. When the circuit is left floating, electromagnetic coupling is more desirable. The only point where effective magnetic coupling over the entire tuning range can be obtained is shown as point 4 in Figure 5. The oscillator shown in Figure 7 uses this type of coupling. By rotating the coupling loop, the amount of coupling can be varied without disturbing the "floating ground" condition of the oscillator.

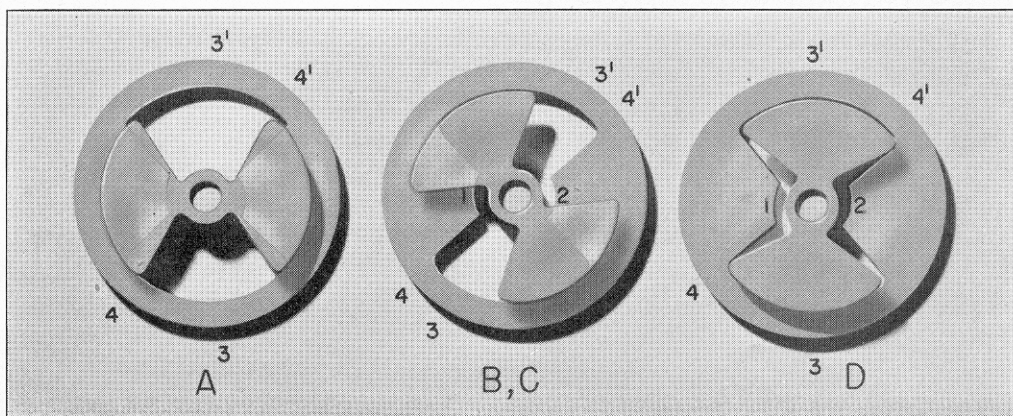
The Butterfly Circuit

A certain amount of electrical symmetry can be obtained by a modification suggested by Dr. A. P. G. Peterson, whereby two circuits are opened up and connected in parallel. The result is the butterfly circuit shown in Figure 8, which received its name from the shape of the rotor.

Figure 8 *A*, *B*, *C*, and *D* correspond to the positions shown in Figure 5. Points 3 and 3' are the electrical midpoints between terminals 1 and 2, and points 4 and 4' are the locations for magnetic coupling.

Electrically the rotor of a butterfly circuit is always the midpoint between terminals 1 and 2, and the rotation between extreme positions is 90° . With

FIGURE 8. Stator and rotor plates of butterfly circuit. Capacitance and inductance between points 1 and 2 on the stator are at maximum in *A* and at minimum in *D*.



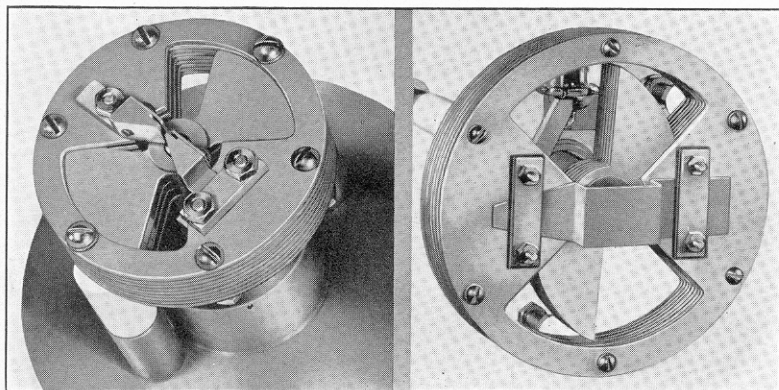


FIG. 9. (Left)
220-1100 Mc
butterfly circuit
with terminals
for TYPE 703-A
door knob type
tube.

FIG. 10. (Right)
100-200 Mc
butterfly circuit
of heterodyne
frequency meter.

these exceptions, the discussion in the preceding paragraphs can be directly applied to the butterfly circuit.

Figure 9 shows a 220 to 1100 Mc oscillator unit intended for use with a TYPE 703-A door-knob type tube. Grid and plate terminals of this tube fit into the spring clips provided on the two stator sections. Figure 10 is the close-up view of the 100 to 200 Mc oscillator of a heterodyne frequency meter. The rotor plates are shaped and provide a logarithmic frequency distribution. The 958 acorn-type oscillator tube can be seen in the rear. Grid and plate are connected to extensions of the two stator sections and the cathode is connected directly to the housing. Figure 11 shows the smallest butterfly circuit built so far, together with an acorn tube and a TYPE 1N22 Detector. This circuit has a tuning range of 900 to 3000 Mc.

Design of Butterfly Circuits

It was mentioned in the introduction that the equivalent Q of a resonant transmission line is of the order of $0.1D\sqrt{f}$. A quarter wave line covering the range from 220 to 1100 Mc would be about $13\frac{1}{2}$ inches long and might have a diameter of 2 inches. The theoretical Q would be 3000 at 220 Mc and 6700 at 1100 Mc. To cover this frequency range, the length of the line would have to be

varied by over 10 inches. The circuit shown in Figure 9 with $2\frac{1}{2}$ inches outside diameter covers exactly this range with a Q of 650 at 220 Mc and of 300 at 1000 Mc.

While the losses in butterfly circuits cannot be made as low as in concentric lines, values of Q can be obtained that are sufficient for many applications. The characteristics of the butterfly circuit shown in Figure 9 are given in Table I, together with the variation in these characteristics that can generally be obtained by varying design factors. In the third column, n is a factor between 1.5 and 3.5, depending on design and size, with the lower values applying for smaller units.

TABLE I

	Circuit of Figure 9		In General	
	Low	High	Low	High
Frequency	220	1100 Mc	1	n^2
Inductance	0.011	0.0041 μ h	1	n^{-1}
Capacitance	48	5 μ mf	1	n^{-3}
Q	650	300	1	n^{-1}
$R = \frac{\omega L}{Q}$	0.023	0.095 Ω	1	n^2
$\sqrt{L/C}$	15.2	28.6 Ω	1	n
$Z = Q\sqrt{L/C}$	9800	8600 Ω	Constant	

The maximum inductance of butterfly circuits in the closed position can be computed by any of the standard inductance formulas by taking $\frac{1}{8}$ of the inductance of the full ring and multiplying by a factor of 1.35 to allow for the



contribution of the stator and rotor plates. For instance,

$$L = 1.35 \times \frac{1}{8} \times$$

$$0.01257r \left(\ln \frac{36r}{t+w} - 2 \right) 10^{-6} \text{ henrys}$$

or

$$L = 2.12r \left(\ln \frac{36r}{t+w} - 2 \right) \text{ centimeters}$$

Another formula suggested by Dr. A. P. G. Peterson that gives fairly accurate results is

$$L = \frac{\pi^2}{6} (r^2 - r_1^2) \times \left[\frac{1}{t+w} + \frac{1}{\sqrt{r^2 - r_1^2}} \right] \text{ centimeters}$$

where r is the inside radius of the band, r_1 is the outside radius of the rotor hub, and t and w the thickness and width of the band, all in centimeters. The minimum inductance depends to a great degree on the clearance between stator and rotor along the circumference of the band and the clearance between the radial plate edges. This latter clearance cannot be reduced very far without adding capacitance and lowering the top frequency. In general, the ratio of maximum to minimum effective inductance will be between 1.5 and 3.5. The highest ratio actually obtained has been 4.6. This circuit had a band diameter of $5\frac{1}{4}$ inches and $1/16$ inch clearance to the rotor. In this particular case the capacitance ratio of the 14-plate unit was 22 — the frequency range 47–470 Mc.

Maximum and minimum capacitance of a butterfly circuit can be computed as in a variable air condenser, but, owing to the butterfly shape, capacitance ratios are considerably less than in well-designed conventional tuning condensers. Figure 12 shows inductance, capacitance, and frequency plotted against dial rotation for the 220 to 1100 Mc butterfly circuit of Figure 9.

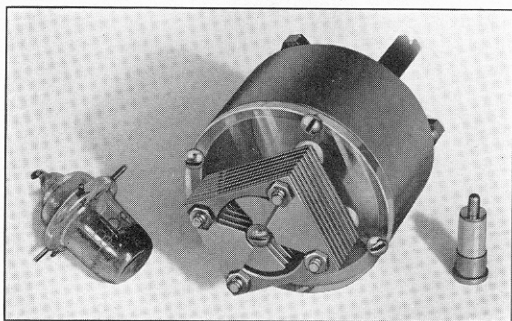
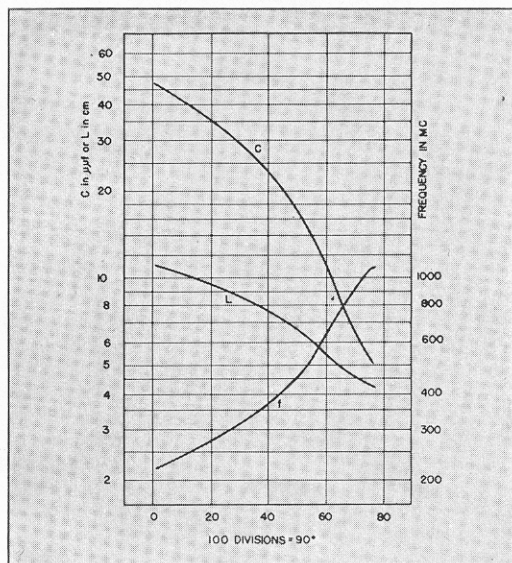


FIGURE 11. 900–3000 Mc butterfly circuit. Acorn tube and 1N22 detector are shown for size.

One effect of changing the number of plates deserves special mention, although it has not been fully explored. As expected, the ratio of maximum to minimum capacitance increases with the addition of plates and the inductance of the stator decreases due to its increased width (designs where the band differs in width from the stator, as shown in Figure 6B and Figure 6C have only rarely been used in practice). For very low numbers of plates, the inductance can decrease more rapidly than the minimum capacitance increases, and a butterfly circuit with four plates, for instance, may have a lower low- and a

FIGURE 12. L, C and f variation of unit shown in Figure 9.



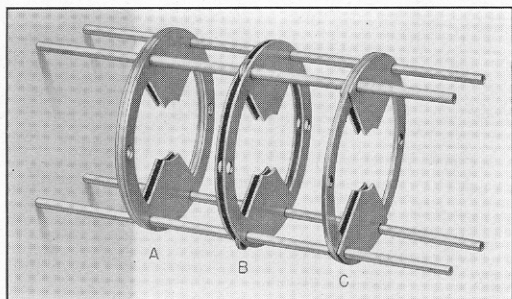


FIGURE 13. Assembly of butterfly stators.

higher high-frequency limit than with two plates. As further plates are added, the range is shifted to lower frequencies and the span slightly increased, but soon odd modes of resonance appear that were not previously observed. These modes are not always noticed in oscillator circuits where a single dominant frequency is produced, but they are easily found by other methods. It appears that these modes can be suppressed or at least can be shifted by additional shorting connections between the individual plates of the four groups of plates. Two connections can be applied at the inside of the stator and two at the outside of the rotor on the radial edges of the plates where they do not interfere with the rotation of the rotor. In practice, shorting of the plates can be avoided by using a larger unit with fewer plates instead of a small unit that requires too many plates to cover the desired frequency range.

The equivalent series resistance of a butterfly circuit increases linearly with frequency and reduces Q at the high-frequency end, where the Q of a transmission line is increased. This increase in equivalent resistance is due to the fact that the change of inductance in the butterfly circuit is brought about by

eddy currents in the rotor and not by a reduction of conductor length. Additional losses originate at the junction of the inductive and capacitive parts of the circuit where currents pass from one plate to another in a direction perpendicular to the plates. Three methods of stator construction with individual plates and spacers are shown in Figure 13. From an electrical point of view, Method A appears to be the best. Method B has the advantage that the rotor can be observed through the slots in the stator and can be centered after assembly. Method C may be desirable in experimental work since it uses parts that can be obtained more easily without special tools. For lowest losses stator and rotor should be soldered after assembly and plated with a material of high conductivity. Dielectric losses can be almost completely avoided in butterfly circuits, if two supports are used, placed at the two points of equal potential corresponding to 3 and 3' in Figure 8A. In practice, however, the supports are often placed at other points, more desirable for mechanical reasons, and on large units three supports may be required. It is interesting to note that, in the circuit shown in Figure 9, about $\frac{1}{4}$ of the 0.023Ω equivalent series resistance, measured at 220 Mc, is accounted for by the metallic losses in the silver-plated inductance band.

The resonant impedance of butterfly circuits is well within the range of plate resistance of vacuum tubes and changes very little from the low to the high frequency end. This makes it possible to produce uniform oscillations over the tuning range.

— EDUARD KARPLUS

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THE

General Radio EXPERIMENTER



VOLUME XIX No. 6

NOVEMBER, 1944

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A 500-MC OSCILLATOR

Also IN THIS ISSUE

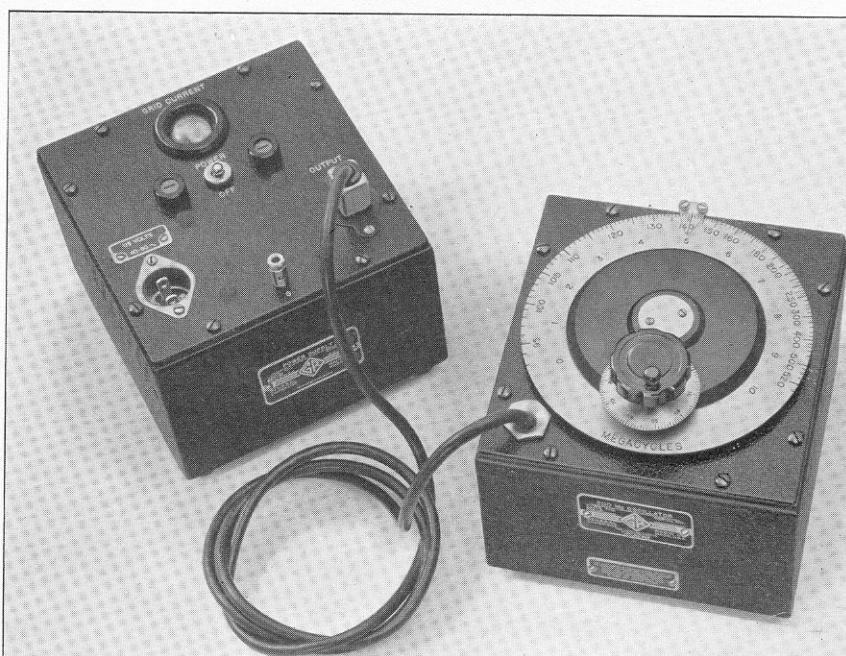
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● A NEW TYPE 857-A OSCILLATOR replaces the TYPE 757-A U-H-F Oscillator announced in August, 1941. Both oscillators use the same tube, but, while frequency range and power are very much the same, the circuits used are radically different. The new oscillator is only one-half the size and weight of the older type and should give much better performance in continued use, since all sliding contacts have been eliminated through the use of a butterfly circuit as the frequency determining element. This radically new type of tuned

contacts have been eliminated through the use of a butterfly circuit as the frequency determining element. This radically new type of tuned

FIGURE 1. View of the TYPE 857-A Oscillator with TYPE 857-P1 Power Supply.



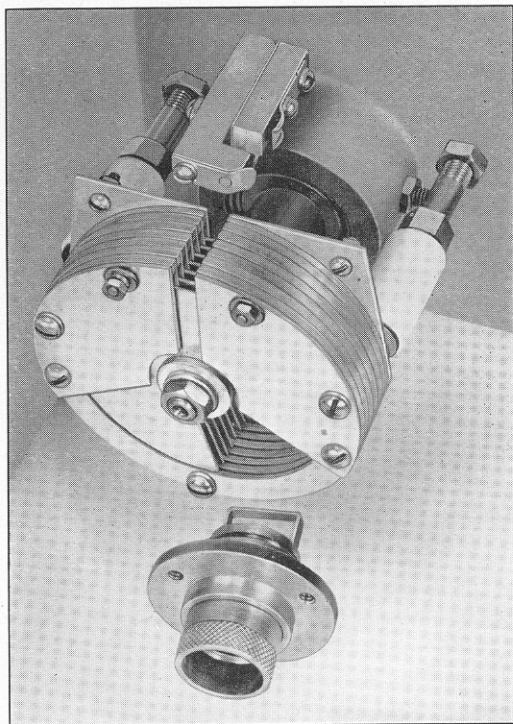


FIGURE 2. View of the tuned circuit of the TYPE 857-A Oscillator, showing also the output coupling loop and output jack in approximately their normal position.

circuit was discussed in detail in a previous issue of the *Experimenter*.*

The TYPE 857-A Oscillator together with the TYPE 857-P1 Power Supply are shown in Figure 1. The oscillator is intended for general laboratory use wherever a convenient source of radio-frequency power between 100 Mc and 500 Mc is required. The oscillator is contained in a metal housing that provides a moderate amount of shielding. Output is obtained at a coaxial jack at the side of the cabinet. Output coupling is inductive and can be varied continuously from maximum to practically zero.

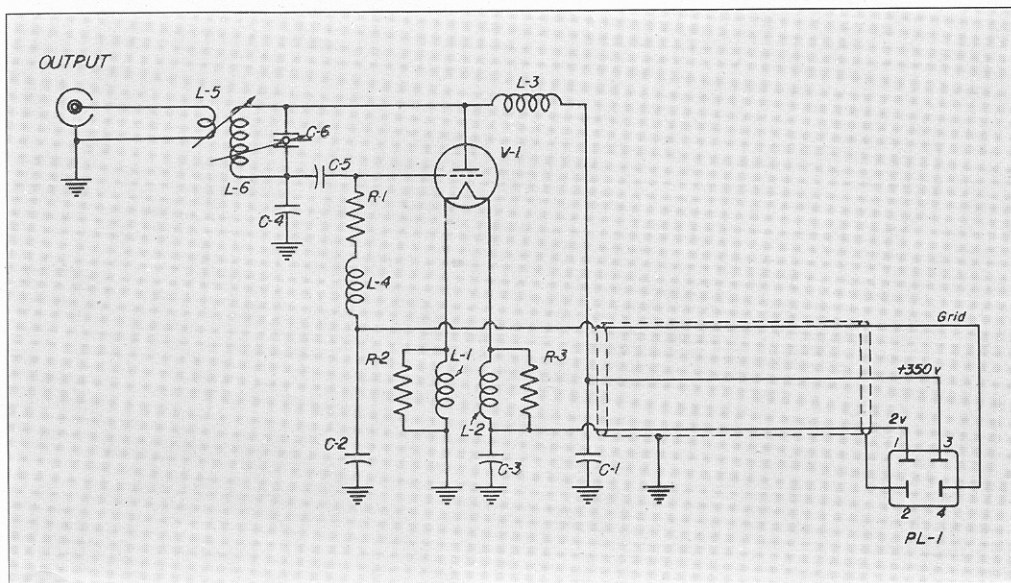
The tuned circuit and its associated output coupling loop are shown in Figure 2.

For a 120° rotation of the rotor the effective capacitance of this circuit varies by a factor of 10.6 from 94 μmf to 8.8 μmf , and the effective inductance by a factor of 2.9 from 30 cm to 10.3 cm, giving a frequency range of 95 Mc to 525 Mc. The capacitance variation of

*Eduard Karplus, "The Butterfly Circuit," General Radio *Experimenter*, XIX, 5, October, 1944.

Copyright, 1944, General Radio Company Cambridge, Mass., U. S. A.

FIGURE 3. Complete circuit diagram of the oscillator unit.





the tuning circuit alone is considerably better than these figures indicate since the grid-to-plate capacitance of the oscillator tube and a trimmer condenser are added to the minimum capacitance.

Grid and plate of the oscillator tube are connected directly across the tuned circuit as shown in Figure 3. The filament of the tube is fed by chokes, allowing it to assume a potential determined by the electrode capacitance of the tube and by the capacitance of the tuned circuit to ground. This latter capacitance is adjusted to optimum value by a small grounding condenser. To compensate for variations in tube-electrode capacitance, a trimmer condenser is provided.

The main dial is calibrated directly in megacycles. The vernier dial carries 100 uniform divisions and covers the tuning range in approximately 10 revolutions. An auxiliary scale on the main dial indicates revolutions of the vernier. A Western Electric TYPE 316-A Vacuum Tube is used in the oscillator. This tube has a thoriated tungsten filament, requiring 2 volts and 3.6 amperes. The maximum allowable plate voltage is 450 volts at 30 watts plate dissipation. Connections to the power supply are made through a

4-conductor shielded cable permanently attached to the oscillator.

The TYPE 857-P1 Power Supply is recommended as a power source. The plate voltage supplied is 350 volts, which is the highest voltage which can be used unless precautions are taken to prevent destruction of the oscillator tube if oscillations are stopped by overcoupling to the load. In addition to filament and plate supply circuits, the power supply contains an electron-ray tube connected to indicate the grid current of the oscillator. The eye of the indicator tube is closed in operation and opens when grid current is interrupted.

The maximum power output obtainable from the TYPE 857-A Oscillator is approximately 1 watt at 500 Mc and is greater at lower frequencies. When the TYPE 857-P1 Power Supply is used, the maximum power output is 0.5 watt.

Coupling to the load is varied by turning the shield of the output jack between two positions marked maximum and minimum, and further variation of the coupling is obtained by moving the output jack in and out through the metal housing. Three positions have been provided.

— EDUARD KARPLUS

TYPE 857-A OSCILLATOR SPECIFICATIONS

Frequency Range: 100 Mc to 500 Mc.

Frequency Calibration: The frequency dial is calibrated directly in megacycles with an accuracy of $\pm 1\%$.

Output Power: The oscillator can deliver a maximum of approximately 1 watt at the highest frequency. At lower frequencies the output power increases. If the oscillator is used with the TYPE 857-P1 Power Supply, the maximum power at the highest frequency is 0.5 watt.

Output Coupling: Output coupling can be varied from maximum to practically zero.

Dial and Drive: The main dial carries the fre-

quency scale and counts in addition the revolutions of the vernier dial. The gear reduction is 15:1 and, with 100 divisions on the vernier, approximately 1000 divisions are provided for the complete scale.

Vacuum Tubes: 1 Western Electric TYPE 316-A.

Accessories Supplied: 1 TYPE 775-M Cable Jack to fit the output connector of the oscillator.

Power Supply: Filament, 2.5 volts at 3.65 amperes. Plate, 450 volts maximum at 30 watts plate dissipation. The TYPE 857-P1 Power Supply has



been designed for this oscillator and its use is recommended. The plate power supplied by this unit is 350 volts at 50 milliamperes. The oscillator carries a permanently attached power cable which plugs into the power supply unit.

Mounting: The oscillator is mounted in a small metal cabinet.

Dimensions: $6\frac{7}{8} \times 7\frac{7}{8} \times 7\frac{1}{4}$ inches, over-all.

Net Weight: $6\frac{1}{4}$ pounds.

TYPE 857-P1 POWER SUPPLY SPECIFICATIONS

Power Input: 115 or 230 volts, 42 to 60 cycles. Total power input, 50 watts with TYPE 857-A Oscillator connected.

Power Output: 2.5 volts AC at 3.65 amperes and 350 volts DC at 50 milliamperes.

Indicator: An electron-ray tube is mounted on the front panel to indicate grid current of the oscillator.

Vacuum Tubes: 1 TYPE 543GT Rectifier and 1 TYPE 6E5 Electron-Ray Tube. Both are supplied.

Accessories Supplied: Spare fuses and 1 line connector cord.

Dimensions: $5\frac{1}{2} \times 6\frac{3}{8} \times 6\frac{7}{8}$ inches, over-all.

Net Weight: $9\frac{1}{2}$ pounds.

Type		Code Word	Price
857-A	U-H-F Oscillator.....	OFTEN	} \$250.00
857-P1	Power Supply.....	OFTENAPACK	
Patent Applied For			

VARIACS FOR 400-CYCLE OPERATION

● **EVER INCREASING DEMANDS** for voltage control devices at alternating-current frequencies of 400 cycles per second and higher have resulted in new additions to the Variac line. TYPES 60-AU (uncased) and 60-AM (cased) Variacs perform, at these higher fre-

quencies, the functions of the familiar TYPE 200-C Variac at 60 cycles. The nominal rating of 5 amperes, 0 to 135 volts when operating from a 115-volt line, may be increased to 7.5 amperes near the zero and line voltage regions.

Important features are:—

● Greatly decreased weight (60-AU, 3 lbs., 2 ozs. versus 200-CU, 8 lbs., 4 ozs.), accomplished by the use of a cast aluminum base, a hollow shaft, and a much smaller core made possible by the higher frequency and by the use of a "scroll" construction.

● General Radio's new Unit Brush, requiring no tools for brush change, and, in cooperation with a rigid radiator structure, preventing contact between brush holder and winding, a frequent cause of burnout in the past.

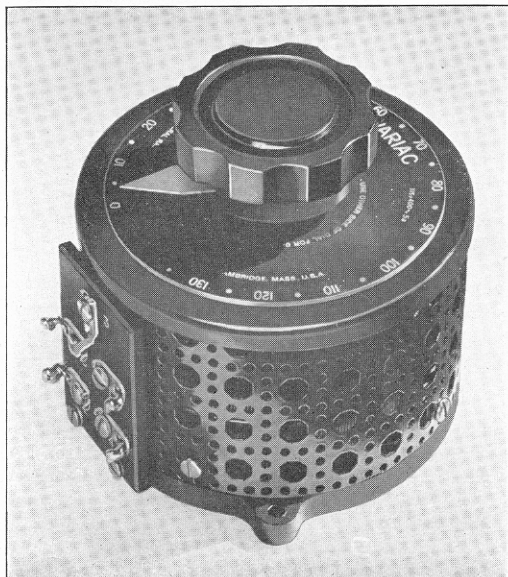


FIGURE 1. View of the TYPE 60-AM Variac showing terminal plate and dial.



- Combination solder and screw terminals for greater convenience in wiring.
- Elimination of the brush "pigtail."
- A single screw adjustment for shaft reversal when changing from panel to table mounting or vice versa.

● Fully insulated hollow steel shaft with no set screws bearing on bakelite.

● Bearing adequate for motor-driven applications. Like other Variacs, TYPES 60-AU and 60-AM are applicable to either adjustable load voltages or to line voltage regulation.

These new Variacs are not in quantity production. All of our Variac manufacturing facilities are fully occupied and scheduled far in advance on the production of the regular 60-cycle units. For the time being, therefore, the new 400-cycle Variacs can be manufactured in relatively very small quantities, which will be available as samples to those interested in determining their suitability for particular applications. The 60-cycle Variac has taken its place as a vital component in wartime equipment as it did

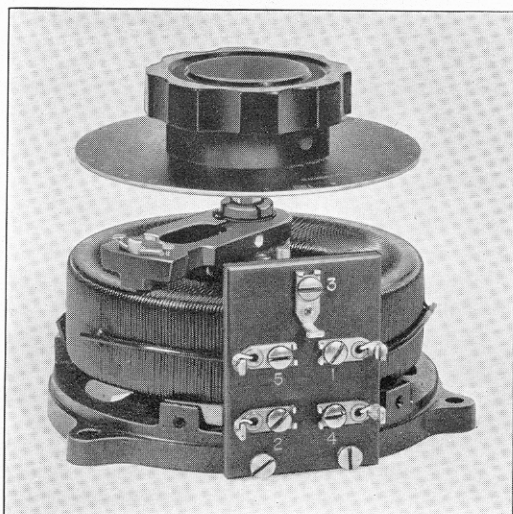
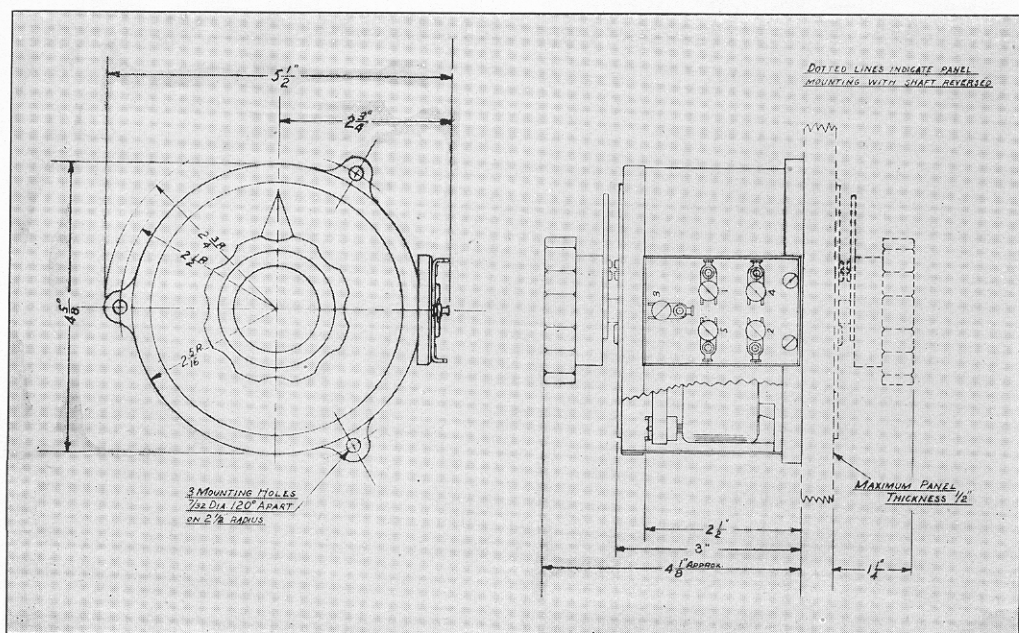


FIGURE 2. View of the uncased model, TYPE 60-AU, showing the new Unit Brush. The brush element is easily removed and replaced if necessary without the use of tools.

before the war in hundreds of civilian products. These top priority war applications are now, and have been since 1941, occupying all of our production facilities. When war requirements per-

FIGURE 3. Dimension sketch of the cased model with cutaway showing dimensions of uncased model.



mit, some machines can be converted for producing the new 400-cycle Variacs. In the meantime, the small preproduc-

tion run will be available soon after the first of the year.

— GILBERT SMILEY

SPECIFICATIONS

Load Rating: 860 va.

Input Voltage: 115 volts.

Frequency: 400 to 2600 cycles per second.

Output Voltage: 0 to 135 volts, or 0 to 115 volts.

Load Current: 5 amperes, rated; 7.5 amperes near zero and line voltages.

No-Load Loss: 7 watts at 400 cycles.

Dimensions: Over-all height for table mounting, $4\frac{1}{8}$ inches; depth behind panel for panel mounting, 3 inches; over-all diameter of base, $5\frac{1}{2}$ inches.

Net Weight: TYPE 60-AU, 3 pounds, 2 ounces. TYPE 60-AM, $3\frac{1}{2}$ pounds.

Type		Code Word	Price
60-AU	Variac (uncased model)	BEGOT	\$13.00
60-AM	Variac (cased model)	BEFOG	15.00

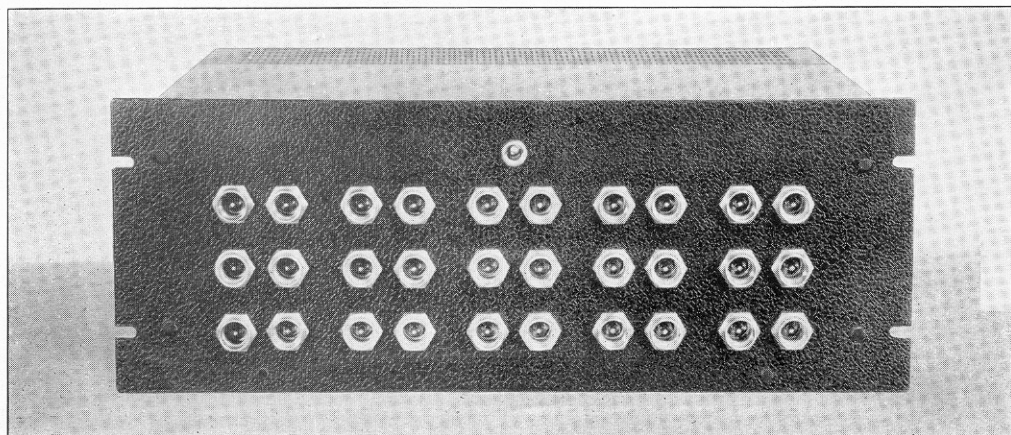
Variacs are manufactured and sold under U. S. Patent No. 2,009,013

DISTRIBUTION OF STANDARD FREQUENCIES

● **DEVELOPMENT** and manufacturing of electronic equipment require the availability of standard frequency facilities for testing and calibrating. If a large quantity of identical equipments is to be calibrated, a permanent arrangement of standard frequency facilities can be arranged for the purpose. However, if large quantities of equipments of a diversified nature are to be calibrated, a considerable degree of flexibility in the arrangement of standard frequency facilities is necessary.

The distribution arrangement described here is one which has served diversified requirements fairly well at the General Radio laboratories. At the central frequency standard a distribution amplifier is provided for supplying sine-wave outputs to the Engineering Department. Each of the output frequencies of 0.1, 1.0, 10.0, and 50.0 kc is filtered and connected to the inputs of four amplifiers in parallel. The outputs, at approximately 500 ohms impedance level, are individually available at jacks

FIGURE 1. Panel view of the distribution amplifier.





on the panel. Lines from the individual engineering laboratories terminate on jacks on a panel mounted below the amplifiers. By use of concentric patch cords, any line can be connected to any amplifier output.

The filtered output at each frequency is available at a pair of jacks on the distribution amplifier, where two "roving" amplifier inputs can be connected. Using these two amplifiers, a total of six channels can be made available on any one standard frequency, or five channels on any two.

Provision is also made, by two jacks in the upper right corner of the amplifier panel, for plugging in either, or both, "roving" amplifiers on the output of the TYPE 614 Selective Amplifier mounted in the standard frequency assembly. This selective amplifier provides output frequencies of any multiple of 1.0 kc from 1 to 10.

At the central frequency standard provision is made for 48 lines running to the individual engineering laboratories and to the manufacturing department.

Since the demand in the manufacturing department for standard frequencies includes both sine-wave and harmonic outputs, and since the number of operating positions is far larger than could be handled conveniently from the central standard, use is made of banks of multivibrators set up at various points in the department. These multivibrators are controlled by 50 kc from the central standard, so the output frequencies are identical with those available at the central location. At each multivibrator bank a distribution amplifier and line panel is provided for distributing the standard frequencies to workers near by.

In the crystal grinding department three similar banks of multivibrators are set up, with various frequency measuring

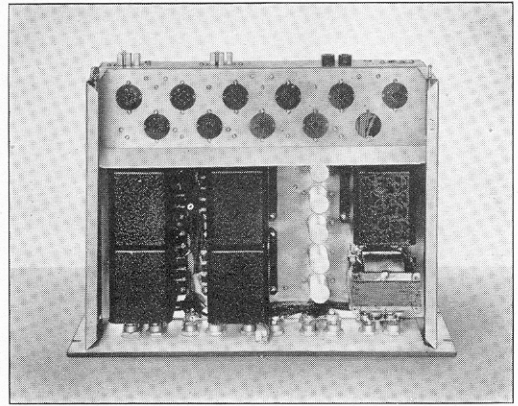


FIGURE 2. Top view of amplifier, with dust cover removed.

assemblies adjacent to them. The choice of multivibrators and units of the measuring equipment is made to meet the particular requirements of operations carried out at each location.

— J. K. CLAPP

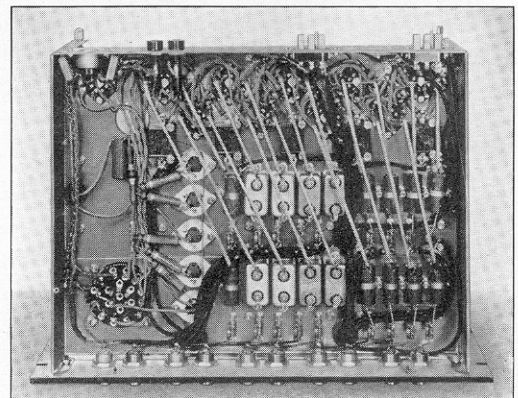
PARTS LIST FOR THE DISTRIBUTION AMPLIFIER

(See circuit diagram on page 8)

RESISTORS

R-1 = 10 k Ω	R-12 = 10 k Ω	R-24 = 1.0 M Ω
R-2 = 10 k Ω	R-13 = 10 k Ω	R-25 = 20 k Ω
R-3 = 10 k Ω	R-14 = 10 k Ω	R-26 = 0.1 k Ω
R-4 = 10 k Ω	R-15 = 10 k Ω	R-27 = 5 k Ω
R-5 = 10 k Ω	R-16 = 10 k Ω	R-44 = 1 k Ω
R-6 = 10 k Ω	R-17 = 10 k Ω	R-45 = 1 k Ω
R-7 = 10 k Ω	R-18 = 10 k Ω	R-46 = 1 k Ω
R-8 = 10 k Ω	R-19 = 1.0 M Ω	R-47 = 1 k Ω
R-9 = 10 k Ω	R-20 = 1.0 M Ω	R-48 = 1 k Ω
R-10 = 10 k Ω	R-21 = 1.0 M Ω	R-49 = 1 k Ω
R-11 = 10 k Ω	R-22 = 1.0 M Ω	
	R-23 = 1.0 M Ω	

FIGURE 3. View from below with dust cover removed, showing arrangement of resistors and capacitors.



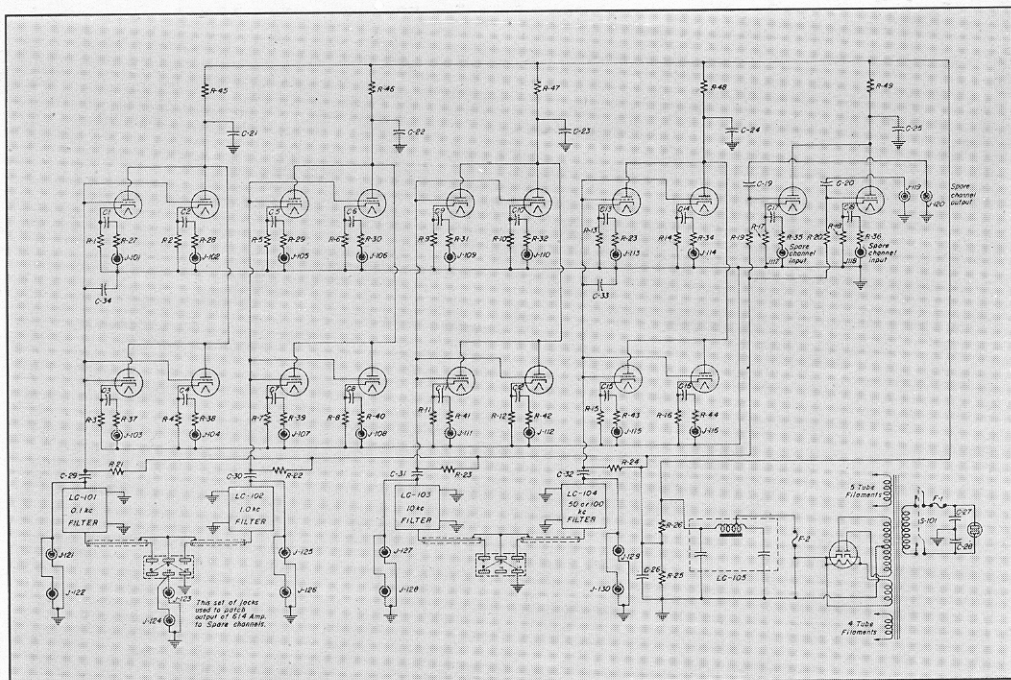


FIGURE 4. Complete circuit diagram of the distribution amplifier.

CONDENSERS

C-1 = 16 μ f
C-2 = 16 μ f
C-3 = 16 μ f
C-4 = 16 μ f
C-5 = 16 μ f
C-6 = 16 μ f
C-7 = 16 μ f
C-8 = 16 μ f
C-9 = 0.5 μ f

C-10 = 0.5 μ f
C-11 = 0.5 μ f
C-12 = 0.5 μ f
C-13 = 0.1 μ f
C-14 = 0.1 μ f
C-15 = 0.1 μ f
C-16 = 0.1 μ f
C-17 = 16 μ f

C-18 = 16 μ f
C-19 = 0.03 μ f
C-20 = 0.03 μ f
C-21 = 20 μ f
C-22 = 20 μ f
C-23 = 20 μ f
C-24 = 20 μ f
C-25 = 20 μ f

C-26 = 16 μ f
C-27 = 0.01 μ f
C-28 = 0.01 μ f
C-29 = 0.03 μ f
C-30 = 0.03 μ f
C-31 = 0.03 μ f
C-32 = 0.03 μ f
C-33 = 0.0001 μ f
C-34 = 0.0001 μ f

J-101 } = G.R. TYPE 774-G
to
J-130 }

LC-105 Rectifier Filter
LC-101 to LC-104 Low-Pass Filters
T-101 Power Transformer
F-1 = 2.0 amp.
F-2 = 0.2 amp.

Note: This amplifier was built for use in the General Radio laboratories and is not for sale as a standard catalog instrument. Owing to the pressure of essential war production the General Radio Company cannot furnish either complete amplifiers of this type or parts from which they can be constructed. This description is published merely as a guide to those who may wish to build similar equipment using such parts as are generally available. — Editor.

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VOLUME XIX No. 7

DECEMBER, 1944

ELECTRICAL MEASUREMENTS AND THEIR INDUSTRIAL APPLICATIONS

A D-C VACUUM-TUBE VOLTMETER

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test and hence that the indicated voltage is the actual circuit voltage. The TYPE 728-A D-C Vacuum-Tube Voltmeter shown in Figure 1 maintains this advantage over a voltage range of approximately 60,000:1. The minimum voltage that can be measured with this meter is 0.05 volt, and the maximum is 3000 volts. For voltages up to 30 volts the input resistance is over 5000 megohms; for voltages above 30, it is 1000 megohms. Operated from internal batteries, it is a companion instrument to the TYPE 727-A Vacuum-

●CIRCUIT DESIGN AND TESTING in radio and electronic equipment involves the measurement of a number of d-c voltages, ranging in magnitude from a fraction of a volt to several thousand volts, associated with impedance from a few ohms to many megohms. The advantage of using a vacuum-tube voltmeter for these measurements is the assurance that no power is drawn from the circuit under

FIGURE 1. View of the TYPE 728-A Vacuum-Tube Voltmeter with cover open.



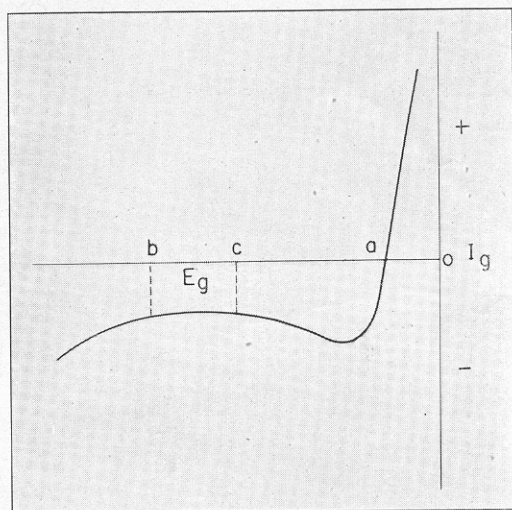


FIGURE 2. Typical grid current characteristic for the amplifier tube.

Tube Voltmeter,¹ for a-c voltage measurements up to 100 megacycles and the TYPE 729-A Megohmmeter² (2000 ohms to 50,000 megohms). These three meters comprise a group of small, portable, accurate instruments for use in the laboratory or in the field where power lines are not accessible.

An analysis of the input circuit of a d-c vacuum-tube voltmeter discloses that grid current³ in the amplifier tube must be kept at a minimum to obtain high input resistance, but of equal importance is the leakage path between input terminals due to wiring, anchor terminals, switches, etc. Here, a choice of materials that maintain high resistance in the face of adverse humidity⁴ conditions is imperative. The grid current in the amplifier tube is a function of tube type, operating conditions, and circuit arrangement. If the grid current of an individual tube is plotted against grid bias, a curve of the general shape shown in Figure 2 will be obtained. It will be noted that at one point (*a*) there is no

grid current, and the input resistance, in so far as grid current is concerned, is infinite. But, to operate the tube over a range of input voltage, the grid bias must change. It is therefore advisable to select a suitable range over which the bias change occurs and, as is done in the circuit of the TYPE 728-A D-C Vacuum-Tube Voltmeter, to introduce a large amount of degeneration in order to limit the required bias increment (e.g., *b* to *c* in the figure) and so to prevent operation over a region of high grid current.

Over the three lower voltage ranges of the TYPE 728-A D-C Vacuum-Tube Voltmeter (0-3, 0-10, and 0-30 volts), the input terminals are connected directly to the open grid of the tube. While manufacturing controls assure an input resistance in excess of 5000 megohms, the usual input resistance is about 20,000 megohms. This high resistance has an interesting effect which may at first seem strange: if a potential of, say, ten volts is applied to the input terminals and then removed, the meter will continue to indicate the ten volts for perhaps a minute or longer. This occurs because the capacitance from grid to ground becomes charged and, in view of the very large leakage resistance, it simply takes time for the capacitor to discharge ($RC = 10,000 \times 10^6 \times .01 \times 10^{-6} = 100$ seconds). Of course, as soon as the input terminals are shorted, or shunted with the external circuit under test, the meter reading drops to zero or to the value of the potential being measured.

To extend the voltage range upward, a voltage divider is switched between high-voltage terminals and the vacuum-tube voltmeter circuit. The input resistance of the instrument for the 0-100, 0-300, 0-1000, and 0-3000 volt ranges is 1000 megohms. The choice, treatment,



adjustment, and mounting of the resistors used for such a high-voltage divider are of paramount importance. There are considerations of voltage and temperature coefficient, of aging and moisture proofing. The resistors in the TYPE 728-A Voltmeter were carefully selected with all of these considerations in mind. They are spirally grooved and, before final selection, are subjected to artificial aging cycles and to impregnation.

In addition to the major design considerations, various features have been incorporated to increase the usefulness of the instrument. An R-C filter permits the measurement of d-c voltages in spite of superimposed a-c potentials as high as 200 volts. A polarity switch obviates the necessity of transposing meter connections and, when desirable, permits grounding either pole to the meter panel. This is extremely useful when checking an electronic circuit where various potentials of different polarity to ground

must be determined. All batteries are contained in the small cabinet, and a simple switch manipulation determines the battery condition and warns the operator when the batteries must be replaced, whether because of low voltage or excessive battery resistance. To safeguard the batteries, the power switch is automatically turned off when the instrument cover is closed.

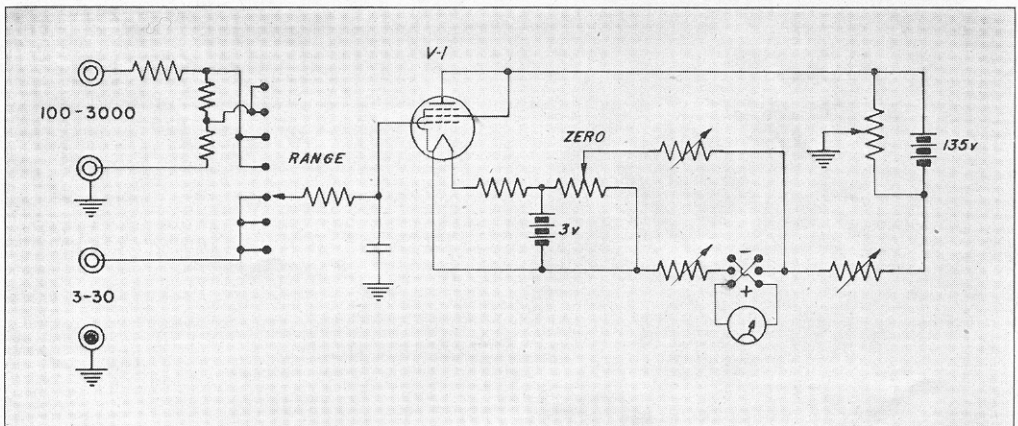
The TYPE 728-A D-C Vacuum-Tube Voltmeter fills the need for a voltmeter of very high input resistance that can be used for checking small tube bias potentials or high cathode-ray accelerating voltages, automatic volume control, or automatic frequency control circuits, f-m discriminator circuits or, in general, the d-c voltages that appear in any setup in the laboratory or in the field. Furthermore, these voltages can be measured with the assurance that the input resistance of the voltmeter will cause no error.

— A. G. BOUSQUET

REFERENCES

1. W. N. Tuttle, "The Type 727-A Vacuum-Tube Voltmeter," *General Radio Experimenter*, Vol. XVI, No. 12, May, 1942.
2. W. N. Tuttle, "A Portable Megohmmeter," *General Radio Experimenter*, Vol. XV, No. 2, July, 1940.
3. W. B. Nottingham, "Measurement of Small D-C Potentials and Currents in High Resistance Circuits by Using Vacuum Tubes," *Journal of the Franklin Institute*, Vol. 209, No. 3, March, 1930, pp. 287-348.
4. R. F. Field, "The Effect of Humidity on Electrical Measurements," *General Radio Experimenter*, Vol. XVIII, No. 3, August, 1943.

FIGURE 3. Schematic circuit diagram of the TYPE 728-A Vacuum-Tube Voltmeter.



SPECIFICATIONS

Range: 0.05 to 3000 volts in seven ranges (3, 10, 30, 100, 300, 1000, 3000 volts, full scale).

Accuracy: Within $\pm 3\%$ of full scale for the lower voltage ranges (3, 10, 30 volts, full scale). Within $\pm 5\%$ of full scale for the higher voltage ranges (100, 300, 1000, 3000 volts, full scale). Battery aging may cause an additional error of 2% of full scale on the 3-volt range.

Input Resistance: 1000 megohms on the higher voltage ranges (100, 300, 1000, 3000 volts, full scale). Greater than 5000 megohms on the low voltage ranges.

Terminals: Two sets of input terminals are provided on the panel. One set is used for measurements at the low voltage end of the range (0 to 30 volts) and the other set is used for the higher voltage measurements (30 to 3000 volts).

Polarity: A reversing switch on the panel permits measurements with either the positive or

the negative terminal of the source grounded to the panel of the instrument.

Effect of A-C: A superimposed a-c voltage of as high as 200 volts has a negligible effect on the meter indication.

Tube: The tube, a type 1E5-GP, is supplied.

Batteries: The batteries required are three Burgess W30BP or equivalent and one Burgess F2BP or equivalent. A compartment is provided in the case of the instrument for holding all batteries. A set of batteries is supplied with the instrument.

Mounting: The instrument is supplied in a walnut case with cover and is mounted on an engraved black crackle-finish aluminum panel.

Dimensions: With cover closed, (length) 11 x (width) $6\frac{5}{8}$ x (height) $5\frac{7}{8}$ inches, over-all.

Net Weight: $9\frac{3}{4}$ pounds, including batteries.

Type		Code Word	Price
728-A	D-C Vacuum-Tube Voltmeter. .	PILOT	\$110.00

MUTUAL INDUCTANCE MEASUREMENT WITH THE TYPE 650-A IMPEDANCE BRIDGE

● IN BOTH THE LABORATORY and the factory, the TYPE 650-A Im-

pedance Bridge provides a rapid means of checking inductors, capacitors, and resistors with an accuracy sufficient for nearly all commercial requirements. In addition, this bridge can be adapted for many specialized measurements beyond the scope of its original design. A good example is furnished by a problem recently encountered in the measurement of the mutual inductance of automotive ignition coils and magnetos.

A TYPE 650-A Impedance Bridge was available for this measurement, and a consideration of the circuit showed that the instrument could easily be converted

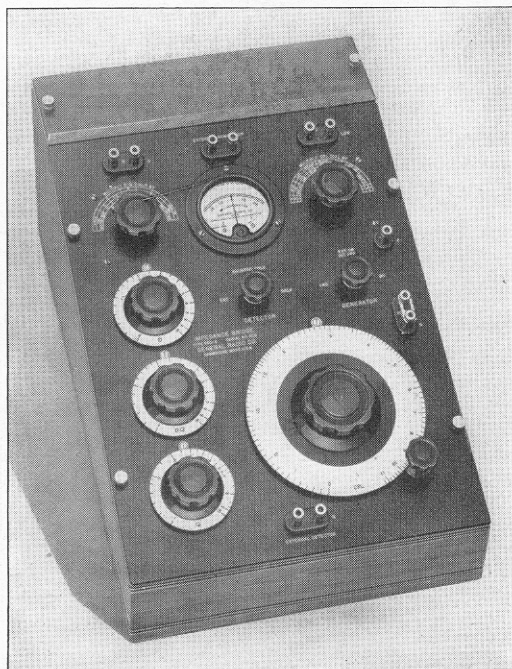


FIGURE 1. View of the TYPE 650-A Impedance Bridge.

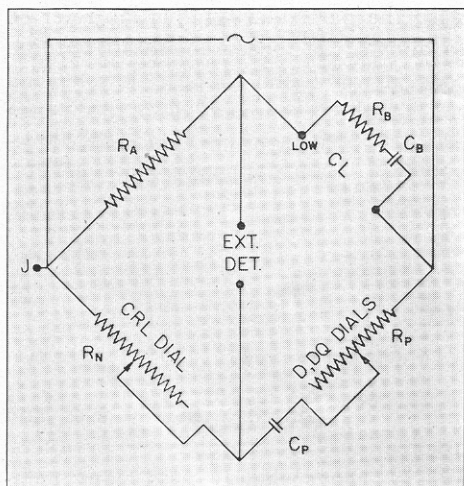


FIGURE 2. Schematic circuit diagram for the TYPE 650-A Impedance Bridge as used for capacitance measurements.

to a Carey-Foster type of mutual inductance bridge.*

The schematic bridge circuit for *capacitance* measurements is shown in Figure 2. To convert to a Carey-Foster bridge, the A-arm is shorted, and the unknown inductor and an external resistor are connected as shown in Figure 3. Shorting of the A-arm is accomplished by connecting the junction *J* to the high EXT DET terminal or to the low CL terminal. One winding of the ignition coil is connected to the CL terminals in series with the external resistor. The other winding is placed in series with the head telephones. The external resistor may, for convenience, be a decade resistance box of 100,000 ohms maximum setting. A known fixed resistor of suitable magnitude is equally satisfactory.

Balance is obtained by varying the CRL dial and either the D or the DQ dial. The mutual inductance, *M*, is then given by

$$M = R_N(R_B + R_L)C_P$$

*Other methods of measuring mutual inductance were discussed in an article by R. F. Field, "The Measurement of Mutual Inductance," *General Radio Experimenter*, January, 1937.

where R_N is the reading of the CRL dial multiplied by 1000, R_B is the external resistor in the B-arm, R_L is the effective resistance of the inductor, and C_P is the capacitance of the standard mica condenser in the P-arm, 10^{-8} farads.

The inductance of the winding connected in the P-arm can be determined from the expression

$$L = (R_N + R_P)(R_B + R_L)C_P.$$

The resistance R_P depends upon the setting of the D or DQ dial at balance. The choice of the dial to be used is determined by the resistance needed for balance. The maximum resistance of the D dial is 1650 ohms, while that of the DQ dial is 16,500 ohms. The magnitude of R_P is calculated from the expression

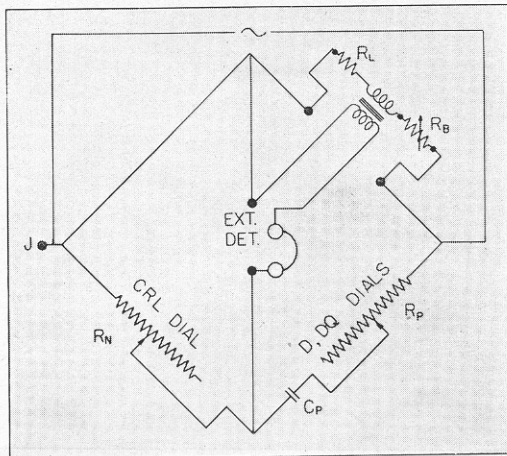
$$R_P = \frac{D}{6.28 \times 10^{-5}}$$

where *D* is the dial reading multiplied by the factor indicated by the multiplier switch setting.

The resistance, R_B , of the decade box should be large compared to R_L , the effective resistance of the coil under measurement, so that R_L can be neglected. For this condition,

$$M = R_N R_B C_P$$

FIGURE 3. Modification of the circuit of Figure 2 for measurements of mutual inductance.



and

$$L = (R_N + R_P)(R_B C_P).$$

When R_L is not negligible compared to R_B , the inductance and storage factor Q of the coil can be first measured with the bridge connected for inductance measurement, and the value R_L then calculated.

In the particular application referred to above, the measurement of ignition coils, it was found that the low tension coil should be connected to the P-arm of the bridge in order that balance be within range of the bridge controls.

— L. E. PACKARD

CHICAGO DISTRICT OFFICE

920 South Michigan Avenue

● AS MANY OF OUR READERS in the Chicago area know, the General Radio Company has been operating an Engineering and Sales Office since December 1, 1943, on South Michigan Avenue. Mr. L. E. Packard from our Cambridge engineering staff is in charge, and is already well known to many of our Chicago friends. Mr. Packard is thor-

oughly acquainted with the uses and applications of General Radio products, and is always glad to pass along technical information gained through his experience in the measurement field. The facilities of the office are available not only to our customers in greater Chicago, but to all in the Mid-West area who find it more convenient to communicate with

View of the interior of the General Radio Company's Chicago Office.





our Chicago office than with the main office at Cambridge.

A demonstration laboratory is maintained in which many of our latest instruments may be inspected and tested under operating conditions. New equipment will be on display as it becomes available. Arrangements can also be made to have some of the smaller instruments demonstrated in customers' plants.

Our office in Chicago is in a position to give valuable and prompt assistance regarding price and delivery, as well as to advise regarding Government restrictions on the purchase of equipment.

Occasionally an engineering conference meeting at the office on a subject of interest to a particular group of engi-

neers will be arranged. Space permits a maximum of about fifty people to be accommodated. Meetings on the subject of sound and vibration, as well as on new broadcast station monitoring equipment, were held early in October of this year.

Mr. Packard is often able to offer suggestions that are useful in keeping General Radio equipment in satisfactory operating condition, or, when repairs are necessary, to arrange for the work to be done at our factory. A small shop for minor emergency repair work is maintained at Chicago.

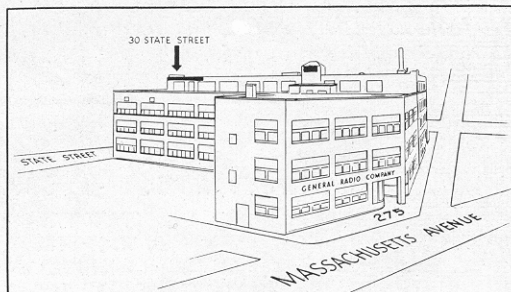
Parking space, an important consideration in congested cities, is available at the rear of the office building, and may be used by visitors who wish to drive to the Chicago office.

MISCELLANY

● **TO THOSE WHO CALL** on us at our new address the accompanying sketch may be helpful. It shows our new front door in relation to our old one.

Congratulations to W. L. Everitt on his election to the presidency of the Institute of Radio Engineers. Dr. Everitt's recent article in the *Proceedings of the I. R. E.* on engineering education has proved a timely one, and has stimulated a good deal of discussion. At a recent meeting of the Boston Section, representatives of both education and industry gave their views on the subject for over two hours, with plenty of people asking to be heard when the meeting adjourned.

is a subject of considerable importance to the radio engineer, and research on dielectrics has naturally been stimulated by the war. One of the leaders in this field is General Radio's Robert F. Field, who has delivered several papers on the subject before technical societies. His latest: "The Behavior of Dielectrics Over Wide Ranges of Frequency and Temperature," at a meeting of the Basic Science Group of the A.I.E.E. at New York on November 9, 1944. We hope to



The nature and behavior of dielectrics



publish some of this material in the *Experimenter* soon.

At the National Electronics Conference held recently in Chicago, papers were presented by two General Radio engineers. H. H. Scott's paper, entitled "Audible Audio Distortion," was devoted to distortion caused by intermodulation and to methods by which it can be measured. D. B. Sinclair discussed "Wide Range Tuned Circuits for High Frequencies," covering not only the butterfly circuits recently described by Eduard Karplus in the *Experimenter*, but also some later cylindrical circuits that we hope will soon appear in the *Experimenter*. Mr. Scott also held two conferences at the General Radio Chicago Office, one on vibration measurements and the other on broadcast station monitoring equipment. At a meeting of the Chicago Radio Engineers Club, Mr. Scott spoke on "The Pulsing of Standard-Signal Generators."

Representatives from our district engineering offices, F. Ireland from Los Angeles, L. E. Packard from Chicago, and M. A. Gilman from New York, have recently completed a visit of several

days to the main office in Cambridge, getting posted on new developments in engineering and manufacturing. Up-to-date information on instrument performance and delivery dates is available at each of our branch offices. In addition, the engineer in charge has a considerable background of experience in the application of GR instruments to customers' problems. If you are located in or near New York, Chicago, or Los Angeles, we urge you to make use of our district office facilities.

You may not have heard of our "reservation-order" system, which has been in effect for some time on broadcast equipment and which has now been extended to cover all of our products. We are glad to receive your non-priority order now for shipment later on when war conditions permit. We are receiving a substantial number of these reservation-orders which we file in chronological sequence and which we will fill before any new orders for similar material. If you are anticipating your postwar requirements, send us your reservation-order now. It is understood, of course, that these orders may be canceled up to sixty days before shipment is to be made.

THE General Radio EXPERIMENTER is mailed without charge each month to engineers, scientists, technicians, and others interested in communication-frequency measurement and control problems. When sending requests for subscriptions and address-change notices, please supply the following information: name, company name, company address, type of business company is engaged in, and title of position of individual.

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